



NAVAL POSTGRADUATE SCHOOL

MONTEREY, CALIFORNIA

THESIS

**MODELING AND CONTROL OF POWER
CONVERTERS FOR ENERGY STORAGE INTERFACE
ON A MICROGRID**

by

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June 2019

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REPORT DOCUMENTATION PAGE			<i>Form Approved OMB No. 0704-0188</i>	
Public reporting burden for this collection of information is estimated to average 1 hour per response, including the time for reviewing instruction, searching existing data sources, gathering and maintaining the data needed, and completing and reviewing the collection of information. Send comments regarding this burden estimate or any other aspect of this collection of information, including suggestions for reducing this burden, to Washington headquarters Services, Directorate for Information Operations and Reports, 1215 Jefferson Davis Highway, Suite 1204, Arlington, VA 22202-4302, and to the Office of Management and Budget, Paperwork Reduction Project (0704-0188) Washington, DC 20503.				
1. AGENCY USE ONLY (Leave blank)		2. REPORT DATE June 2019	3. REPORT TYPE AND DATES COVERED Master's thesis	
4. TITLE AND SUBTITLE MODELING AND CONTROL OF POWER CONVERTERS FOR ENERGY STORAGE INTERFACE ON A MICROGRID			5. FUNDING NUMBERS	
6. AUTHOR(S) Theodoros Karapas				
7. PERFORMING ORGANIZATION NAME(S) AND ADDRESS(ES) Naval Postgraduate School Monterey, CA 93943-5000			8. PERFORMING ORGANIZATION REPORT NUMBER	
9. SPONSORING / MONITORING AGENCY NAME(S) AND ADDRESS(ES) N/A			10. SPONSORING / MONITORING AGENCY REPORT NUMBER	
11. SUPPLEMENTARY NOTES The views expressed in this thesis are those of the author and do not reflect the official policy or position of the Department of Defense or the U.S. Government.				
12a. DISTRIBUTION / AVAILABILITY STATEMENT Approved for public release. Distribution is unlimited.			12b. DISTRIBUTION CODE A	
13. ABSTRACT (maximum 200 words) Power DC-to-DC converters are critical to AC and DC microgrids. Among the DC-to-DC power converters, the boost converter has a right half plane zero, which makes it difficult to control. The load voltage must be kept constant, no matter the variations in the input voltage or the change of the load; to achieve this, a negative feedback loop and a control strategy are necessary. This research presents five different digital control strategies to regulate the output of a non-ideal DC-DC boost converter, operating either in continuous conduction mode (CCM) or discontinuous conduction mode (DCM). Four of them are innovative control strategies, implemented by modifying regular K-factor and regular voltage mode control. The fifth is current mode, which is widely used as the industrial standard. A simple algorithm to set up the parameters for each compensator was introduced. In an ideal system, these control strategies performed well, but when time delays were added the performance of the controllers changed significantly, and the stability of the system could be affected. The closed-loop performance of all controllers was evaluated and compared. Simulation results are provided to show the effectiveness of each controller and compare their performance. A laboratory experiment and parametric sweep measurements were made to validate the boost converter small-signal model.				
14. SUBJECT TERMS energy storage, boost converter, microgrid, controller, control strategies, current control, K-factor, PI compensator, field programmable gate arrays (FPGAs), power electronics			15. NUMBER OF PAGES 125	
			16. PRICE CODE	
17. SECURITY CLASSIFICATION OF REPORT Unclassified	18. SECURITY CLASSIFICATION OF THIS PAGE Unclassified	19. SECURITY CLASSIFICATION OF ABSTRACT Unclassified	20. LIMITATION OF ABSTRACT UU	

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**MODELING AND CONTROL OF POWER CONVERTERS FOR ENERGY
STORAGE INTERFACE ON A MICROGRID**

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Submitted in partial fulfillment of the
requirements for the degrees of

ELECTRICAL ENGINEER

and

MASTER OF SCIENCE IN ELECTRICAL ENGINEERING

from the

**NAVAL POSTGRADUATE SCHOOL
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ABSTRACT

Power DC-to-DC converters are critical to AC and DC microgrids. Among the DC-to-DC power converters, the boost converter has a right half plane zero, which makes it difficult to control. The load voltage must be kept constant, no matter the variations in the input voltage or the change of the load; to achieve this, a negative feedback loop and a control strategy are necessary. This research presents five different digital control strategies to regulate the output of a non-ideal DC-DC boost converter, operating either in continuous conduction mode (CCM) or discontinuous conduction mode (DCM). Four of them are innovative control strategies, implemented by modifying regular K-factor and regular voltage mode control. The fifth is current mode, which is widely used as the industrial standard. A simple algorithm to set up the parameters for each compensator was introduced. In an ideal system, these control strategies performed well, but when time delays were added the performance of the controllers changed significantly, and the stability of the system could be affected. The closed-loop performance of all controllers was evaluated and compared. Simulation results are provided to show the effectiveness of each controller and compare their performance. A laboratory experiment and parametric sweep measurements were made to validate the boost converter small-signal model.

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LIST OF ACRONYMS AND ABBREVIATIONS

AC	alternating current
DC	direct current
A/D	analog-to-digital
CCM	continuous conduction mode
DCM	discontinuous conduction mode
EKF	extended Kalman filter
FGPA	field-programmable gate array
PI	proportional-integral
PV	photovoltaic
PWM	pulse-width modulation
SA	switch position
SC	supercapacitor
LQR	linear quadratic regulator

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ACKNOWLEDGMENTS

First of all, I need to give thanks to the Lord, who has blessed me with the opportunity to attend the Naval Postgraduate School.

I sincerely wish to express my gratitude to Professor Giovanna Oriti, Professor Roberto Cristi, and Doctor Alexander Julian, for their guidance, assistance, and patience over the last several months.

I also would like to acknowledge all of the professors and teachers by whom I have been taught in the past. They gave me the academic foundation I needed to understand the science to conclude my studies at NPS.

To my wife, Fotini, and my daughter, Eleni, thank you for helping and motivating me during this wonderful journey.

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I. INTRODUCTION

This chapter introduces the needs that motivate this research, the research purpose, and a description of previous work.

A. MOTIVATION

Electrical energy is an important and vital part of modern life. Every modern navy should be interested in distributing effectively the electrical energy of a ship. Each ship is a microgrid, having all necessary equipment such as generators, power lines, inverters, batteries, and converters. A microgrid is a small network of electricity, including a local source of supply, which is able to function independently (islanding mode) but also may be connected to a centralized national grid (grid-connected mode). Distributing the electrical energy and controlling power converters plays an important part in a modern navy, since it increases a ship's autonomy, reduces operation cost, and reduces environmental pollution. One type of power converter is the DC-DC boost converter, which converts the low voltage of a battery to the higher voltage required by the load. This component is becoming more and more important as we see ships with integrated electric propulsion (like the U.S. Navy's Makin Island and the UK Royal Navy's Type 45 destroyer class) and a fully electric ship (U.S. Navy DDG-1000 Zumwalt) [1].

B. RESEARCH OBJECTIVES

A DC-to-DC power converter is a critical building block required in a wide variety of applications such as AC and DC microgrids. A microgrid operating in islanding mode uses batteries in addition to other energy sources to produce the required voltage for the loads. Batteries require DC-to-DC power converters, to regulate the voltage to the level required by the loads. When operating in this "grid-forming" mode, the load voltage must be kept constant, no matter the variations in the input voltage or the change of the load; this is achieved by feedback control designed to compensate for various perturbations.

In both DC and AC application, there is often the need to shift the DC voltage of the energy sources by a proper amount. This is particularly important when the energy

source is connected to a battery, which has specific requirements on the input voltage. This shift in DC voltage is obtained by either a “buck” or a “boost” converter, which provides a lower or higher voltage, respectively [2], [3].

Of particular interest to this research is the boost converter, since its control is particularly challenging, due to a right half plane zero in its small-signal model. This makes it difficult to control due to the inherent time delay associated with this dynamic. Furthermore, in practical applications, digital controllers, such as field programmable gate arrays (FPGAs), are used to control the converter, so analog-to-digital (A/D) converters are necessary, but A/D converters introduce time delays in the control system. When those time delays are combined, the performance of the controller may change significantly, thus affecting the stability of the system.

The goal of this research was to study the modeling and the control of a boost converter for energy storage interface on a microgrid. The first objective was to analyze a non-ideal DC-to-DC boost converter and to derive the linearized small-signal model. The second objective was to implement and compare different control strategies to regulate the non-ideal boost converter output during all the modes of operation. The second objective was conducted in two steps: first, implementing controllers that achieve the desired performance, and second, comparing performance.

Different controller implementations were examined. First, existing control strategies were implemented: regular K-factor control method, regular voltage mode control method, and current mode control method. The K-factor is a technique to design a compensator to provide desired gain and phase margins [4]. For each of those controllers a simple algorithm was introduced to set the required parameters. Next, two innovative modifications for each K-factor and voltage mode control were introduced to improve their performances.

In order to evaluate the performance of each controller (Mod 1 K-factor control, Mod 2 K-factor control, Mod 1 voltage mode control, and Mod 2 voltage mode control), they were compared with the current mode control, which is widely used as the industry standard. To evaluate the effectiveness of each controller, three performance indexes were

tested: the command voltage tracking capability, the robustness to the disturbances, and the time to reach the steady state.

This research included a computer simulation (using MATLAB and Simulink software) as well as a laboratory experiment and Simulink experiment to verify the boost converter small-signal model.

C. COMPARISON TO RELATED WORK

The study of power converters is a well-documented topic [2], [3]. In particular, the boost converter, designed to increase the voltage of a source, presents a challenge in its control due to the “non-minimum phase” frequency response. This is due to a zero in the right half plane of the transfer function of its small-signal model, which cannot be overcome by feedback. Several studies involving boost converters have addressed different control methodologies, the effect of time delays, and open-loop frequency response measurement. Some of these works are highlighted below.

Several different control methodologies have been examined in previous works [2]–[15], such as the regular K-factor, the regular voltage mode control, and the current mode control. There are three types of regular K-factor compensator that have been introduced in the past [4]. The regular K-factor controller has already been applied in a boost converter operating in continuous conduction mode (CCM) [5], [6], [7] to regulate the output voltage of sources such as solar photovoltaic (PV) arrays [5], or battery voltage. The need to improve the performance of the regular K-factor has already been realized, and different algorithms to optimize the regular K-factor performance have been introduced [6], [7]. Therefore, in this thesis two innovating modifications of the regular K-factor are introduced, giving the ability to the boost converter to operate in both CCM and discontinuous conduction mode (DCM).

Compensators using the regular voltage mode control have been used to regulate the boost converter output, operating in CCM. Similar with the regular K-factor, regular voltage mode has been applied so that the input voltage is coming from either a stable source such as a battery [8] or a source with big variations like a solar PV array [9]. Different methods have been introduced to tune the proportional-integral (PI) controllers,

such as trial and error, the Ziegler-Nichols method, or introducing an arithmetic method like Bode's integrals [10]. In this thesis, an arithmetic method is used to effectively set up the PI compensators. Then, due to load limitations of the regular voltage mode control compensator, two innovative modifications are introduced. That way, the boost converter operates in the wide range of loads, in both CCM and DCM modes of operation.

A compensator using current mode control is a popular and effective method to regulate the converter output in the CCM. Several studies present the implementation of this control method [8], [11]–[15]. In this thesis, an arithmetic method is used to set up the necessary parameters, so that the boost converter operates in both CCM and DCM.

The effect of the time delays in a second-order closed loop has already been analyzed in the past [16]. A DC-DC boost converter is a second-order system, so for a digital implementation of a controller, time delays should be a consideration. The robustness of a boost converter with a time delay control has also been examined in the past [17]. In this thesis, time delays have been introduced in the system, each controller is set up to regulate the boost converter output, regardless the time delays, and each control methodology performance is evaluated.

Several studies present the experimental measurement of the frequency response of a boost converter [18], [19], [20]. The sweep analysis can be implemented either to verify the boost converter model transfer function [18] or as a method to measure the parasitic elements of a converter [19]. In this thesis, a hardware experiment in the laboratory and a simulation experiment are conducted to verify the mathematical model of the open-loop frequency response of a boost converter.

D. ORGANIZATION

A detailed analysis of the boost converter architecture and state-space model is introduced in Chapter II. A typical boost converter architecture is introduced, and its components and the possible modes of operation are explained. A circuit analysis is made to derive the boost state-space model. Then, this state-space model is used to derive the small-signal state-space model, and eventually the small-signal transfer function.

Chapter III provides an analysis on different control strategies. First, the regular K-factor compensator is analyzed, providing a simple algorithm to tune the controller parameters. Then, two different modification of the regular K-factor compensator are introduced, to improve its performance during the change of the mode of operation of the boost converter. Next, the regular voltage mode control methodology is analyzed, providing a simple algorithm to tune the PI controller parameters. Due to a limitation in changing the mode of operation of the regular voltage mode control, two modifications are introduced to overcome this limitation. Finally, the current mode control methodology is introduced, providing a simple algorithm to tune the compensator parameters.

Chapter IV provides a laboratory implementation and experimental measurements of the boost converter open-loop transfer function. First, the hardware experiment is explained, with the hardware setup, the equipment used, the procedure used, and the test results. Then, the simulation experiment is explained, with the procedure and the test results. Finally, a comparison is made between the theoretical open-loop transfer function from Chapter II, the measured open-loop transfer function from the hardware experiment, and the measured open-loop transfer function from the simulation experiment.

Chapter V provides a comparison between different control strategies: the two modifications of the regular K-factor compensator, the two modifications of the regular voltage mode compensator, and the current mode compensator. A scenario that includes various step load changes is simulated for every control methodology, and the performance of the method is evaluated. Finally, Chapter VI provides the conclusions of this research and recommendations for future work.

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II. BOOST CONVERTER ARCHITECTURE AND STATE-SPACE MODEL

Before analyzing the control methodologies of a boost converter, it is essential to describe a non-ideal boost converter and derive its small-signal model. The necessary background information is provided in this chapter.

A. BOOST CONVERTER ARCHITECTURE

A DC-DC boost converter is a power electronic device made up from inductors, capacitors, switching devices, and diodes, and it is used to convert the low voltage of a DC source to a higher voltage that is required by the load. A typical boost converter architecture is depicted in Figure 1. Both the source and the inductor are represented by ideal elements in series with an internal resistor, while the capacitor is represented as ideal, disregarding its internal resistance.

The inductor and the capacitor operate as a low-pass filter, with a cut-off frequency capable to remove the undesired harmonics of the switching frequency. The inductor and the capacitor values should be selected in such a way that the maximum voltage and current ripple will not exceed the desired thresholds [3]. The IGBT is acting like a switch; during the OFF state it is an open circuit, and during the ON state it is modeled by a forward-biased diode junction in series with an effective resistance R_{on} . Similarly, the diode in reverse-bias junction behaves like an open circuit, and in forward-bias junction it is modeled by a voltage source V_D in series with an effective resistance R_{on} . Finally, the load is represented by an effective load resistance R .

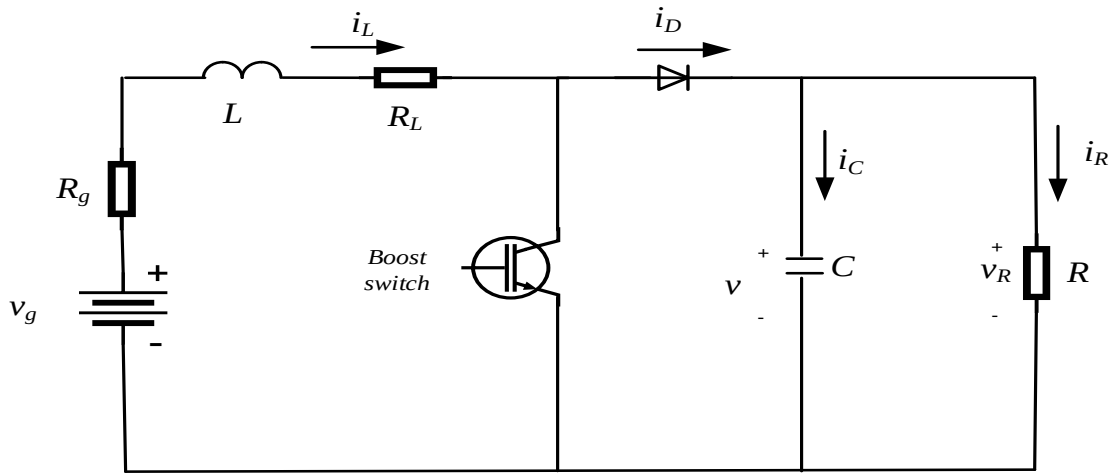


Figure 1. DC-DC Boost Converter Architecture

The circuit during the ON state of operation of IGBT is depicted in Figure 2.

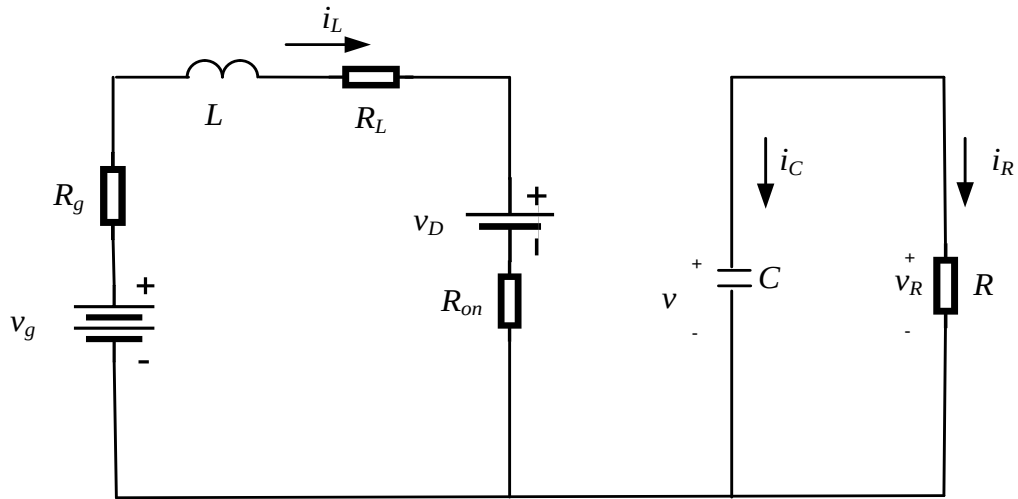


Figure 2. Boost Converter during the ON State of Operation of IGBT

The circuit during the OFF state of operation of IGBT is depicted in Figure 3.

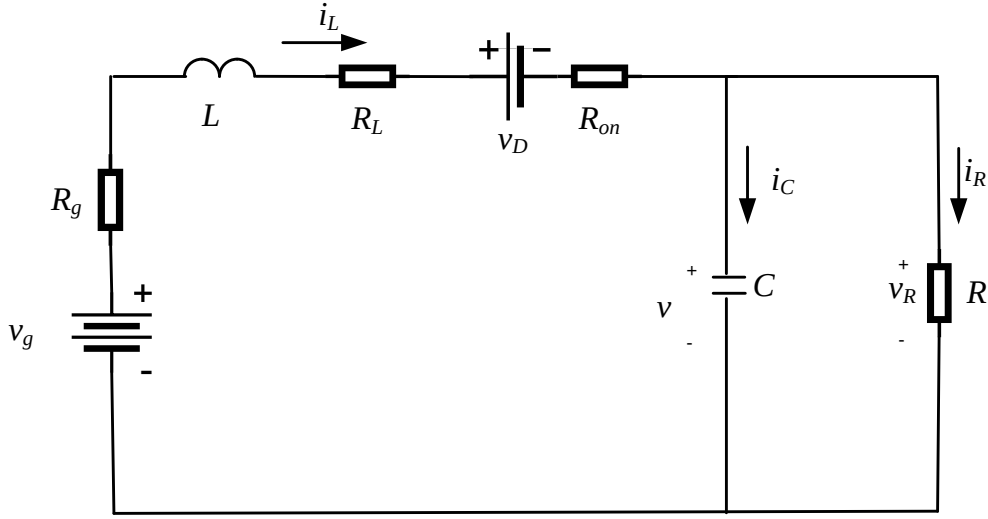


Figure 3. Boost Converter during the OFF State of Operation of IGBT

The duty cycle is defined as the percentage of the period in which the switch is ON, and it is equal to the on-time divided by the period of the switch

$$Dt = \frac{t_{on}}{T_{sw}}, \quad (1)$$

where Dt is the duty cycle, t_{on} is the on-time of the switch, and T_{sw} is the switching period.

The boost converter IGBT duty cycle is equal to a constant value with the addition of the perturbations

$$d(t) = D + Dm \cdot \cos(\omega_m t) = D + \hat{d}(t), \quad (2)$$

where $d(t)$ is the IGBT duty cycle, D is a constant with value equal to the average duty cycle, Dm is a constant with value equal to the amplitude small-signal duty cycle, ω_m is the circular frequency of the small signal, and $\hat{d}(t)$ are the AC variations of the duty cycle.

The output spectrum contains harmonics of both the small-signal frequency and the switching frequency. The switching ripple is small in converters operating in CCM, so it is ignored, and only the AC variations of the waveforms will be considered here [2], [3].

B. MODES OF OPERATION

There are two possible modes of DC-DC boost converter operation [3]. The first is the continuous conduction mode of operation (CCM), and the second is the discontinuous conduction mode of operation (DCM) [3]. The CCM is the normal mode of a boost during operation [3]. During this mode there is only one direction of inductor current and capacitor voltage [3]. When the ripple of the inductor current or the capacitor voltage is large enough to result in a polarity change of the inductor current or the capacitor voltage, the DCM boost converter mode of operation happens [3]. The DCM typically occurs when the load resistance exceeds the critical value [2], [3].

$$R_{crit}(D) = \frac{2L}{D(1-D)^2 T_{sw}} \quad (3)$$

C. STATE-SPACE AVERAGING TECHNIQUE THEORY

The state-space averaging method is a control theory technique [3]. To apply this method, the converter is analyzed during the two subintervals of the switch [3]. At the ON position of the switch the boost converter can be simplified/reduced to a linear circuit [3]. Its state-space equations are of the following forms [3]

$$K_B \frac{dx(t)}{dt} = A_{B1}x(t) + B_{B1}u(t) \quad (4)$$

$$y(t) = C_{B1}x(t) + E_{B1}u(t) \quad (5)$$

Similarly, at the OFF position of the switch the boost converter is represented by a linear circuit with state-space equations of the following forms [3]

$$K_B \frac{dx(t)}{dt} = A_{B2}x(t) + B_{B2}u(t) \quad (6)$$

$$y(t) = C_{B2}x(t) + E_{B2}u(t) \quad (7)$$

At the equilibrium, the derivative of the states are equal to zero [3]

$$0 = A_B X + B_B U \quad (8)$$

$$Y = C_B X + E_B U, \quad (9)$$

where

K_B = constant gain matrix

$$A_B = DA_{B1} + D' A_{B2} \quad (10)$$

$$B_B = DB_{B1} + D' B_{B2} \quad (11)$$

$$C_B = DC_{B1} + D' C_{B2} \quad (12)$$

$$E_B = DE_{B1} + D' E_{B2}, \quad (13)$$

The values of the equilibrium states can be calculated by equations (14) and (15) [21]

$$X = -A_B^{-1} B_B U \quad (14)$$

$$Y = (-C_B A_B^{-1} B_B + E_B) U. \quad (15)$$

The linearized AC small-signal model is represented by the following equations [3].

$$K \frac{d\hat{x}(t)}{dt} = A_B \hat{x}(t) + B_B \hat{u}(t) + \left[(A_{B1} - A_{B2}) X + (B_{B1} - B_{B2}) U \right] \hat{d}(t) \quad (16)$$

$$\hat{y}(t) = C_B \hat{x}(t) + E_B \hat{u}(t) + \left[(C_{B1} - C_{B2}) X + (E_{B1} - E_{B2}) U \right] \hat{d}(t) \quad (17)$$

D. APPLYING STATE-SPACE AVERAGING TECHNIQUES

When applying the state-space averaging technique, two steps are required. The first step is to select the states and then perform an analysis of the circuit during the two subintervals of the switch. The second step is to derive the small-signal transfer function.

1. Circuit Analysis

The circuit is analyzed to apply the state-space averaging technique. The capacitor voltage and the inductor current are selected as states. The DC sources are selected as input. Finally, two different outputs are selected. The first output is the load voltage. The second output is the inductor current

$$x(t) = \begin{bmatrix} i_L(t) \\ v_C(t) \end{bmatrix} = \begin{bmatrix} i(t) \\ v(t) \end{bmatrix} \quad (18)$$

$$u(t) = \begin{bmatrix} v_g(t) \\ V_D \end{bmatrix} \quad (19)$$

$$y_1(t) = v_R(t) = v(t) \quad (20)$$

$$y_2(t) = i(t). \quad (21)$$

First, the load voltage is used as the output of the model. During the first subinterval (switch in position 1) the boost converter can be simplified to the linear circuit in Figure 2.

$$\begin{bmatrix} L & 0 \\ 0 & C \end{bmatrix} \frac{d}{dt} \begin{bmatrix} i(t) \\ v(t) \end{bmatrix} = \begin{bmatrix} -R_i & 0 \\ 0 & -\frac{1}{R} \end{bmatrix} \begin{bmatrix} i(t) \\ v(t) \end{bmatrix} + \begin{bmatrix} 1 & -1 \\ 0 & 0 \end{bmatrix} \begin{bmatrix} v_g(t) \\ V_D \end{bmatrix} \quad (22)$$

$$\begin{bmatrix} v_R(t) \end{bmatrix} = \begin{bmatrix} 0 & 1 \end{bmatrix} \begin{bmatrix} i(t) \\ v(t) \end{bmatrix} + \begin{bmatrix} 0 & 0 \end{bmatrix} \begin{bmatrix} v_g(t) \\ V_D \end{bmatrix} \quad (23)$$

where:

$$R_i = R_g + R_L + R_{on} \quad (24)$$

Similarly, during the second subinterval (switch in position 2) the converter can be simplified to the linear circuit in Figure 3.

$$\begin{bmatrix} L & 0 \\ 0 & C \end{bmatrix} \frac{d}{dt} \begin{bmatrix} i(t) \\ v(t) \end{bmatrix} = \begin{bmatrix} -R_i & -1 \\ 1 & -\frac{1}{R} \end{bmatrix} \begin{bmatrix} i(t) \\ v(t) \end{bmatrix} + \begin{bmatrix} 1 & -1 \\ 0 & 0 \end{bmatrix} \begin{bmatrix} v_g(t) \\ V_D \end{bmatrix} \quad (25)$$

$$\begin{bmatrix} v_R(t) \end{bmatrix} = \begin{bmatrix} 0 & 1 \end{bmatrix} \begin{bmatrix} i(t) \\ v(t) \end{bmatrix} + \begin{bmatrix} 0 & 0 \end{bmatrix} \begin{bmatrix} v_g(t) \\ V_D \end{bmatrix} \quad (26)$$

Assuming CCM, the average state-space model [2], [3] is represented as follows

$$\begin{bmatrix} L & 0 \\ 0 & C \end{bmatrix} \frac{d}{dt} \begin{bmatrix} i(t) \\ v(t) \end{bmatrix} = \begin{bmatrix} -R_i & -D' \\ D' & -\frac{1}{R} \end{bmatrix} \begin{bmatrix} i(t) \\ v(t) \end{bmatrix} + \begin{bmatrix} 1 & -1 \\ 0 & 0 \end{bmatrix} \begin{bmatrix} v_g(t) \\ V_D \end{bmatrix} \quad (27)$$

$$\begin{bmatrix} v_r(t) \end{bmatrix} = \begin{bmatrix} 0 & 1 \end{bmatrix} \begin{bmatrix} i(t) \\ v(t) \end{bmatrix} + \begin{bmatrix} 0 & 0 \end{bmatrix} \begin{bmatrix} v_g(t) \\ V_D \end{bmatrix}, \quad (28)$$

where D is the duty cycle and D'=1-D is the duty cycle complement value.

The steady state equilibrium values can be derived from the above state-space model [3], [21]

$$\begin{bmatrix} I \\ V \end{bmatrix} = \frac{1}{\frac{R_i}{R} + D'^2} \begin{bmatrix} \frac{1}{R} & -\frac{1}{R} \\ D' & -D' \end{bmatrix} \begin{bmatrix} V_g \\ V_D \end{bmatrix}. \quad (29)$$

Since, the desired voltage output is known, the required duty cycle can be calculated

$$D' = \frac{(V_g - V_D) + \sqrt{(V_g - V_D)^2 - 4V^2 \frac{R_i}{R}}}{2V}. \quad (30)$$

The small-signal AC state equations can be derived from the above state-space model [3]. The diode forward-voltage drop and IGBT forward voltage are modeled by a constant voltage source V_D , so there is no AC variation in this source.

$$\frac{d}{dt} \begin{bmatrix} \hat{i}(t) \\ \hat{v}(t) \end{bmatrix} = \begin{bmatrix} -\frac{R_i}{L} & -\frac{D'}{L} \\ \frac{D'}{C} & -\frac{1}{RC} \end{bmatrix} \begin{bmatrix} \hat{i}(t) \\ \hat{v}(t) \end{bmatrix} + \begin{bmatrix} \frac{1}{L} \\ 0 \end{bmatrix} \hat{v}_g(t) + \begin{bmatrix} \frac{V}{L} \\ -\frac{I}{C} \end{bmatrix} \hat{d}(t) \quad (31)$$

$$\begin{bmatrix} \hat{v}_r(t) \end{bmatrix} = \begin{bmatrix} 0 & 1 \end{bmatrix} \begin{bmatrix} \hat{i}(t) \\ \hat{v}(t) \end{bmatrix} \quad (32)$$

Similarly, using the inductor current as the output, the corresponding small-signal state equations are the following

$$\frac{d}{dt} \begin{bmatrix} \hat{i}(t) \\ \hat{v}(t) \end{bmatrix} = \begin{bmatrix} -\frac{R_i}{L} & -\frac{D'}{L} \\ \frac{D'}{C} & -\frac{1}{RC} \end{bmatrix} \begin{bmatrix} \hat{i}(t) \\ \hat{v}(t) \end{bmatrix} + \begin{bmatrix} \frac{1}{L} \\ 0 \end{bmatrix} \hat{v}_g(t) + \begin{bmatrix} \frac{V}{L} \\ -\frac{I}{C} \end{bmatrix} \hat{d}(t) \quad (33)$$

$$\begin{bmatrix} \hat{i}(t) \end{bmatrix} = \begin{bmatrix} 1 & 0 \end{bmatrix} \begin{bmatrix} \hat{i}(t) \\ \hat{v}(t) \end{bmatrix} \quad (34)$$

The observability matrix for a second-order system is given by Equation (35) [21]

$$O_M = \begin{bmatrix} C_B \\ C_B A_B \end{bmatrix}. \quad (35)$$

The controllability matrix for a second-order system is given by Equation (36) [21]

$$C_M = \begin{bmatrix} B_B & A_B B_B \end{bmatrix}. \quad (36)$$

Since the controllability matrix and the observability are full rank, the system is both controllable and observable.

2. Small-Signal Transfer Function

The small-signal state-space model is used to calculate the necessary transfer functions. A transfer function of a linear system is a mathematical model that gives the output for every possible value of the input. The duty cycle control to output load voltage transfer function $G_{vd}(s)$ can be found by setting the voltage variations equal to zero [3]

$$G_{vd}(s) = \frac{\hat{v}(s)}{\hat{d}(s)} \Big|_{\hat{v}_g(s)=0} = -\frac{1}{C} \frac{V_g - V_D}{R_i + RD^2} \frac{s - \frac{(D'^2 R - R_i)}{L}}{s^2 + s \left(\frac{R_i}{L} + \frac{1}{RC} \right) + \frac{D'^2}{CL} + \frac{R_i}{RCL}}. \quad (37)$$

where s is the Laplace operator.

The right half zero of the transfer function $G_{vd}(s)$ can be found by setting up the numerator equal to zero

$$\omega_{RHP} = \frac{D'^2 R - R_i}{L}. \quad (38)$$

The transfer function $G_{vd}(s)$ was calculated using the parameter values from Table 1. The corresponding gain and phase (Bode) diagram is depicted in Figure 5. Similarly, the corresponding root locus is depicted in Figure 6. The root locus design procedure of open-loop transfer function $G_{vd}(s)$ is depicted in Figure 4 [22].

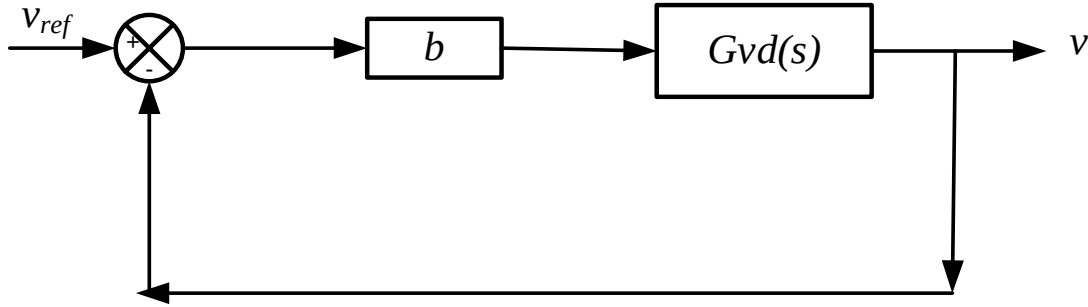


Figure 4. Root Locus Design Procedure

To design the root locus, an open-loop gain, b , is added to the system [22]. Then the closed-loop pole trajectories are plotted in the complex plane as this gain b varies in a continuous range of values [22]. To plot the root locus of the transfer function $G_{vd}(s)$, the closed-loop poles are the roots of [22]

$$rl(s) = 1 + bG_{vd}(s). \quad (39)$$

Table 1. Parameter Values for the Simulink/MATLAB Model

Device	Value
V_g	100 V
V_D	0.7 V
L	940 μ H
C	990 μ F
R_L	0.1 Ω
R_{on}	0.2 Ω
R_g	0.1 Ω
R	50 Ω
f_s	15 kHz

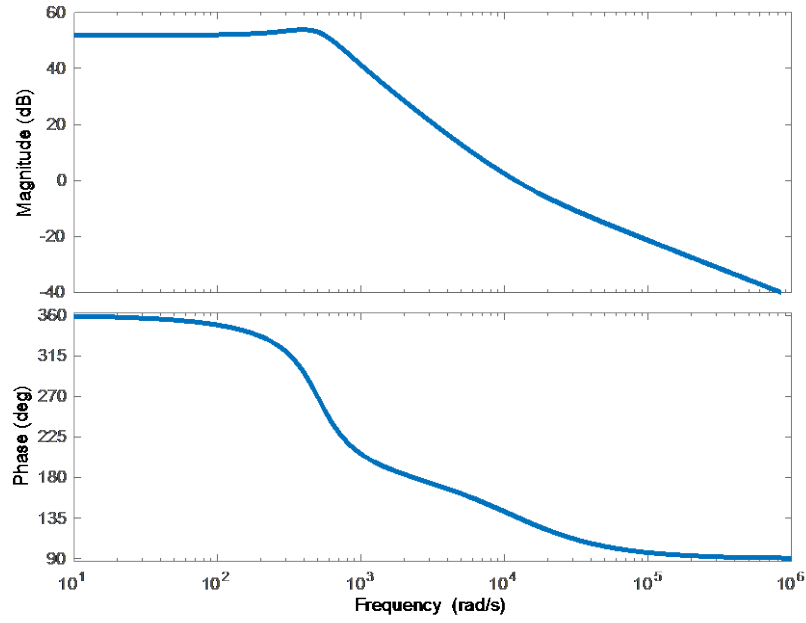


Figure 5. Gain and Phase of Open-Loop Transfer Function $G_{vd}(s)$

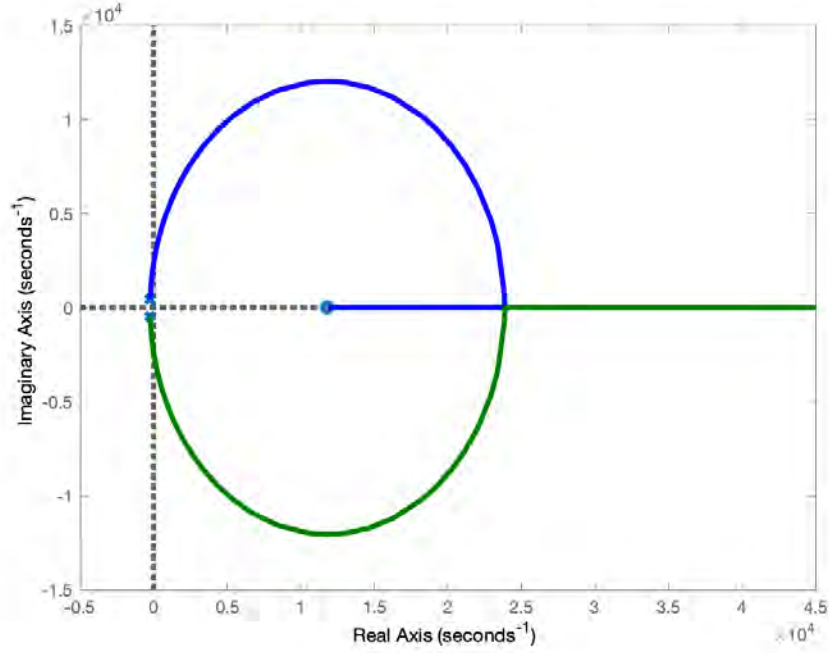


Figure 6. Root Locus of Open-Loop Transfer Function $G_{vd}(s)$

The input line voltage to output load voltage transfer function $G_{vg}(s)$ can be found by setting the duty cycle variations equal to zero [3]

$$G_{vg}(s) = \frac{\hat{v}(s)}{\hat{v}_g(s)} \Big|_{\hat{d}(s)=0} = \frac{\frac{D'}{LC}}{s^2 + s \left(\frac{R_i}{L} + \frac{1}{RC} \right) + \frac{D'^2}{CL} + \frac{R_i}{RCL}}. \quad (40)$$

The transfer function $G_{vg}(s)$ was calculated using the parameter values from Table 1. The corresponding gain and phase (Bode) diagram is depicted in Figure 7. Similarly, the corresponding root locus is depicted in Figure 8.

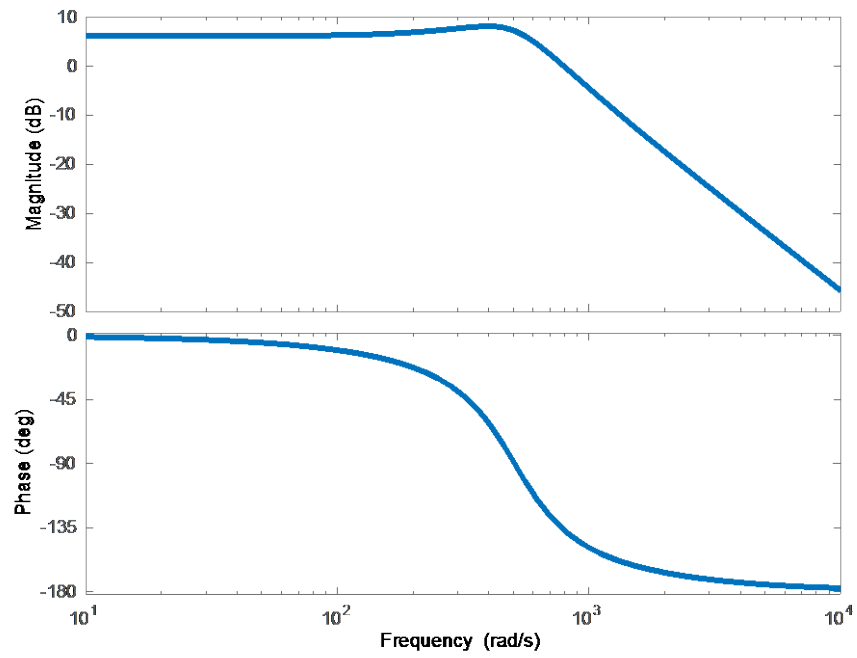


Figure 7. Gain and Phase of Open-Loop Transfer Function $G_{vg}(s)$

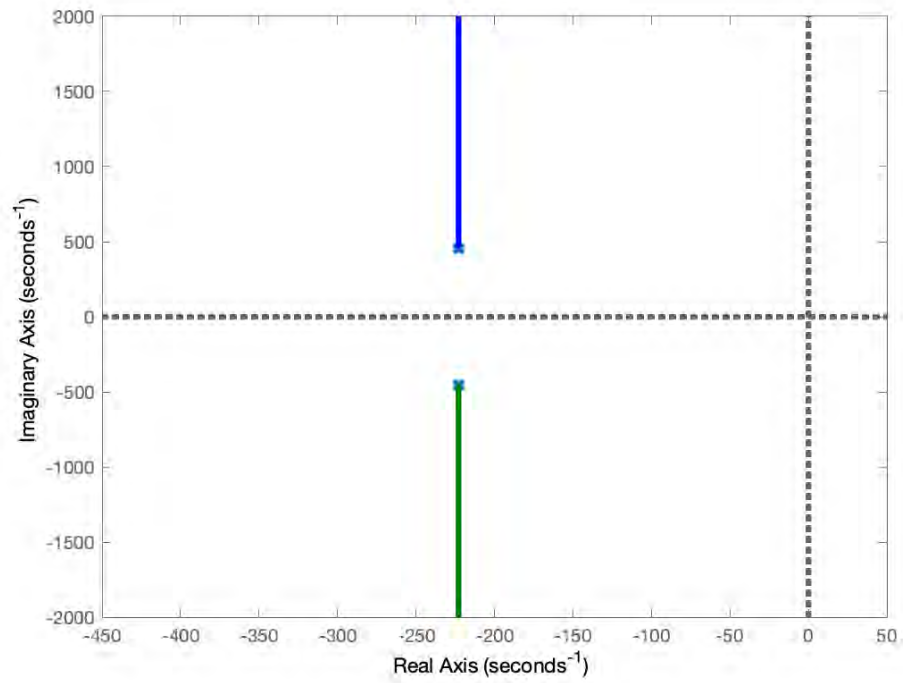


Figure 8. Root Locus of Open-Loop Transfer Function $G_{vg}(s)$

The duty cycle control to output inductor current transfer function $G_{id}(s)$ can be found by setting the voltage variations equal to zero [3]

$$G_{id}(s) = \frac{\hat{i}(s)}{\hat{d}(s)} \Big|_{\hat{v}_g(s)=0} = \frac{1}{L} \frac{D' R (V_g - V_D)}{R_i + R D'^2} \frac{s + \frac{2}{RC}}{s^2 + s \left(\frac{R_i}{L} + \frac{1}{RC} \right) + \frac{D'^2}{CL} + \frac{R_i}{RCL}}. \quad (41)$$

The transfer function $G_{id}(s)$ was calculated using the parameter values from Table 1. The corresponding gain and phase (Bode) diagram is depicted in Figure 9. Similarly, the corresponding root locus is depicted in Figure 10.

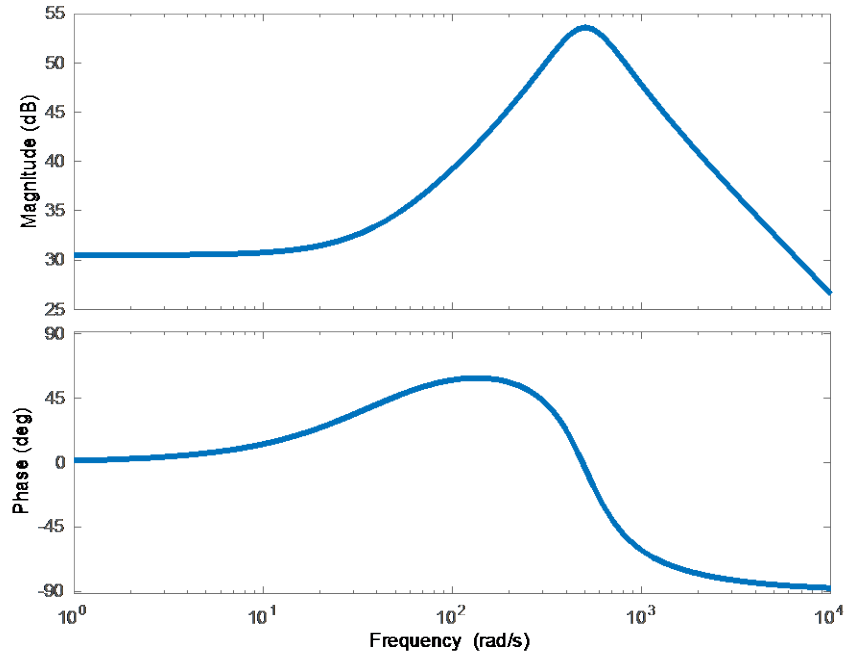


Figure 9. Gain and Phase of Open-Loop Transfer Function $G_{id}(s)$

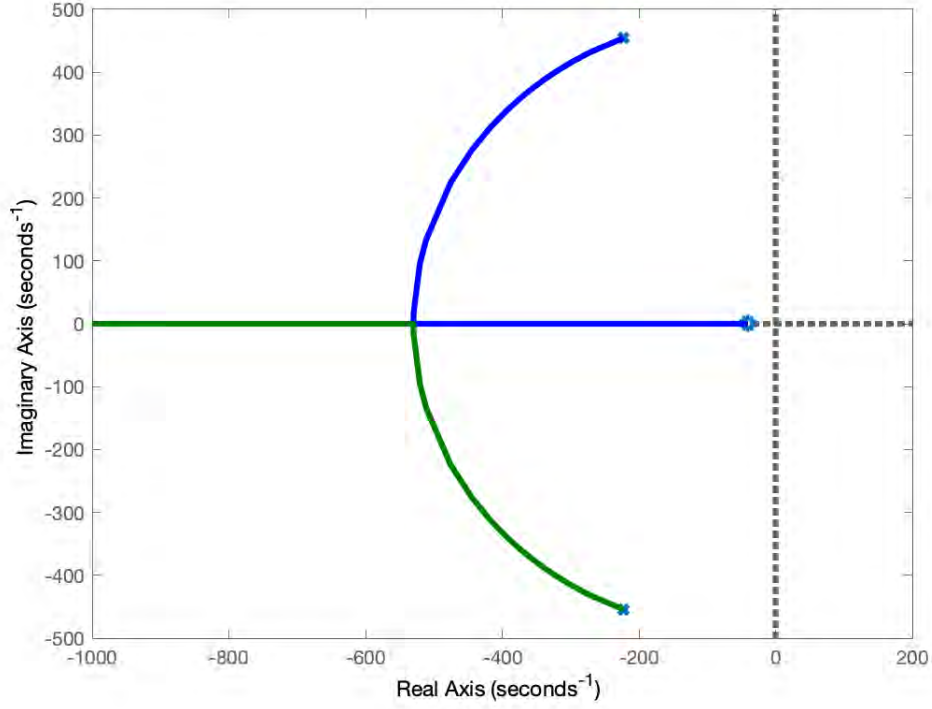


Figure 10. Root Locus of Open-Loop Transfer Function $G_{id}(s)$

The input line voltage to output inductor current transfer function $G_{ig}(s)$ can be found by setting the duty cycle variations equal to zero [3]

$$G_{ig}(s) = \frac{\hat{i}_L(s)}{\hat{v}_g(s)} \Big|_{\hat{d}(s)=0} = \frac{1}{L} \frac{s + \frac{1}{RC}}{s^2 + s \left(\frac{R_i}{L} + \frac{1}{RC} \right) + \frac{D^2}{CL} + \frac{R_i}{RCL}} \quad (42)$$

The transfer function $G_{ig}(s)$ was calculated using the parameter values from Table 1. The corresponding gain and phase (Bode) diagram is depicted in Figure 11. Similarly, the corresponding root locus is depicted in Figure 12.

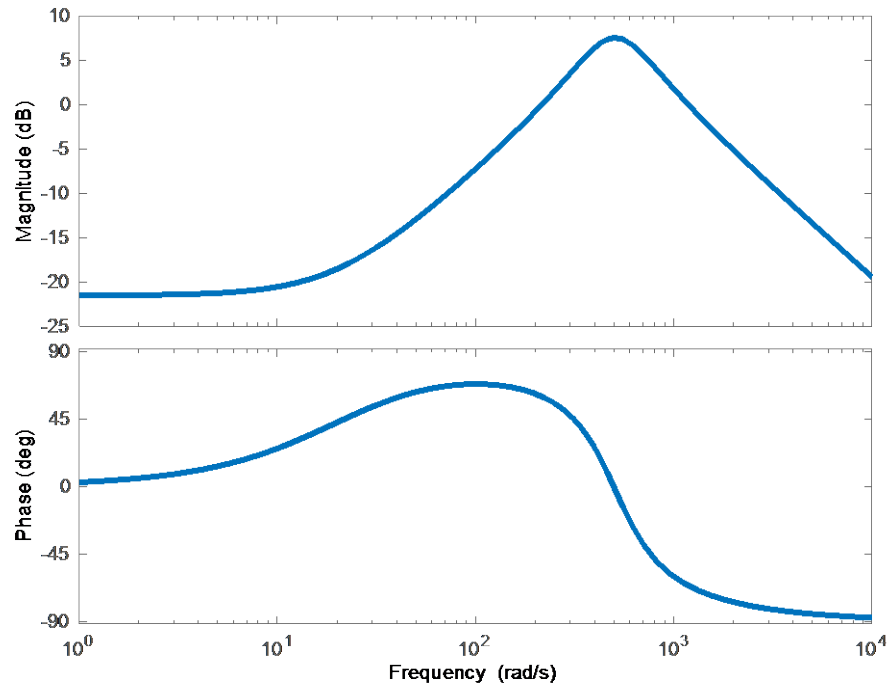


Figure 11. Gain and Phase of Open-Loop Transfer Function $G_{ig}(s)$

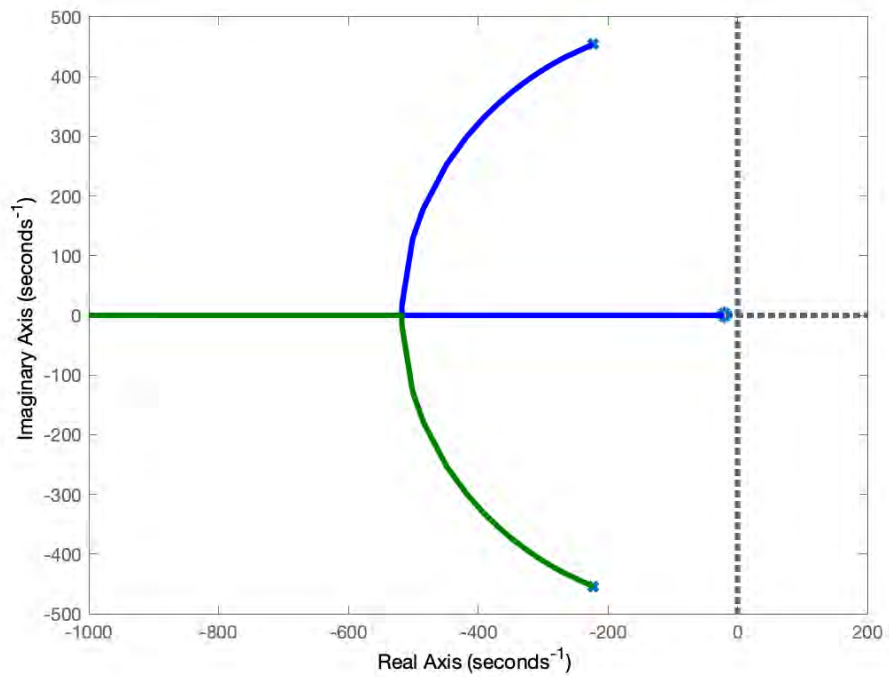


Figure 12. Root Locus of Open-Loop Transfer Function $G_{ig}(s)$

In the next chapter, the transfer functions will be used to set up control methodologies to maintain a constant output, no matter the variations in the load or the input voltage.

3. Boost Converter Simulink Model

A Simulink model of a non-ideal boost converter can be created using (27). A block diagram of this boost converter is shown in Figure 13 .

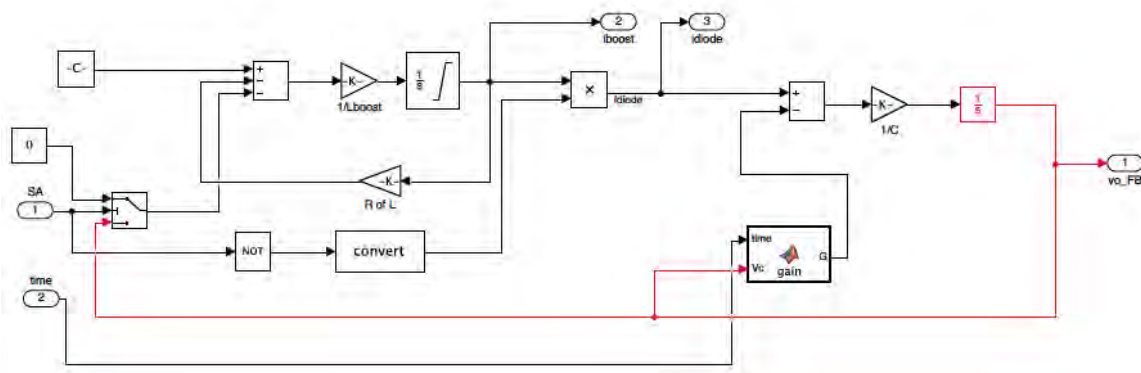


Figure 13. Boost Converter Simulink Model

The implementation of the Simulink model was a straightforward process. The input of the system was the switch position (SA) and the simulation time. The load was simulated by a separate MATLAB function that set different load resistors depending on the simulation time. This approach achieved the desired step load changes to evaluate the performance of each controller.

III. BOOST CONVERTER CONTROL METHODOLOGIES AND DESIGN ANALYSIS

During the operation of the boost converter, the output voltage must be kept constant, no matter the variations in the input voltage or the change of the load. This is accomplished with a negative-feedback control loop. The feedback loop attenuates any output voltage variations caused by the input voltage variation or by the load current variation [3].

Since there is no control on the input voltage variations, the duty cycle is to be used by the controller to maintain a constant output. The PI controller can be used to improve the performance of a control system [21], but it is difficult for the PI controller used in a boost converter to exhibit the desired performance [6], [8]. The major problem in controlling a boost converter is caused by a right half plane zero of the transfer function $G_{vd}(s)$ [4]. The frequency at which the right half plane zero occurs is given by (38) and it is related to the effective value of the filter inductance, the parasitic resistance, and the load resistance. This right half plane zero causes a phase reversal [3]. The magnitude response of the right half plane zero is similar to the response of the left half plane zero, but it exhibits a negative phase because it acts as a time delay [3]. Unlike the poles of a system that can be moved by feedback, the zeros can only be moved by pole/zero cancellation, which, in the case of a zero on the right half plane, would add instability in the system. Therefore, in this thesis, different control strategies are implemented, and their performance is evaluated.

A. K-FACTOR CONTROL METHODOLOGY

K-factor compensator can be used to effectively control a boost converter. This section describes regular K-factor control methodology, along with two innovative modifications to improve the performance of the regular methodology.

1. Regular K-Factor Control Methodology

The K-factor originated as a tool for the synthesis of amplifiers [4], and there are three basic amplifier types. Type 1 is used to control loops where there is minimum phase shift of the modulator [4]. Type 2 is used to compensate loops where the phase shift of the

plant is approximately -90° at the desired crossover frequency [4]. Type 3 is used to compensate loops where the phase shift is approximately -180° at the desired crossover frequency [4]. A type 3 amplifier is an amplifier with a pole at the origin and two zeros/poles pairs [4]. In order to regulate a boost converter, a type 3 amplifier is required, and it is defined as the square root of the ratio of the double pole frequency to the double zero frequency [4]. The compensator is of the form [4]

$$G(s) = \frac{B_c(s)}{sA_c(s)} = K_c \frac{(s + \omega_z)^2}{s(s + \omega_p)^2}, \quad (43)$$

$$\text{with } \omega_z = \frac{\omega_{GC}}{\sqrt{K}}, \omega_p = \omega_{GC}\sqrt{K}, K_c = \frac{K\omega_{GC}}{|G_{vd}(j\omega_{GC})|}, K = \left(\tan(\Phi_{Boost} + 45^\circ)\right)^2$$

where, ω_{GC} = Selected crossover frequency

K = K factor parameter

$$\Phi_{Boost} = M - P - 90^\circ$$

M = Desired phase margin (degrees)

P = Modulator phase shift (degrees)

Figure 14 shows a closed-loop output voltage regulation of a boost converter using regular K -factor control method.

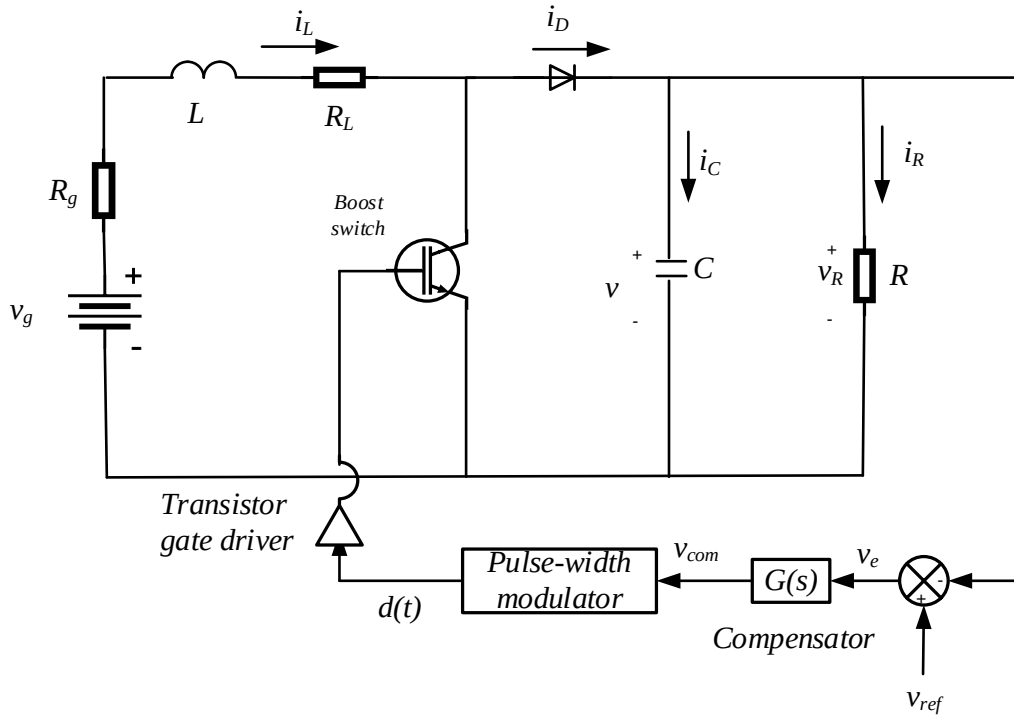


Figure 14. Closed-Loop Output Voltage Regulation of a Boost Converter Using Regular K-Factor Control Method

The advantages of the K-factor control method compared with a PI are that it has faster transient response due to the phase boost from the two zero-pole pair, and it is more stable [5].

A K-factor controller is designed by selecting the crossover frequency and the desired phase boost. The crossover frequency should be chosen to be as high as possible, since this will result in a faster transient response [4]. The higher value of the crossover frequency, however, is bounded by the frequency of the right half plane zero in the Bode plot. The lower value of the crossover frequency is bounded by the resonant frequency, while an appropriate value corresponds to frequency with phase of the transfer function $G_{vd}(s)$ close to 180° . The second selection is the desired phase margin. A high value of desired phase margin will result in a very stable system with slow transient response, while the opposite occurs for a low selected value; a desired phase margin of 60° will be used, since it achieved a sufficient level of both fast transient response and stability [4].

Figure 15 shows an optimization algorithm for the regular K-factor compensator.

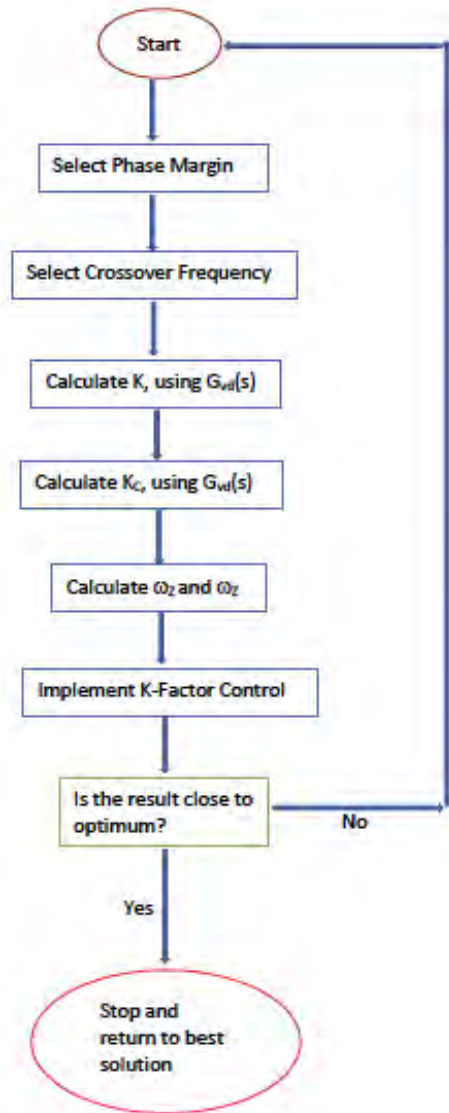


Figure 15. Regular K-Factor Compensator Optimization Algorithm

The boost converter transfer function of the $G_{vd}(s)$ and the K-factor compensator transfer function $G(s)$ were calculated using the parameter values of Table 2. The frequency response of the compensated system is depicted in Figure 16.

Table 2. Parameter Values for the Simulink/MATLAB Model

Device	Value
V_g	100 V
V_D	0.7 V
L	940 μ H
C	990 μ F
R_L	0.1 Ω
R_{on}	0.2 Ω
R_g	0.1 Ω
R	50 Ω
f_s	15 kHz
ω_z (rad/sec)	320.98
ω_p (rad/sec)	27207.99
K_c	20945.41
K	84.76

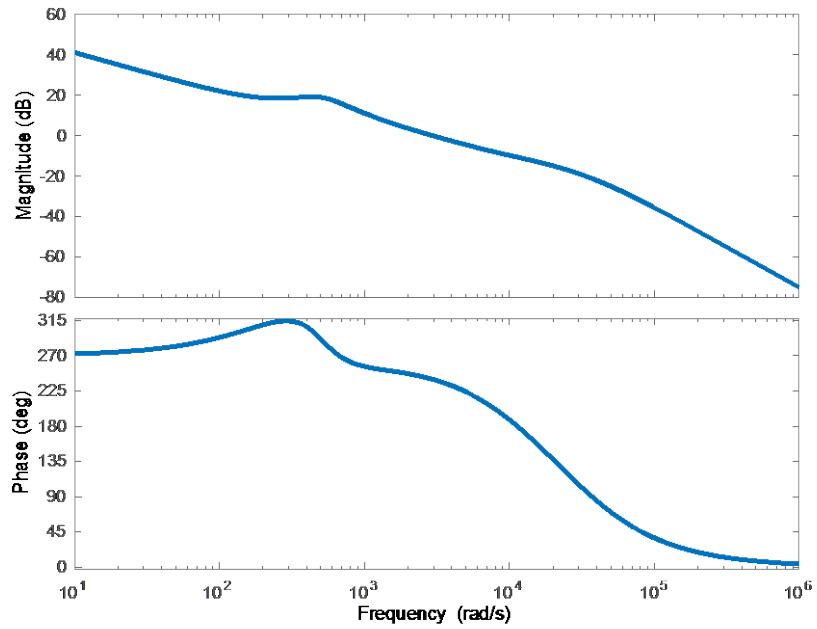


Figure 16. Gain and Phase of Open-Loop Transfer Boost Converter Function $G_{vd}(s)$ in Series with K-Factor Compensator

2. Modifications of the Regular K-Factor Control Methodology

During this research two modifications of the regular K-factor control were implemented. The first modification (Mod 1 K-factor) was implemented by adding a current sensor on the inductor and switch between controllers when the inductor current equaled zero. The second modification (Mod 2 K-factor) was implemented by switching between controllers when the output voltage exceeded a predetermined value.

The first K-factor control modification is depicted in Figure 17. The current sensor was used to detect if the inductor current went to zero, which is equivalent to the boost converter operating in DCM mode. This can occur either when the load resistor exceeds the critical load value [3] or the boost converter switches to DCM mode of operation during transient response due to the occurrence of a big step load change. To implement this modification at first, a comparator was added to detect when the inductor current goes to zero. Then for the duration where the inductor was detected to be zero, a different K-factor compensator was selected, with parameters that corresponded to a DCM load value.

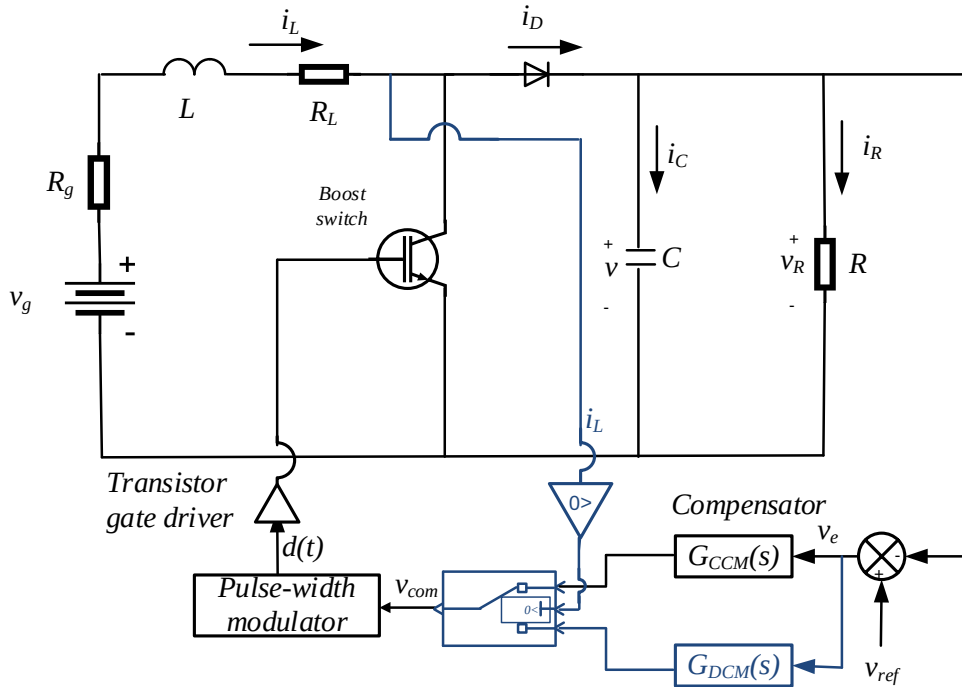


Figure 17. Closed-Loop Output Voltage Regulation of a Boost Converter Using the Mod1 K-Factor Control Method

Given the added complexity and the high cost of a Hall Effect current sensor, a second modification was implemented. This time, only one voltage sensor was used, as it is depicted in Figure 18. When a step load change to a lower value occurs, the output voltage is increased during the transient response. The goal of this modification is to control the peak voltage by using a faster compensator, corresponding to a different K-factor. At first, a comparator was added to detect when the transient response occurs; this was implemented by comparing the output voltage with a predetermined value. The predetermined value is a value slightly higher than the desired output. Then, for the duration where the output voltage exceeds that value, a different K-factor compensator was selected, with parameters that correspond to a DCM load value.

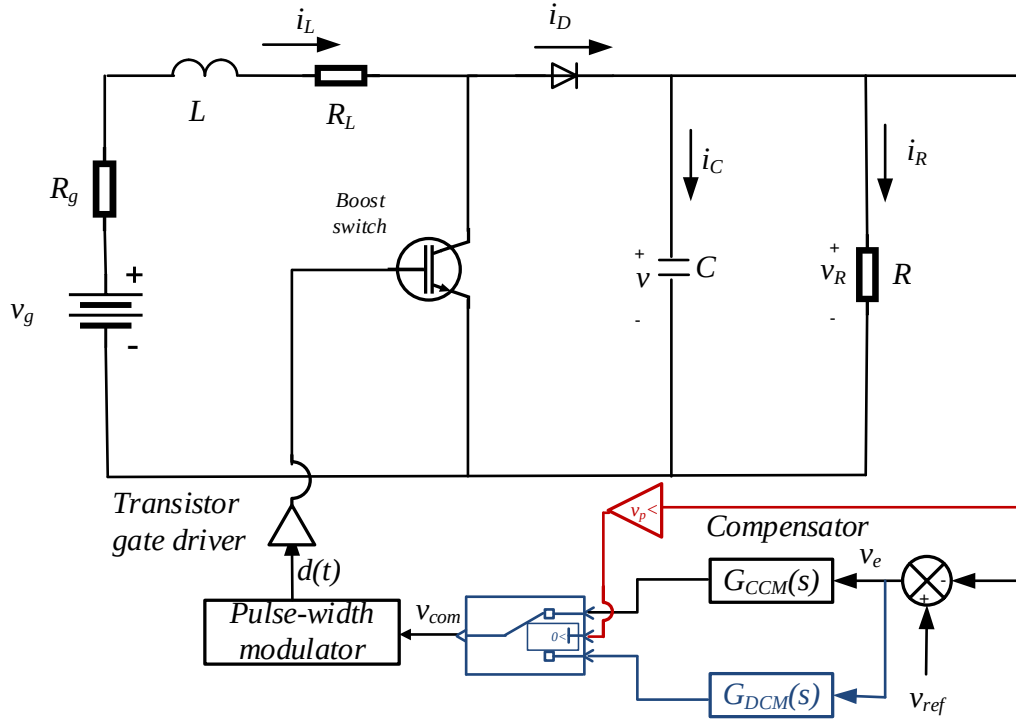


Figure 18. Closed-Loop Output Voltage Regulation of a Boost Converter Using the Mod 2 K-Factor Control Method

3. K-Factor Control Compensator Simulink Model

The K-factor controller was implemented in Simulink. The design was a straightforward progress. The regular K-factor compensator model is depicted in Figure 19.

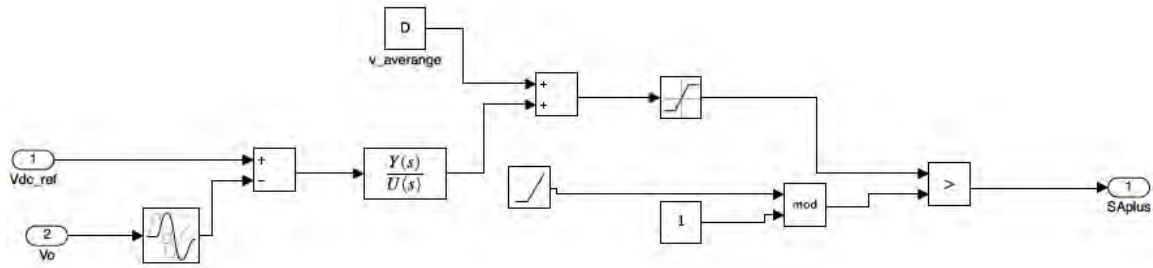


Figure 19. Regular K-Factor Compensator Simulink Model

The time delay block was added due to the A/D converter. An A/D converter is a device that takes as an input an analog signal and converts it into digital format. Using an A/D converter introduces a time delay due to the conversion process. The top saturation block limited the duty cycle to its maximum value. The average value added to the controller output was the initial value for the controller integrator. Finally, the PWM signal driving the switch was generated by comparing a periodic sawtooth signal with a threshold.

The regular K-factor compensator Simulink block diagram was modified to get the corresponding Mod 1 K-factor compensator Simulink block diagram, as depicted in Figure 20.

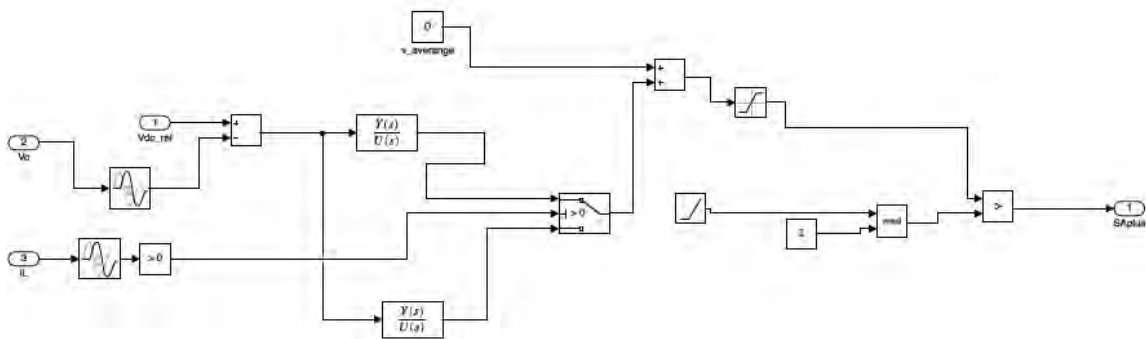


Figure 20. Mod 1 K-Factor Control Compensator Simulink Model

This implementation added a second input at the controller, the inductor current. A comparator was used to compare the inductor current with zero. The output of the comparator was the control signal for a switch, which selected between two controllers. The first controller was set up with parameter values that corresponded to CCM, and the second was set up with parameter values that corresponded to DCM.

The Mod 2 K-factor compensator Simulink model is depicted in Figure 21.

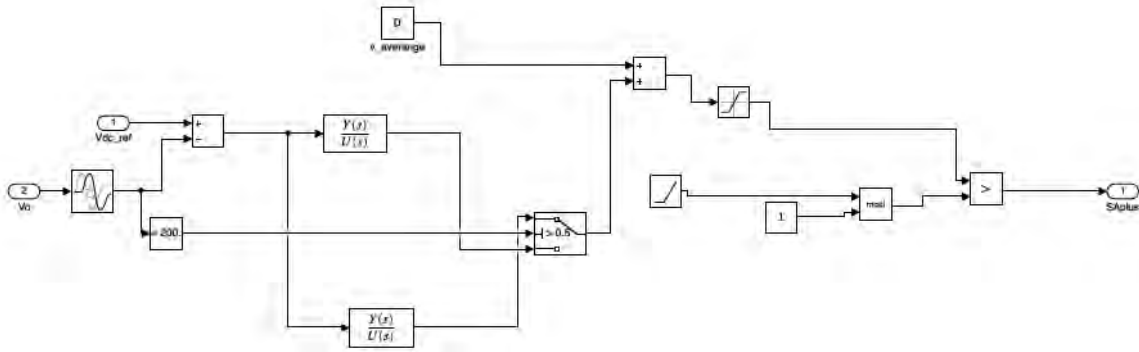


Figure 21. Mod 2 K-Factor Control Compensator Simulink Model

In this modification the only input in the controller was the output voltage. Similar to the first modification, a comparator was used to create the control signal for the switch that selects between two controllers, but this time the comparator took as an input the output voltage and compared it with a predetermined value to detect if a transient due to load reduction occurred.

4. K-Factor Control Methodologies Comparison

A simulation was executed to validate the effectiveness of the proposed modifications. The effectiveness of each version of the K-factor compensator was evaluated for three different step load changes. The parameters that were used in Simulink are shown in Table 3. The gains of the K-factor control are shown in Table 4. The CCM gains refer to the gains that were used by the regular K-factor compensator. The modified K-factor controller used the CCM parameters during the operation and when necessary switched to a compensator with parameter values equal to the DCM ones.

Table 3. Parameter Values for the Simulink/MATLAB Model

Device	Value
V_g	100 V
V_D	0.7 V
L	940 μ H
C	990 μ F
R_L	0.1 Ω
R_{on}	0.2 Ω
R_g	0.1 Ω
R	50 Ω
f_s	15 kHz
A/D time delay	1 μ sec

Table 4. K-factor Compensator Parameters

K-Factor		
Gains	CCM	DCM
ω_z (rad/sec)	320.98	1080.04
ω_p (rad/sec)	27207.99	3902595.88
K_c	20945.41	8342236824.4
K	84.76	3613.37

The first step load change was executed for changing the load resistance from 50 Ω to 185 Ω . All the three versions of the K-factor controller had similar performance. The transient response of each controller is depicted Figure 22.

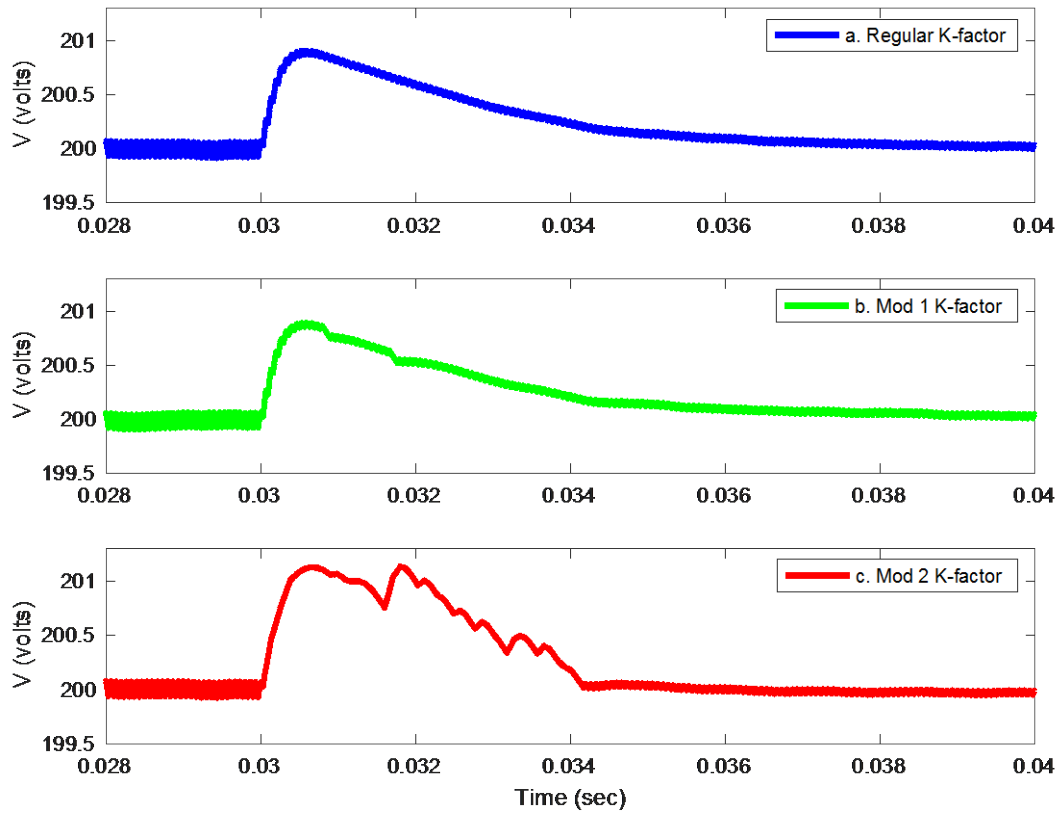


Figure 22. Closed-Loop Step Response for Load Change from $50\ \Omega$ to $185\ \Omega$: (a) Regular K-Factor Control, (b) Mod 1 K-Factor Control, (c) Mod 2 K-Factor Control

The load resistor changed back to $50\ \Omega$. When the system reached steady state, the controller performance was evaluated for another step load change. This step load change was executed for changing the load resistance from $50\ \Omega$ to $500\ \Omega$. The two modified versions of the K-factor performed much better than the regular one. The transient response of each controller is depicted in Figure 23. Actually, the Mod1 K-factor compensator resulted in the best performance, since that modification resulted in a smaller ripple, and faster transient response compared to the second one.

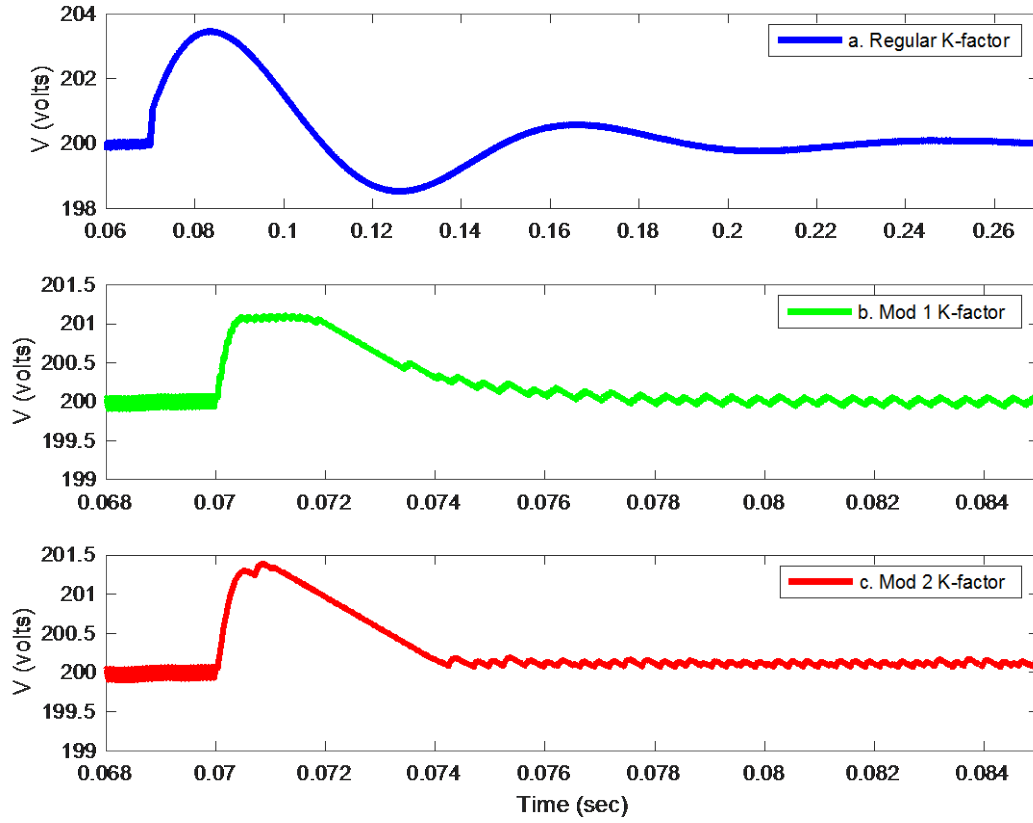


Figure 23. Closed-Loop Step Response for Load Change from 50 Ω to 500 Ω : (a) Regular K-Factor Control, (b) Mod 1 K-Factor Control, (c) Mod 2 K-Factor Control

Finally, the load resistance changed from 500 Ω to 50 Ω , and the compensators were evaluated. The two modified versions of the K-factor performed much better than the regular one, with the first modification resulting in the best performance. The transient response of each controller is depicted in Figure 24.

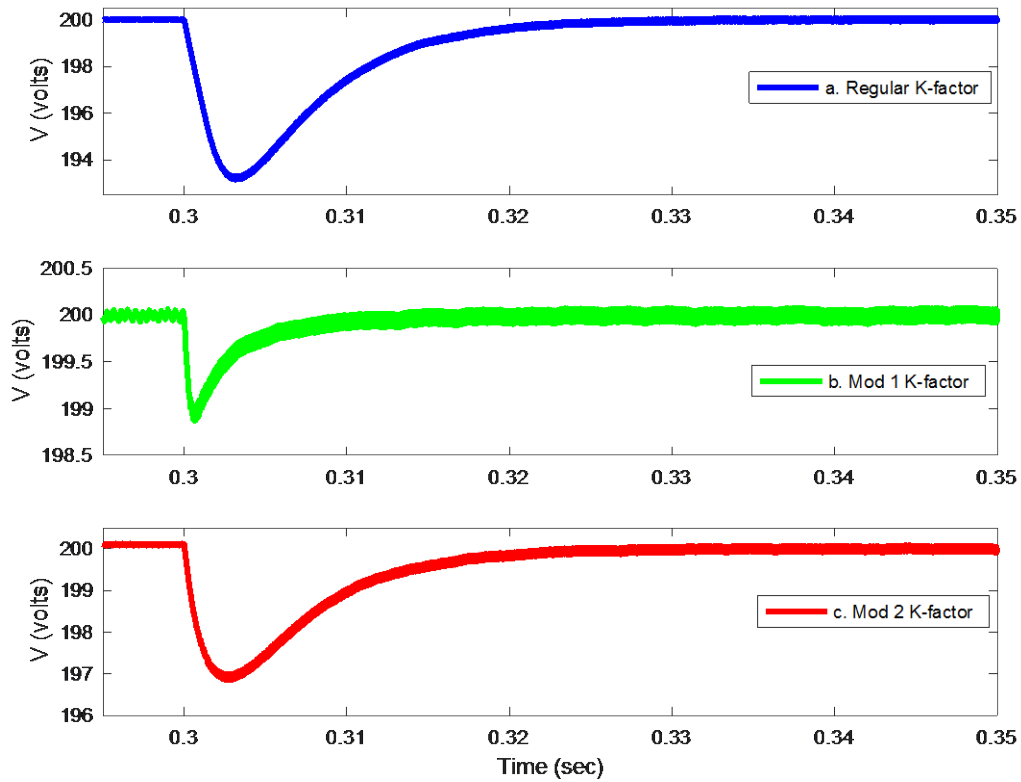


Figure 24. Closed-Loop Step Response for Load Change from $500\ \Omega$ to $50\ \Omega$: (a) Regular K-Factor Control, (b) Mod 1 K-Factor Control, (c) Mod 2 K-Factor Control

Taking all the above simulations into consideration, it is clear that the two modifications are a significant improvement with respect to the regular K-factor compensator. This is particularly evident when the load resistance changes to values that would cause the DC-DC boost converter to operate in DCM. The Mod 1 K-factor compensator, in particular, resulted in the best overall performance of the three cases, evaluating the maximum ripple and the time to reach steady state. On the other hand, however, the Mod 2 K-factor control method is simpler to implement, since it does not require a current sensor. An AC microgrid uses both AC and DC sources to provide the necessary power for the loads. In that microgrid an inverter is used to convert the DC bus voltage to required AC voltage. Given the added complexity and the high cost of a Hall Effect current sensor, and the fact that for an AC microgrid a DC voltage with a ripple of

3% can be considered acceptable as an input for the inverter, the Mod 2 K-factor is suitable to regulate the boost converter, without the extra cost of the current sensor.

B. VOLTAGE MODE CONTROL METHODOLOGY

The regular voltage mode control compensator can be used to control a boost converter, but due to load limitations two innovative modifications to improve the performance of the regular methodology are introduced.

1. Regular Voltage Mode Control Methodology

A description of the regular voltage mode control methodology as well as a simple algorithm to set the required parameters are introduced.

A closed-loop block diagram for the regulation of the output voltage of the DC-DC boost converter using a regular voltage mode control compensator is shown in Figure 25.

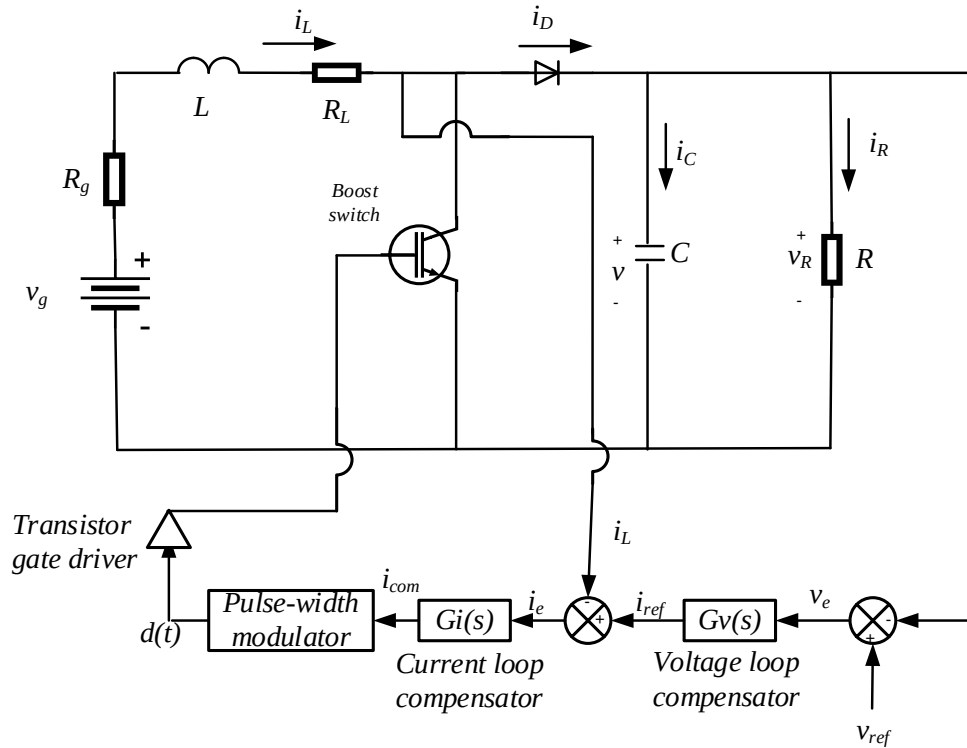


Figure 25. Closed-Loop Output Voltage Control Regulation of a Boost Converter Using Regular Voltage Mode Control Method

Both outer and inner loop consist of a PI controller. The input of the outer PI controller is the reference voltage, and the output is the reference inductor current, which is given from

$$i_{ref} = \left[\frac{k_{pv}s + k_{iv}}{s} \right] (V_{ref} - V). \quad (44)$$

The inductor current to output load voltage transfer function $G_{vi}(s)$ is used to set up this controller, as follows

$$G_{vi}(s) = \frac{\hat{v}(s)}{\hat{i}(s)} \Big|_{\hat{v}_g=0} = \frac{G_{vd}(s)}{G_{id}(s)} = \frac{-sL + RD'^2 - R_i}{RRCD's + 2D'}. \quad (45)$$

The transfer function $G_{vi}(s)$ was calculated using circuit component values of Table 1. The gain and phase diagram of the transfer function $G_{vi}(s)$ is depicted in Figure 26.

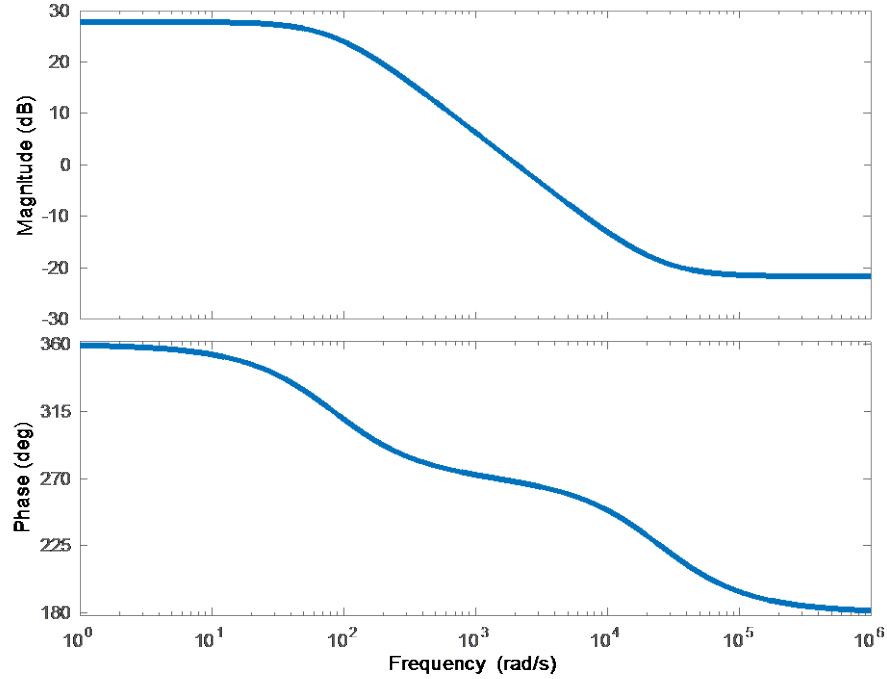


Figure 26. Gain and Phase Open-Loop Boost Converter Transfer Function $G_{vi}(s)$

The input of the inner PI controller is the reference inductor current and the output is the gate signal for duty cycle.

$$i_{com} = \left[\frac{k_{pl}s + k_{il}}{s} \right] (I_{ref} - I) \quad (46)$$

To set up this controller, transfer function $G_{id}(s)$ is to be used. The general form of a PI controller is given from

$$G_{pi}(j\omega) = K_p \left(1 + \frac{1}{j\omega T_i} \right) \quad (47)$$

To obtain the desired phase margin at the crossover frequency, the gains of the controller are given by equations (48) and (49) [10]

$$K_p = \frac{|\cos(\varphi_c - M_c)|}{|G(j\omega_{gc})|} \quad (48)$$

$$T_i = \frac{1}{\omega_{gc} \tan(\varphi_c - M)}, \quad (49)$$

where M = desired phase margin at ω_{gc} , φ_c = the phase of the transfer function at ω_{gc} (using $G_{vi}(j\omega_{gc})$ for the outer loop, and $G_{vd}(j\omega_{gc})$ for the inner loop)

To regulate the PI controllers the desired phase margin is set to 60°. In order to have a fast transient response, a different crossover frequency for each controller must be selected. The outer control loop is regulated using the transfer function $G_{vi}(s)$, so the selected crossover frequency must be lower than the value the right half zero. The inner control loop is regulated using the transfer function $G_{id}(s)$. Since there is a left hand zero there is no limitation on the selection of the crossover frequency, so it should be much higher than the previous one.

Figure 27 shows an optimization algorithm for the regular voltage mode control compensator.

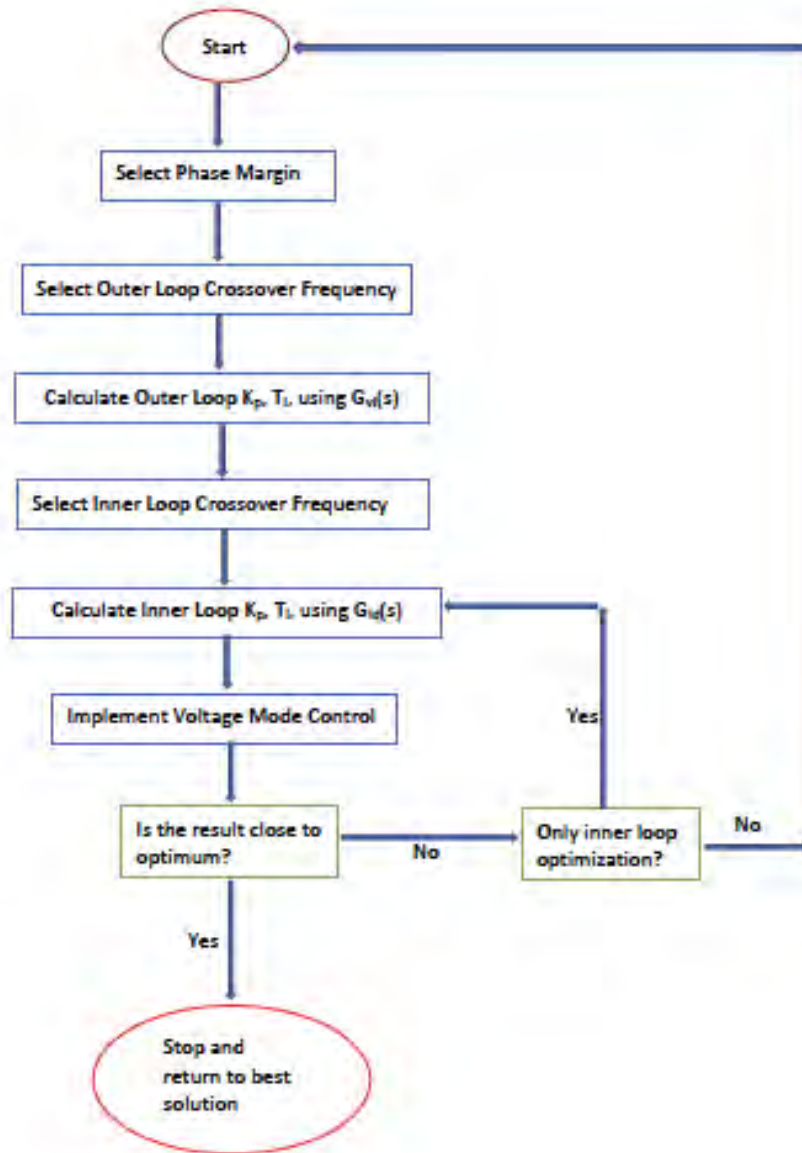


Figure 27. Regular Voltage Mode Control Compensator Optimization Algorithm

2. Modifications of the Regular Voltage Mode Control Methodology

Due to the fact that the voltage mode control is not operated sufficiently during step load changes that switch the operation from CCM to DCM and vice versa, two different

modifications were implemented. Regular voltage mode control has an unwanted limitation in the range of loads that the boost converter can be operated. The first modification (Mod 1 voltage mode control) uses the inductor current to detect if the boost converter operates in DCM. The second modification (Mod 2 voltage mode control) detects the raise in the output voltage to detect if the load has been reduced.

The first modification is depicted in Figure 28. The inductor current was used to detect if the boost converter switches to DCM. This was implemented by adding a comparator to detect when the inductor current went to zero. Then for duration where the inductor was detected to be zero, a different much faster set of inner and outer PI compensators was selected, with parameters that correspond to a DCM load value.

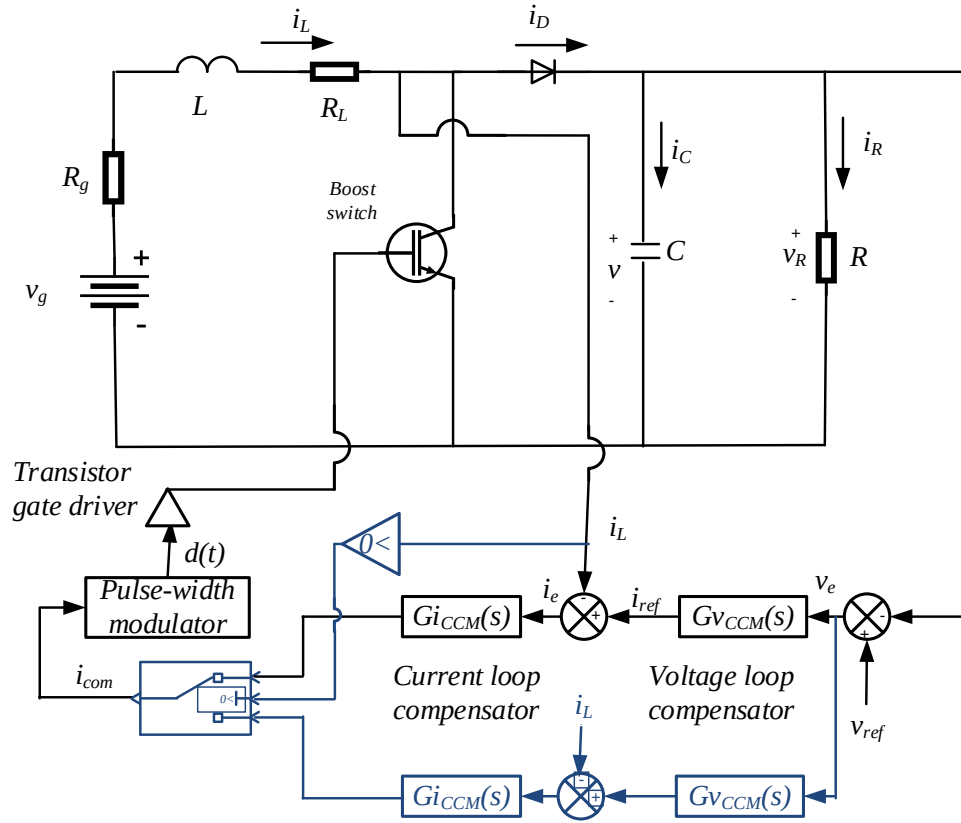


Figure 28. Closed-Loop Output Voltage Control Regulation of a Boost Converter Using the Mod 1 Voltage Mode Control Method

The second modification is depicted in Figure 29. The output voltage was used to detect the transient due to the load reduction of the boost converter. When a step load change to a lower value occurs during the transient response, the output voltage is increased. The goal of this modification is to control the peak voltage by using a faster compensator, corresponding to a different set of PI controllers. At first, a comparator was added to detect when the transient response occurs; this was implemented by comparing the output voltage with a predetermined value. The predetermined value was a value slightly higher than the desired output. Then for duration where the output voltage exceeds that value, a different set of compensators was selected, with parameters that correspond to a DCM load value.

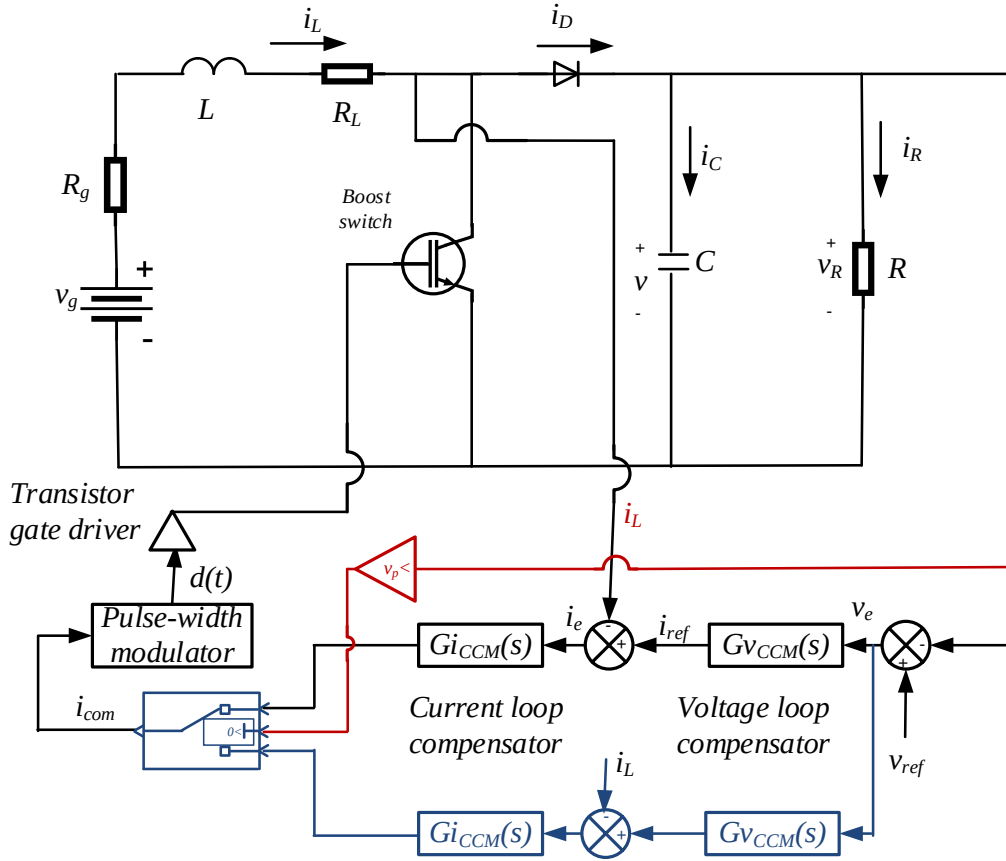


Figure 29. Closed-Loop Output Voltage Control Regulation of a Boost Converter Using the Mod 2 Voltage Mode Control Method

3. Voltage Mode Control Compensator Simulink Model

The voltage mode controller was implemented in Simulink. The design was a straightforward progress. The regular voltage mode control compensator model is depicted in Figure 30.

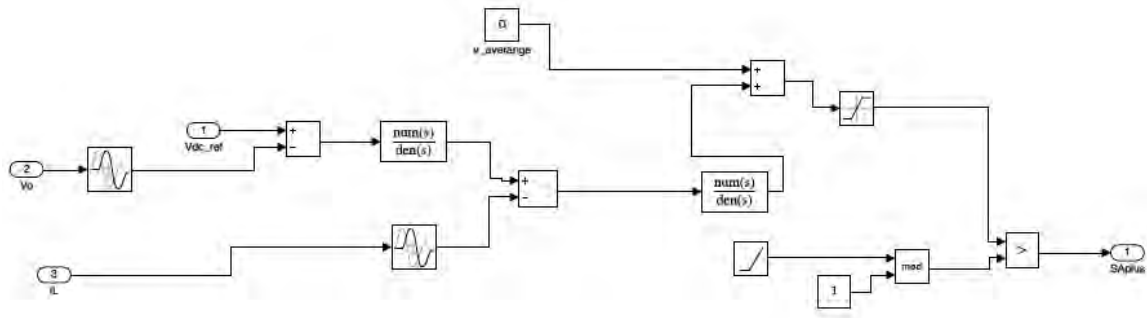


Figure 30. Regular Voltage Mode Control Compensator Simulink Model

The two loops of the regular voltage mode compensator are represented by the two PI controllers. Those two PI controllers are in series, and the output of the later one is the duty cycle.

The regular voltage mode compensator was modified to get the corresponding Mod 1 voltage mode control compensator, as depicted in Figure 31.

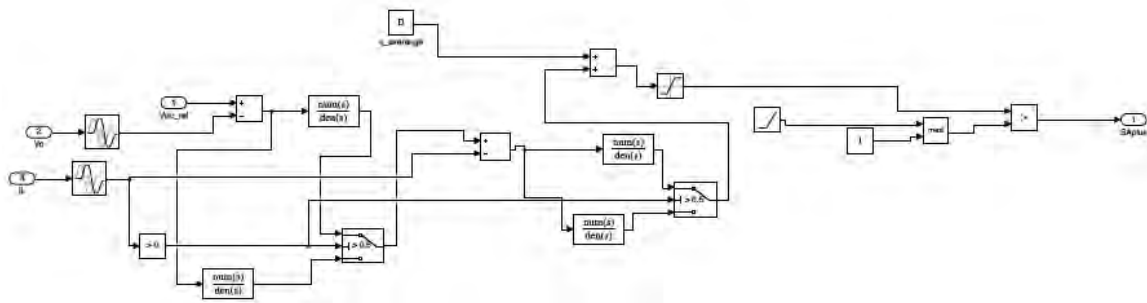


Figure 31. Mod1 Voltage Mode Control Compensator Simulink Model

A comparator was used to compare the inductor current with zero, the output of the comparator was the control signal for a switch, and this switch selected between two

different sets of controllers. The first set of controllers was set up with parameter values that corresponded to CCM, and the second set of controllers was set up with parameter values that corresponded to DCM.

The Mod 2 voltage mode control compensator Simulink model is depicted in Figure 32.

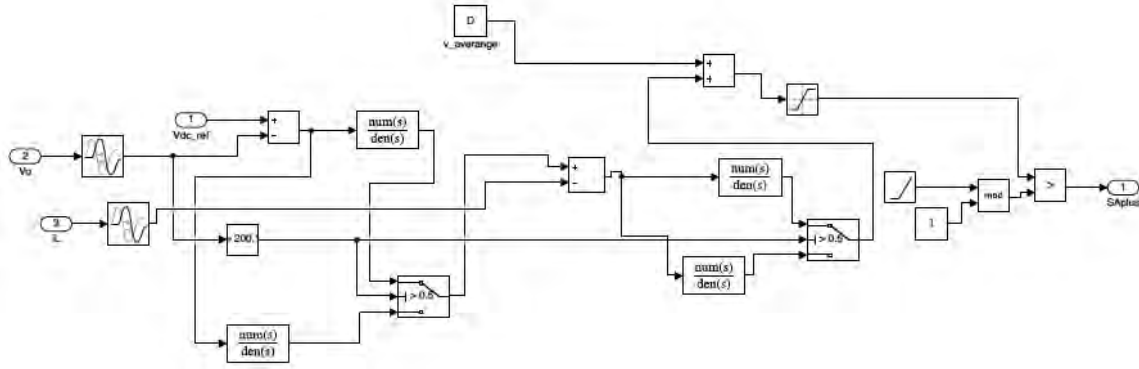


Figure 32. Mod 2 Voltage Mode Control Compensator Simulink Model

Similarly, with the first modification a comparator was used to create the control signal for a switch. The switch selected between two different sets of controllers, but this time the comparator took as an input the output voltage and compared it with a predetermined value to detect if a transient due to load reduction was occurring.

4. Voltage Mode Control Methodologies Comparison

For each modification a simulation was executed to validate its effectiveness. The effectiveness of each version of the voltage mode control was evaluated for three different step load changes. The parameters that were used in Simulink are shown in Table 3. The parameters used to set up the controllers are shown in Table 5. The CCM gains refer to the gains that were used by the regular voltage mode control compensator. The two versions of the modified voltage mode control used the CCM parameters during the normal operation and when necessary switched to a compensator with parameters equal to the DCM ones.

Table 5. Voltage Mode Control Compensator Parameters

Gains	Voltage Mode Control			
	CCM		DCM	
	Inner loop	Outer loop	Inner loop	Outer loop
K_p	0.095	5.53	0.96	11.71
K_i	1351.95	10089.22	131740.74	6850.02

The first step load change was executed for changing the load resistance from $50\ \Omega$ to $185\ \Omega$. All three versions of the voltage mode controller had similar performance; the transient response of each compensator is depicted in Figure 33. Actually, the regular voltage mode control performed slightly better than the modified versions.

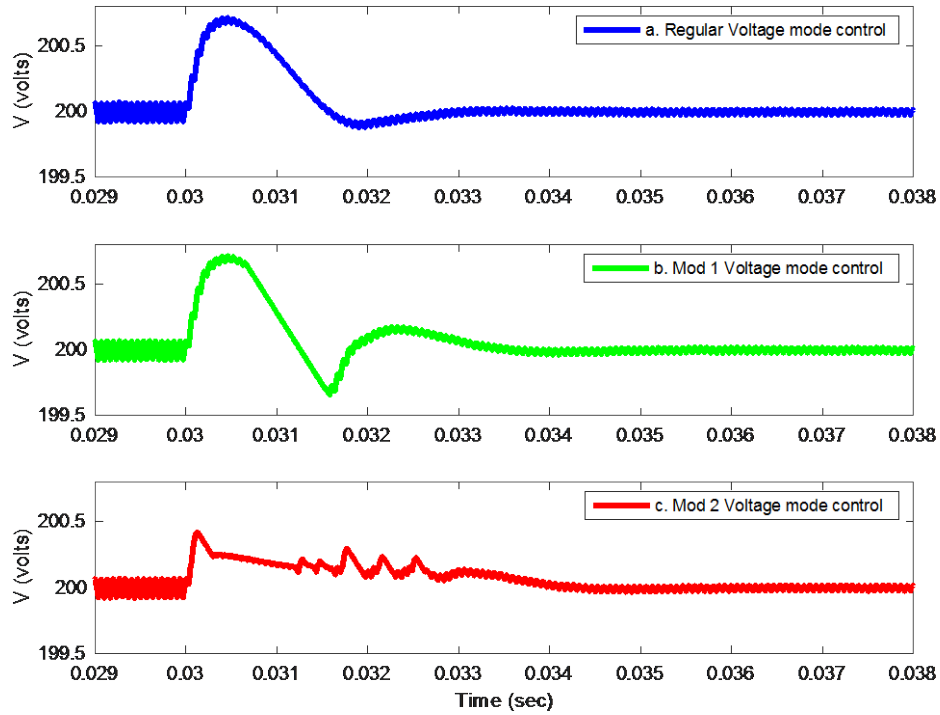


Figure 33. Closed-Loop Step Response for Load Change from $50\ \Omega$ to $185\ \Omega$: (a) Regular Voltage Mode Control, (b) Mod 1 Voltage Mode Control, (c) Mod 2 Voltage Mode Control

The effectiveness of the modified versions of the voltage mode compensators was revealed at the second step load change that was evaluated. After changing the load resistor back to $50\ \Omega$, at steady state the second step load change occurred. This step load change was executed for changing the load resistance from $50\ \Omega$ to $500\ \Omega$. For this step load change the regular control methodology could not follow the command and then became unstable, but the two modified versions of compensators were able to follow the command and actually both performed with a voltage ripple less than 0.5%. The transient response of the two modified compensators is depicted in Figure 34. Comparing the two modified versions we observe that the second one resulted in the better performance.

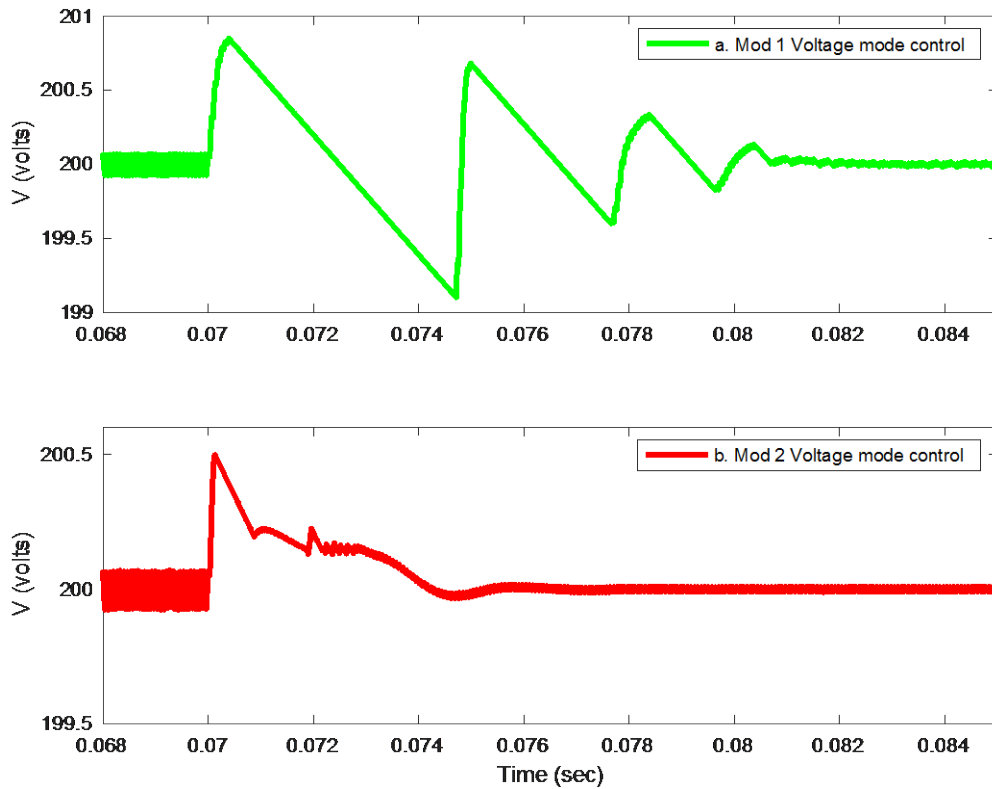


Figure 34. Closed-Loop Step Response for Load Change from $50\ \Omega$ to $500\ \Omega$: (a) Mod 1 Voltage Mode Control, (b) Mod 2 Voltage Mode Control

The third step load change was for changing the load resistance from $500\ \Omega$ to $50\ \Omega$. The regular voltage mode compensator became unstable, but the two versions of the modified controllers were able to maintain a constant voltage output. Both modified versions performed well, and both modifications resulted in an output voltage ripple less than 1%. Actually, the second one had the better performance. The transient response of the two modified compensators is depicted in Figure 35.

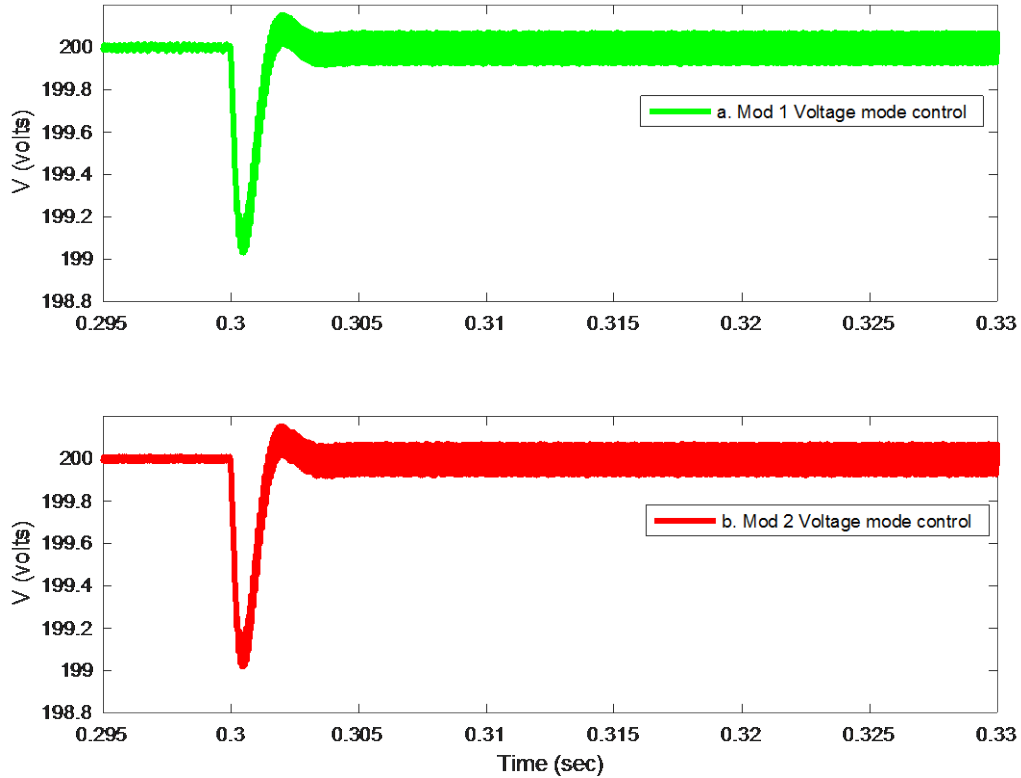


Figure 35. Closed-Loop Step Response for Load Change from $500\ \Omega$ to $50\ \Omega$: (a) Mod 1 Voltage Mode Control, (b) Mod 2 Voltage Mode Control

Taking all the above simulations into consideration, it is clear that the two modifications significantly improved the effectiveness of the regular voltage mode compensator. With the modifications, the boost converter can operate in a load range that was beyond the capabilities of the regular controller. In particular, the second modification of the voltage compensator resulted in the better overall performance, in all the cases, since

both modifications were able to maintain the voltage output with voltage ripple less than 1%, but the second one was faster and had less ripple.

C. CURRENT MODE CONTROL METHODOLOGY

A description of the current mode control methodology as well as a simple algorithm to set the required parameters are introduced.

1. Current Mode Control Methodology

A current mode control compensator consists of two feedback loops. One is an outer loop with a PI regulator that takes as input the error between the reference voltage and the measured output voltage. The output is the reference inductor current. The second is an inner loop that takes as inputs the reference inductor current and the measured inductor current and compares them to produce the gate pulse for the IGBT. This method is unstable for a duty cycle greater than 50% due to subharmonic oscillation [11]. Therefore, a slope compensation is required to maintain stability of the system and to provide a feed-forward property [2], [11], [12]. There is a special case in which the slope compensation is not required; this happens if the gain of the open-loop transfer function $G_{vd}(s)$ at half switching frequency is less than -10dB [12]. The selected slope is equal to one half of the inductor current when the IGBT operates in position 2 [2]. The estimated perturbations of the inductor current have been added to the reference inductor current [13], [14]

$$\hat{i}_{FF} = \hat{v} \frac{G_{Id}}{G_{vd}} \bigg|_{\omega} = \frac{2D'}{D'^2 R - R_i} \hat{v} \quad (50)$$

A closed-loop block diagram for the regulation of the output voltage of the DC-DC boost converter using a current mode control compensator is shown in Figure 36.

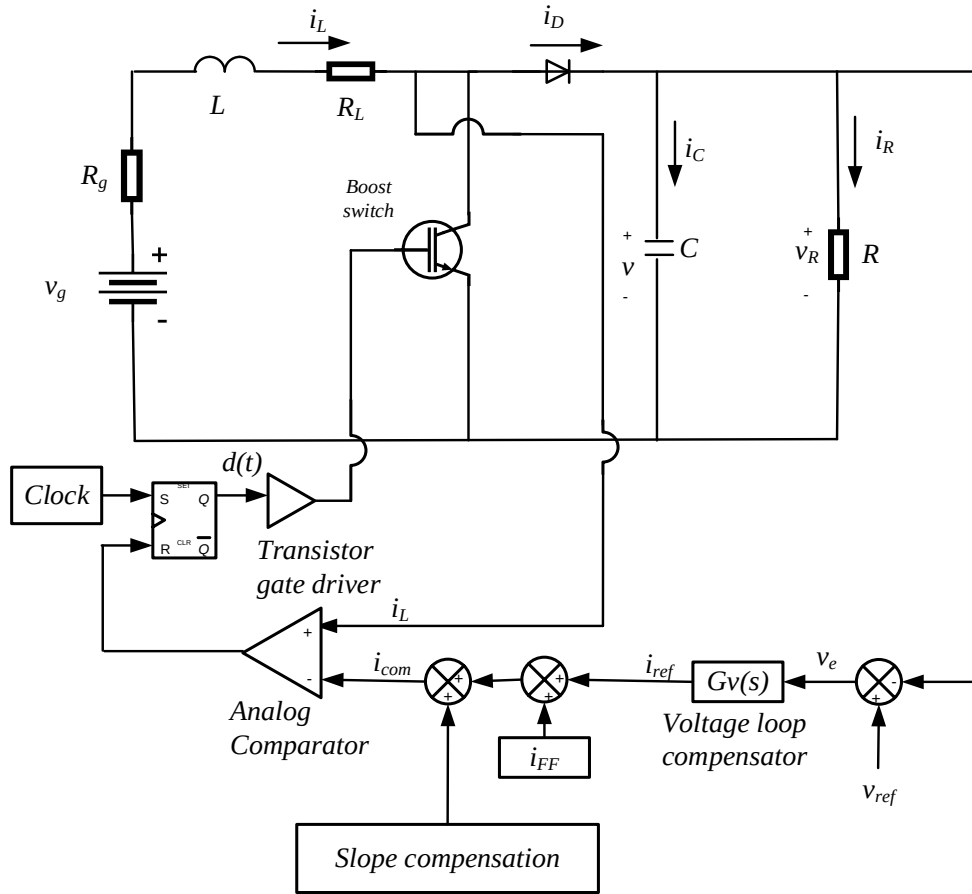


Figure 36. Closed-Loop Output Voltage Control Regulation of a Boost Converter Using Current Mode Control Method

The PI regulator has been set up using the transfer function $G_{vi}(s)$. Similarly, with the regulation of the voltage control methodology (48) and (49) are to be used. A desired phase margin of 60° is selected. The crossover frequency is selected as high as possible with the limitation to be less than the right half zero of the transfer function $G_{vi}(s)$.

Figure 37 shows an optimization algorithm for the current mode control compensator.

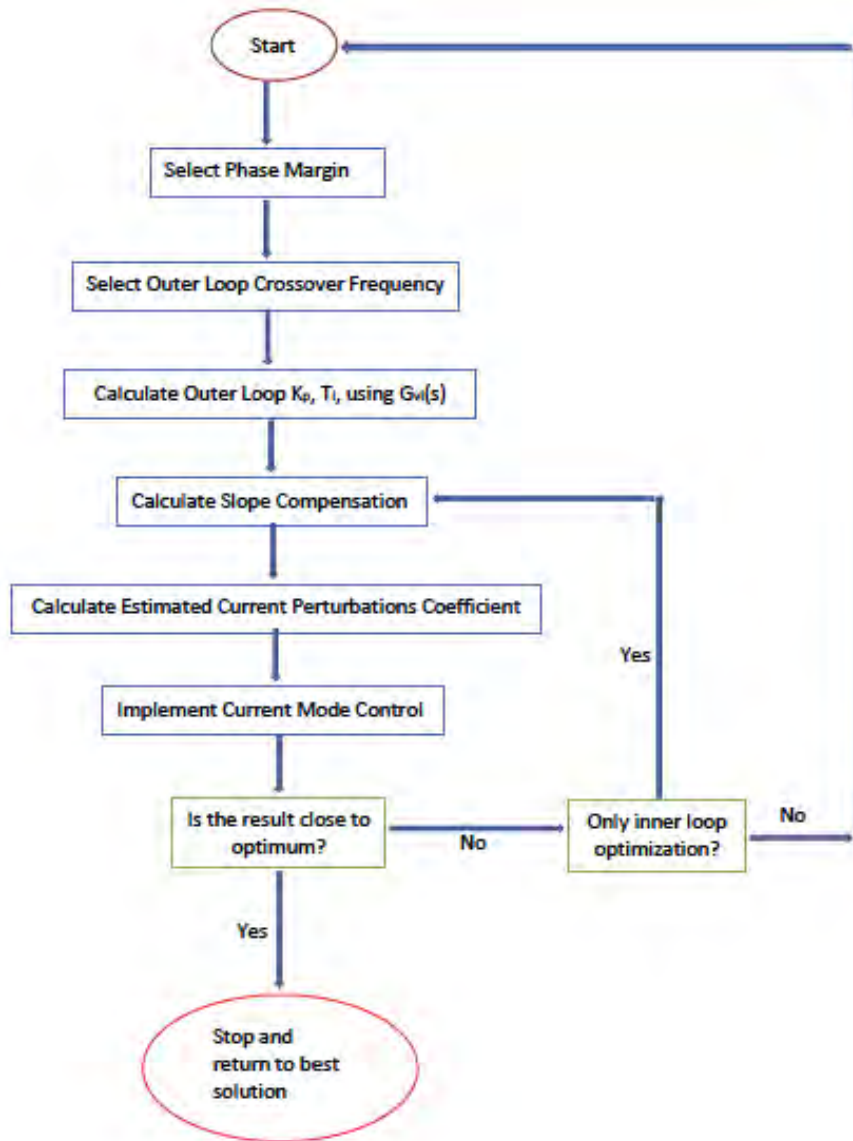


Figure 37. Current Mode Control Compensator Optimization Algorithm

The current mode control method performed well, without any kind of modification. Similar modifications with the other control methodologies were implemented, but the performance of the current mode control was not improved.

2. Current Mode Control Compensator Simulink Model

The current mode controller was implemented in Simulink. The design was a straightforward progress. The current mode control compensator model is depicted in Figure 38.

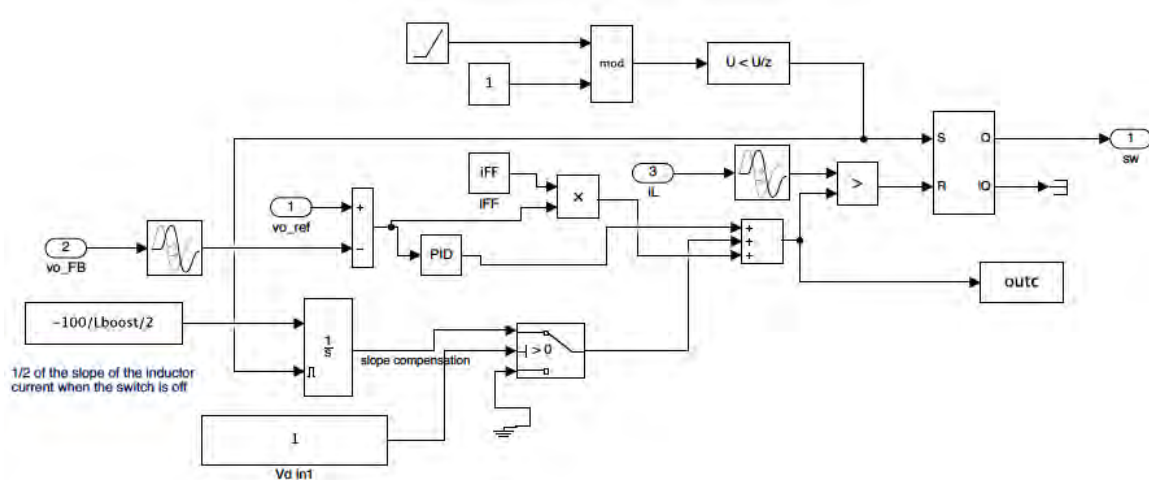


Figure 38. Current Mode Control Compensator Simulink Model

The output of the outer loop was equal to the output of the PI controller with the addition of the slope compensator and the estimated current perturbations. A comparator was used to compare the inductor current with the output of the outer loop, and the output of the comparator was fed as the reset signal to a S-R flip-flop. The set signal was a sawtooth. The output of the flip-flop was the gate drive control signal for the switch.

IV. LABORATORY IMPLEMENTATION AND EXPERIMENTAL MEASUREMENTS

The experimental validation of the transfer function $G_{vd}(s)$ was conducted in two ways. The first way was an experimental hardware validation, by building a boost converter prototype in the laboratory. The second way was an experimental validation using Simulink. In both methods, perturbations were injected in the duty cycle. The resulting perturbations of the voltage output were measured in order to calculate the gain and the phase of the open-loop transfer function $G_{vd}(s)$.

A. HARDWARE LABORATORY EXPERIMENT

This section provides a description of the laboratory setup, the equipment used, and the procedure followed.

1. Hardware Laboratory Setup and the Equipment Used

The hardware validation was conducted in the laboratory using the Semikron hardware box (Semistack-IGBT [23]). The Semistack-IGBT box is a pre-qualified power assembly ready for integration. This box contains diodes, IGBTs, inverter, A/D converter and digital interface, capacitors, IGBT grade drive connection, and motor. The Semistack-IGBT is controlled with Xilinx XC4VLX25 FPGA. The input DC power was provided by BK Precision XLN3640 programmable DC power supply. A sine-triangle pulse-width modulation (PWM) was implemented using a LM311. Clamp-on probes were used to take source voltage and current measurements, which were captured using an oscilloscope. The schematic of the circuit is shown in Figure 39. All discrete components values are shown in Table 6.

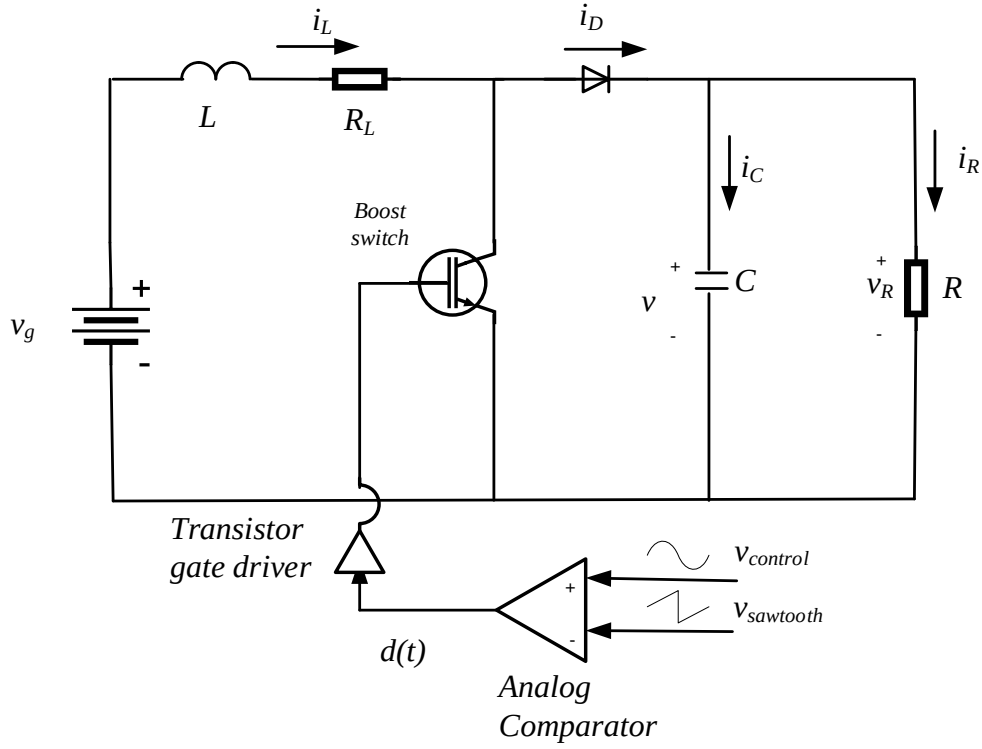


Figure 39. Experimental Validation Circuit Schematic

Table 6. Discrete Components Values for the Hardware Experiment

Device	Value
V_g	10 V
V_D	0.8 V
L	1114 μ H
C	2094 μ F
R_L	0.355 Ω
R_{on}	0.5 Ω
$R_i = R_L + R_{on}$	0.855 Ω
R	101.45 Ω

The boost converter was built in the laboratory using an IGBT, a diode, the capacitors, and the IGBT gate drive connection of the Semikron box (Semistack IGBT). The components of the experimental setup are illustrated in Figure 40.

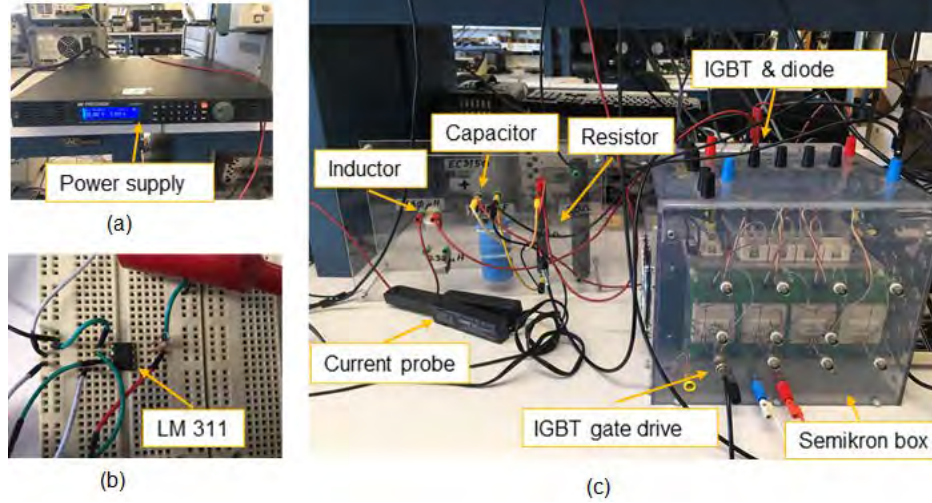


Figure 40. Laboratory Setup with Major Components Labeled (a) DC Power Source, (b) PWM, (c) Boost Converter

The input voltage v_g was set to 10V. The desired output voltage was selected by setting the appropriate duty cycle for the IGBT switch. The inductor value is $1114\mu\text{H}$ and was chosen such that the converter operates in CCM. The filter capacitor value is $994\mu\text{F}$, and the Semikron box capacitor is $1100\mu\text{F}$. Those capacitors are connected in parallel, which results in a total capacitance of $2094\mu\text{F}$. The choice of the capacitance was made to maintain the output voltage ripple less than 1%.

Perturbation in the duty cycle has been injected by using a comparator to create a PWM, as depicted in Figure 40 (b). The PWM was used as the switch control signal for the IGBT. At the beginning the circuit of Figure 41 was implemented. The comparator was used to create a square pulse [2]; this was accomplished by comparing a DC voltage at one input V_{control} with a sawtooth at the other input v_{sawtooth} and the output was the following [24]

$$v_p = 15V, \text{ for } V_{\text{control}} > v_{\text{sawtooth}}$$

$$v_p = 0, \text{ for } V_{\text{control}} < v_{\text{sawtooth}},$$

where v_p is the output of the comparator.

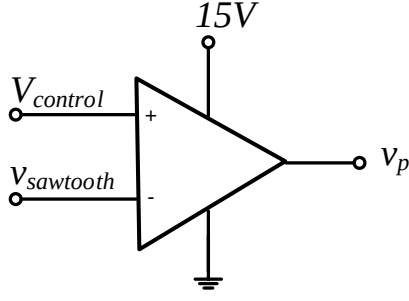


Figure 41. Voltage Comparator Circuit

A PWM can be implemented when the voltage comparator is used to compare a slow varying sinusoidal $\hat{v}_{control}$ with a high-frequency sawtooth [3]. The output of the comparator is a square wave; the frequency is equal to the sawtooth frequency; the duty cycle is controlled by $\hat{v}_{control}$ [3]. The corresponding duty cycle is the following [3]

$$d(t) = 100 \frac{\hat{v}_{control}}{v_{sawtooth}} = D_m \cdot \sin(\omega_m t) = \hat{d}(t) \quad (51)$$

In order to inject perturbation into the duty cycle, the control signal $v_{control}$ consists of a small sinusoid signal $\hat{v}_{control}$ with DC offset $V_{control}$. The DC offset is used to adjust the duty cycle, and the small sinusoid signal $\hat{v}_{control}$ is used to inject the perturbation in the duty cycle, as depicted in Figure 42.

$$d(t) = 100 \frac{v_{control}}{v_{sawtooth}} = 100 \frac{V_{control}}{v_{sawtooth}} + 100 \frac{\hat{v}_{control}}{v_{sawtooth}} = D + D_m \cdot \sin(\omega_m t) = D + \hat{d}(t) \quad (52)$$

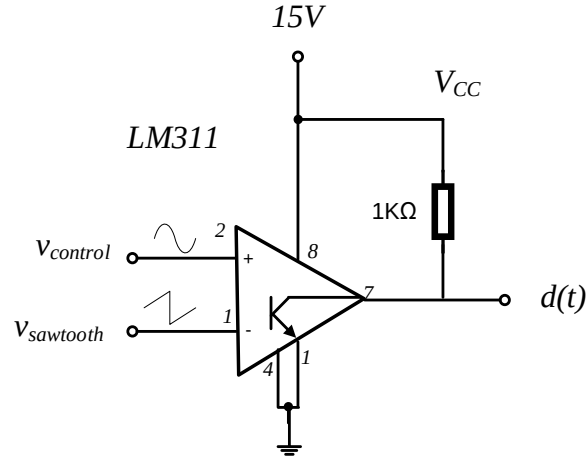


Figure 42. Injecting Perturbations on Duty Cycle via Modulating a High-Frequency Triangular Wave with a Low-Frequency Control Signal

Two Agilent 3320A function generators were used, one to generate a sawtooth and one to generate the control signal. The sawtooth waveform had the following parameter values: frequency 15kHz, low voltage 0 V, and peak-to-peak amplitude 10 V. The control signal was a sinusoidal wave with a DC offset; the DC offset was set to the required value for the desired duty cycle; the amplitude of the AC signal was varied from 20mV to 2.5 V; the frequency was varied from 1 Hz to 550 Hz.

A Keysight infiniiVision DSOX3014T digital oscilloscope was used to measure the magnitude and the phase of the AC portion of the control signal and the corresponding converter output at the injected frequency. The phase shift between those two signals was measured.

2. Hardware Laboratory Procedure and Testing

Implementation of the experiment was a straightforward process. Upon completing the lab setup the DC part of the control signal $V_{control}$ was set to get the desired output voltage, and the AC part of the control signal $\hat{v}_{control}$ was adjusted in two ways. The frequency of the control signal $\hat{v}_{control}$ was adjusted to perform a frequency sweep. Furthermore, its amplitude was selected to observe the perturbation of the output voltage. At low frequencies a small amplitude value was selected, but on higher frequencies a higher

amplitude value was required. Finally, the amplitude and phase values of perturbations for both the duty cycle and the output voltage were recorded to calculate the gain and the phase difference. The open-loop gain for each frequency was acquired by dividing the measured converter output by the input control signal. Due to the fact that for the PWM a sawtooth with amplitude 10V peak to peak was used, for the gain calculation the control signal was scaled by a factor of 10. This procedure for each was repeated for three different duty cycles. For each duty cycle the experiment was repeated three times to calculate the average gain and phase margin for the open-loop transfer function $G_{vd}(s)$.

At first, the offset of the control signal $V_{control}$ was set to get 20V output, which corresponds to a duty cycle of 55.9%. The duty cycle waveform with the injected perturbations is shown in Figure 43. Oscilloscope channel one (yellow line) was the control signal. The control signal consisted of a DC voltage added with a small AC signal at 5Hz. Since the sawtooth frequency was 15kHz, the selected time scale was equal to 50 μ s/div. Therefore, the control looks like a straight line, but actually there was an injected sinusoidal wave. Channel two (green line) was the sawtooth. Channel three (magenta line) was the created PWM. There was a small variation in the PWM pulse width barely visible in the variation of the horizontal segments of the magenta line, depending on the injected perturbations.

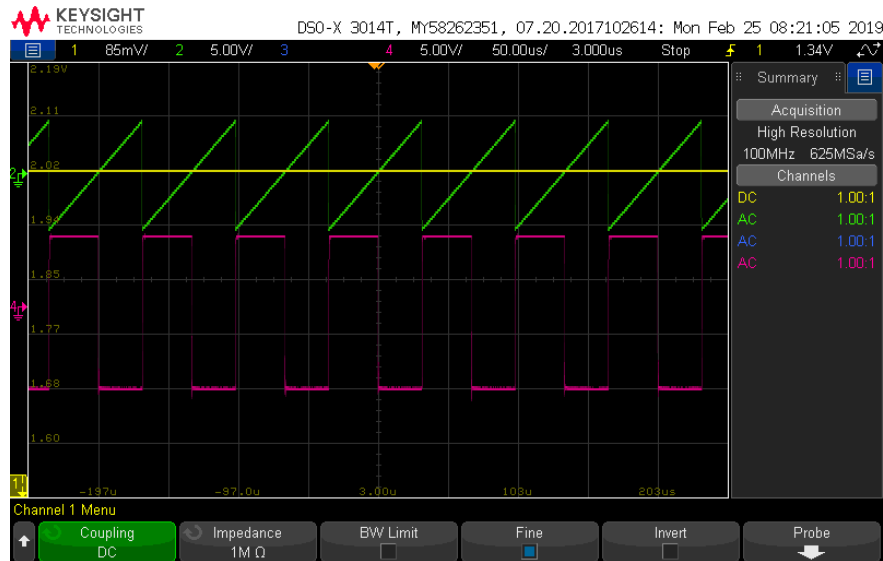
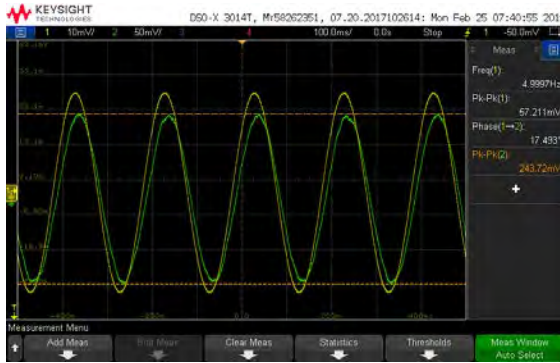
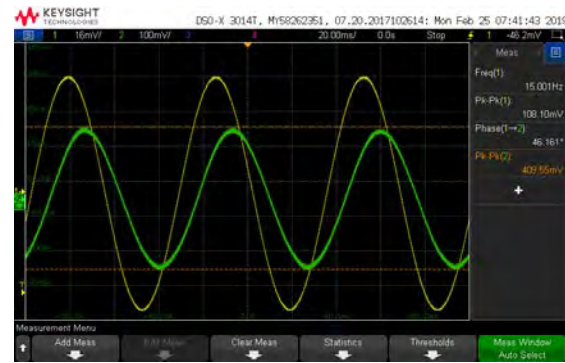


Figure 43. PWM Waveform for a Duty Cycle of 55.9% with the Injected Perturbations at 5Hz (CH1: 85mV/div, CH2: 5V/div, CH4: 5V/div, time: 50 μ s/div)

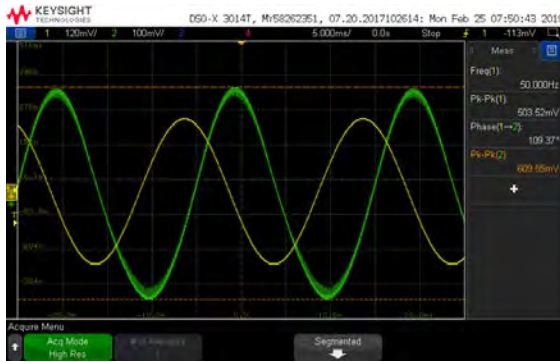
The injected perturbations of the duty cycle and the corresponding perturbations of output voltage are shown in Figure 44.



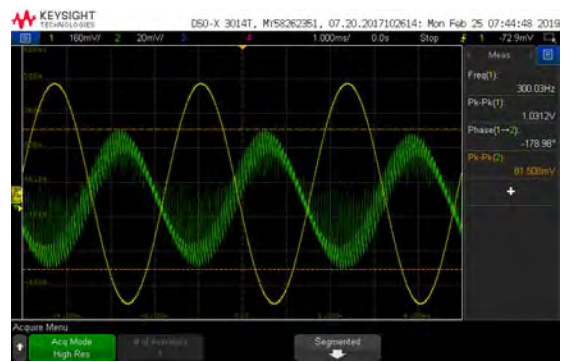
(a)



(b)



(c)



(d)

Figure 44. The Injected Perturbations for a Duty Cycle of 55.9% and the Corresponding Perturbations of the Output Voltage: (a) at 5 Hz (CH1: 10mV/div, CH2: 50mV/div, time: 10ms/div), (b) at 15 Hz (CH1: 16mV/div, CH2: 100mV/div, time: 20ms/div), (c) at 50 Hz (CH1: 120mV/div, CH2: 100mV/div, time: 5ms/div), (d) at 250 Hz (CH1: 160mV/div, CH2: 20mV/div, time: 1ms/div)

When the frequency of the injected perturbations of the duty cycle is relatively high, the output signal is weak, and the phase difference calculation is less accurate (Figure 44[d]).

During the experiment the current of the inductor was monitored with a current probe connected to the oscilloscope to ensure that the boost converter operates in CCM. The current waveforms are depicted in Figure 45.

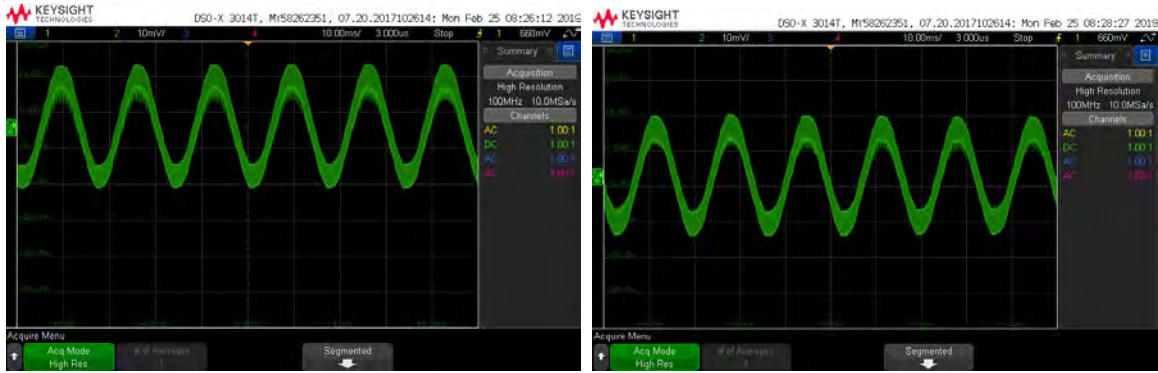


Figure 45. Inductor Current at a Duty Cycle of 55.9% for Injected Perturbations in the Duty Cycle with Frequency: (a) 5 Hz (CH1: 10mV/div, time: 10ms/div), (b) 50 Hz (CH1: 10mV/div, time: 10ms/div)

The experiment was repeated and the offset of the control signal $V_{control}$ was set to get 17.7V output, which corresponds to a duty cycle of 49.7%. The duty cycle waveform with the injected perturbations is shown in Figure 46. The oscilloscope channel configurations and visible results were similar to Figure 43.

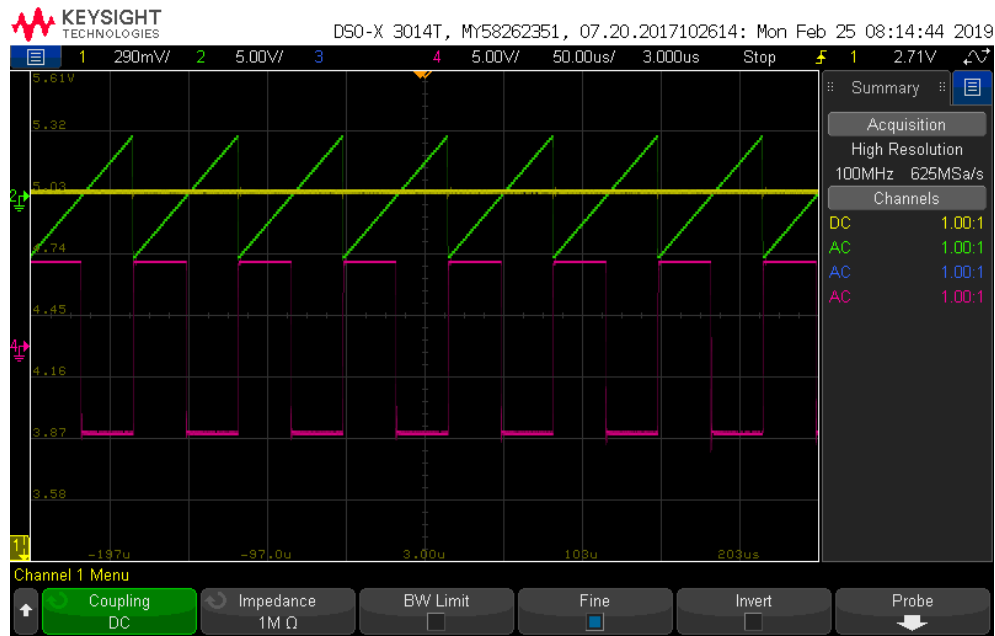
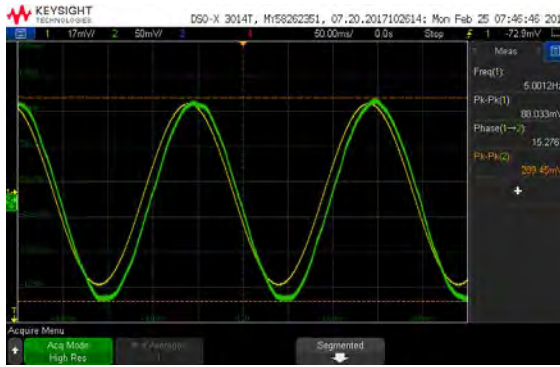
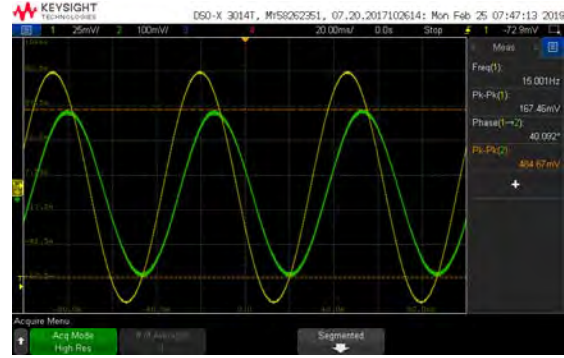


Figure 46. PWM Waveform for a Duty Cycle of 49.7% with the Injected Perturbations at 5Hz (CH1: 290mV/div, CH2: 5V/div, CH4: 5V/div, time: 50μs/div)

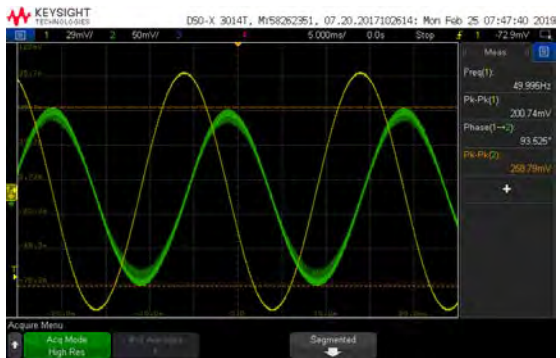
The injected perturbations of the duty cycle and the corresponding perturbations of the output voltage are shown in Figure 47.



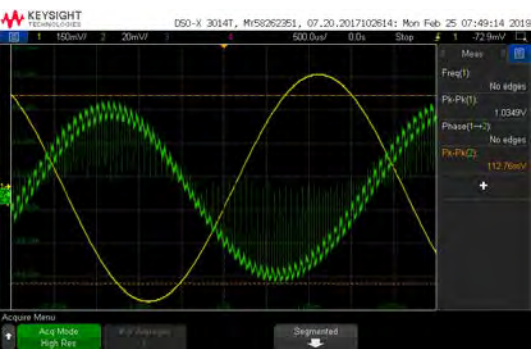
(a)



(b)



(c)



(d)

Figure 47. The Injected Perturbations for a Duty Cycle of 49.7% and the Corresponding Perturbations of the Output Voltage: (a) at 5 Hz (CH1: 17mV/div, CH2: 50mV/div, time: 50ms/div), (b) at 15 Hz (CH1: 25mV/div, CH2: 100mV/div, time: 20ms/div), (c) at 50 Hz (CH1: 29mV/div, CH2: 50mV/div, time: 5ms/div), (d) at 250 Hz (CH1: 150mV/div, CH2: 20mV/div, time: 50 μ s/div)

Finally, the experiment was repeated and the offset of the control signal $V_{control}$ was set to get 22.9V output, which corresponds to a duty cycle of 62.1%. The duty cycle waveform with the injected perturbations is shown in Figure 48. The oscilloscope channel configurations and visible results were similar to Figure 43.

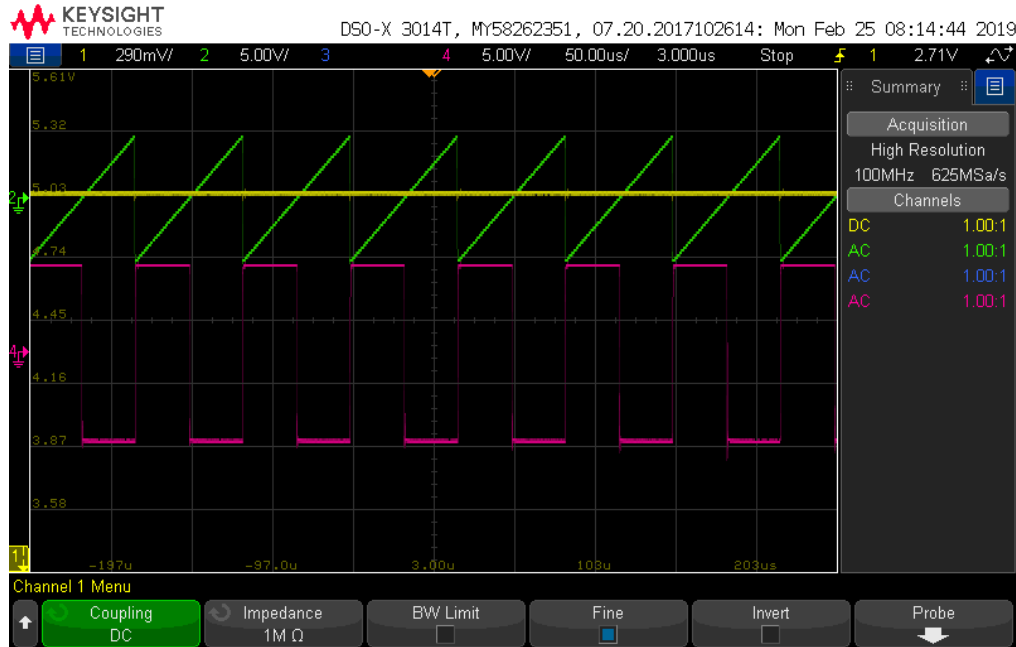
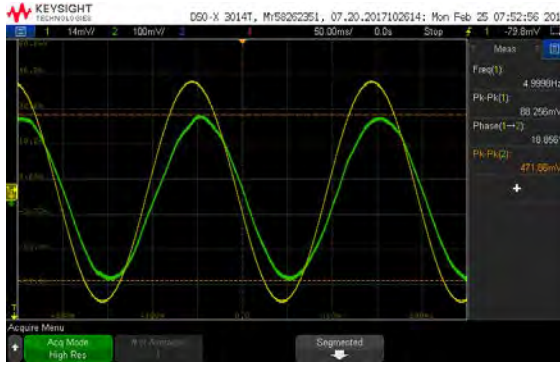
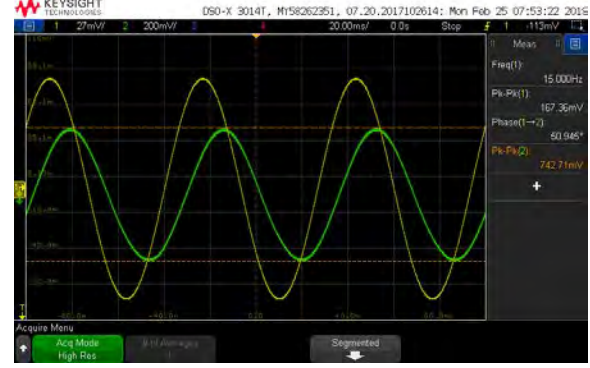


Figure 48. PWM Waveform for a Duty Cycle of 62.1% with the Injected Perturbations at 5Hz (CH1: 290mV/div, CH2: 5V/div, CH4: 5V/div, time: 50 μ s/div)

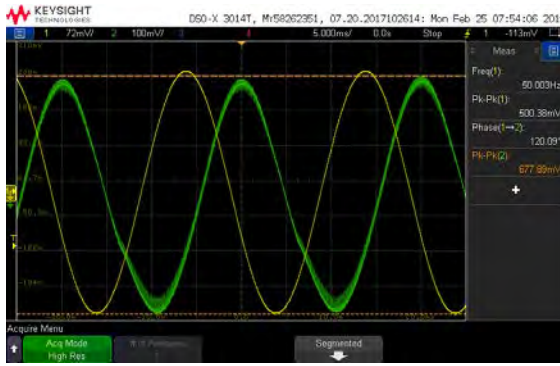
The injected perturbations of the duty cycle and the corresponding perturbations of the output voltage are shown in Figure 49.



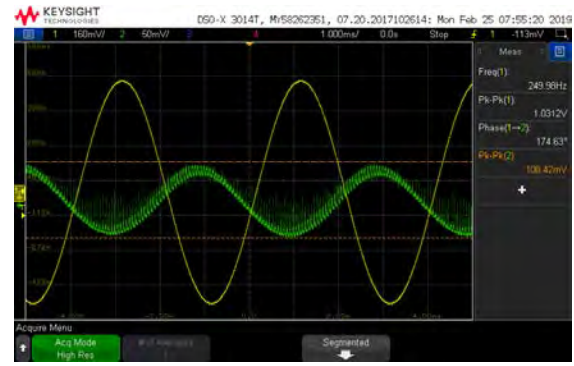
(a)



(b)



(c)



(d)

Figure 49. The Injected Perturbations for a Duty Cycle of 62.1% and the Corresponding Perturbations of the Output Voltage: (a) at 5 Hz (CH1: 14mV/div, CH2: 100mV/div, time: 50ms/div), (b) at 15 Hz (CH1: 27mV/div, CH2: 200mV/div, time: 20ms/div), (c) at 50 Hz (CH1: 72mV/div, CH2: 100mV/div, time: 5ms/div), (d) at 250 Hz (CH1: 160mV/div, CH2: 50mV/div, time: 1ms/div)

3. Simulink Verification Procedure and Testing

The software validation of the open-loop transfer function $G_{vd}(s)$ was conducted in Simulink. The configurations used were the same as those in Figure 39. All parameters were the same as the hardware validation except for the amplitude of the sawtooth. The amplitude of the sawtooth was selected to be 1 volt, so that it would not be required to scale the amplitude of the input perturbations for the calculation of the gain for the open-loop transfer function $G_{vd}(s)$. The amplitude and phase of the injected perturbations in duty cycle

and the corresponding perturbations in the output voltage were measured using a Fourier block. The schematic of the Simulink setup is shown in Figure 50.

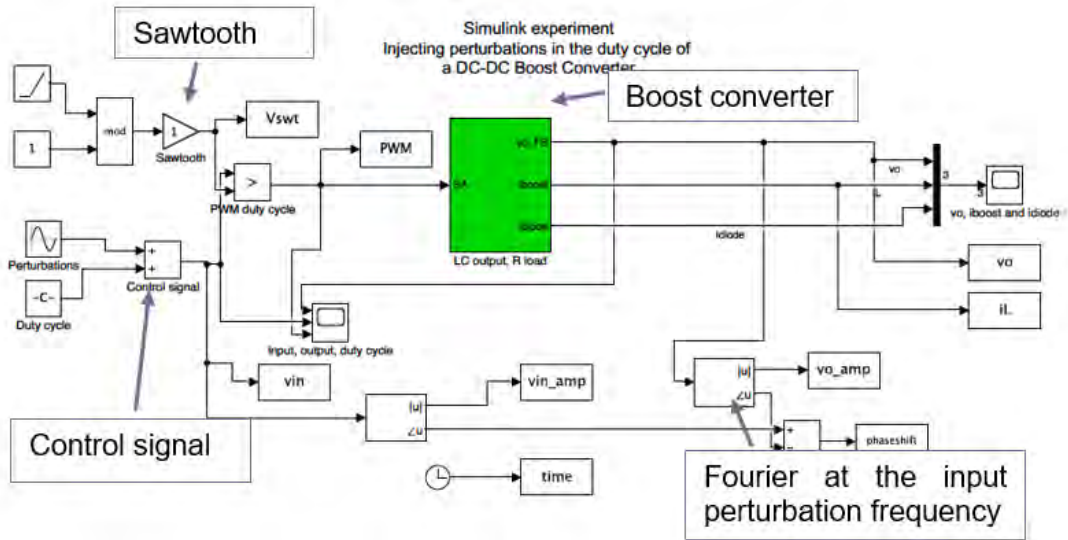


Figure 50. Simulink Setup with Major Components Labeled

The offset of the control signal $V_{control}$ was set to 20V output, which corresponds to a duty cycle of 55.9%. The duty cycle waveform with the injected perturbations is shown in Figure 51. In visible results produced by Simulink, perturbations are barely visible, as in Figure 43.

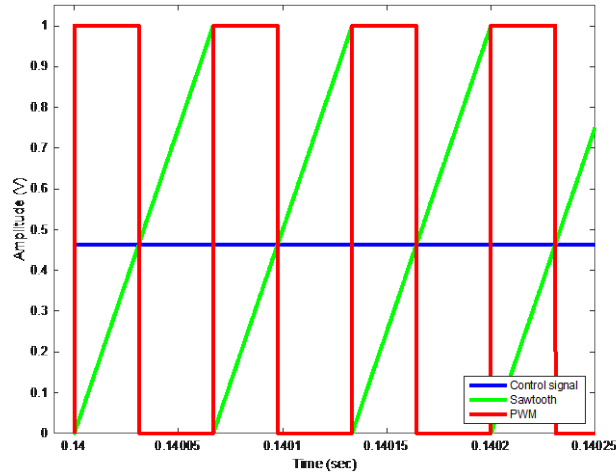


Figure 51. Simulink PWM Waveform for a Duty Cycle of 55.9% with the Injected Perturbations at 5Hz

The injected perturbations of the duty cycle and the corresponding perturbations of the output voltage are shown in Figure 52.

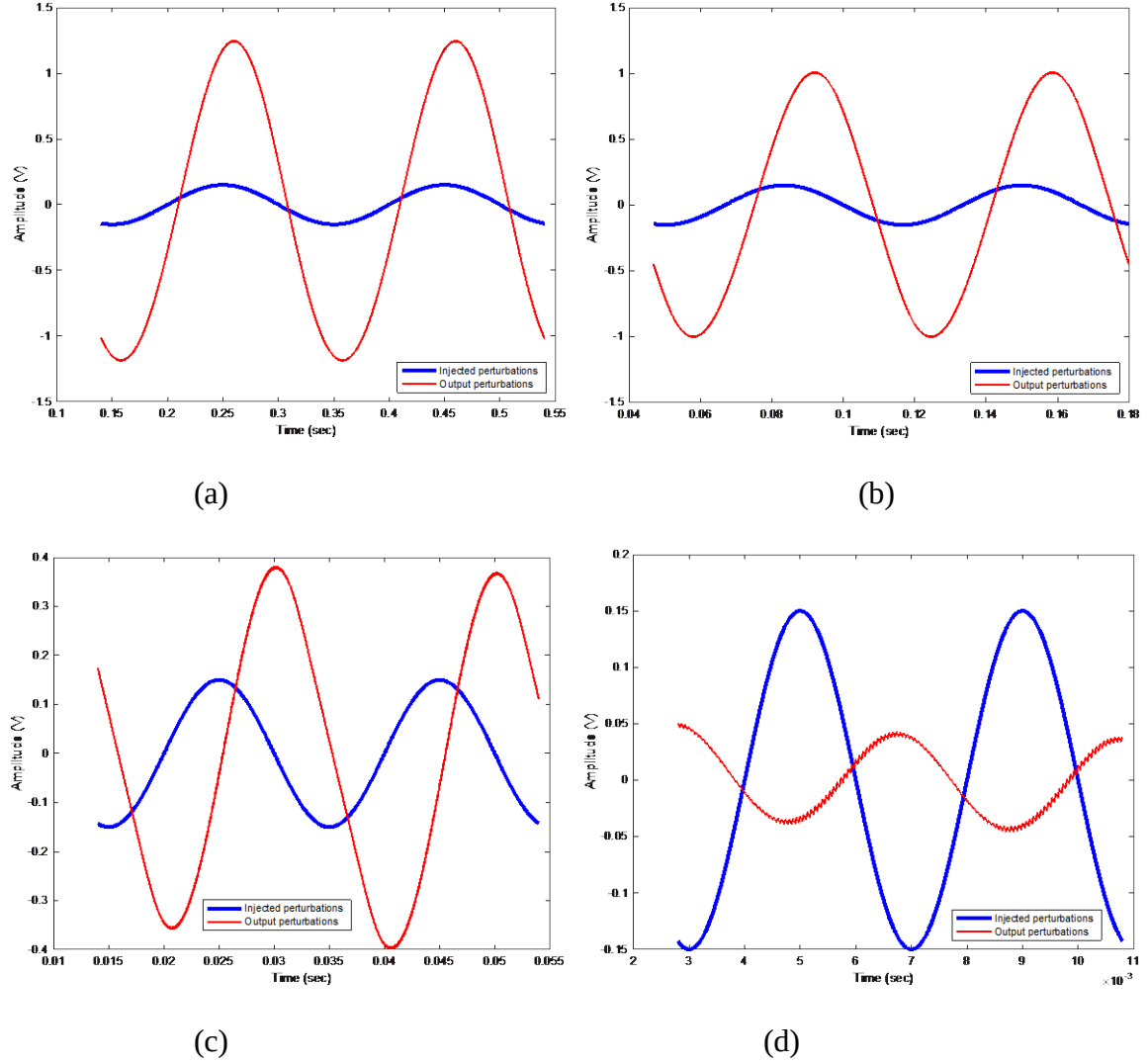


Figure 52. Simulink Model of the Injected Perturbations for a Duty Cycle of 55.9% and the Corresponding Perturbations of the Output Voltage: (a) at 5 Hz, (b) at 15 Hz, (c) at 50 Hz, (d) at 250 Hz

4. Hardware and Simulink Verification Results

Both hardware and software experiments prove that the theoretical open-loop transfer function $G_{vd}(s)$ as given in (37) is a satisfactory representation of the DC-DC boost converter. Figure 53, Figure 54, and Figure 54 show the comparison of the theoretical open-

loop transfer function $G_{vd}(s)$ with the measured values obtained by the hardware experiment for the three tested duty cycles. At the hardware validation experiment results there was a deviation at the phase difference for relatively high frequencies of the injected perturbations because the measured signal in those cases was weak and the phase difference measurement was less accurate compared to the lower-frequency results. Figure 56 shows the comparison of the theoretical open-loop transfer function $G_{vd}(s)$ with the simulated one. In particular, for the simulation experiment the results are almost identical to the theoretical expected values.

The results from using the mean gain and phase difference that were acquired from the hardware experiment for the case of a desired output of 20V and a duty cycle of 55.9% are depicted in Figure 53.

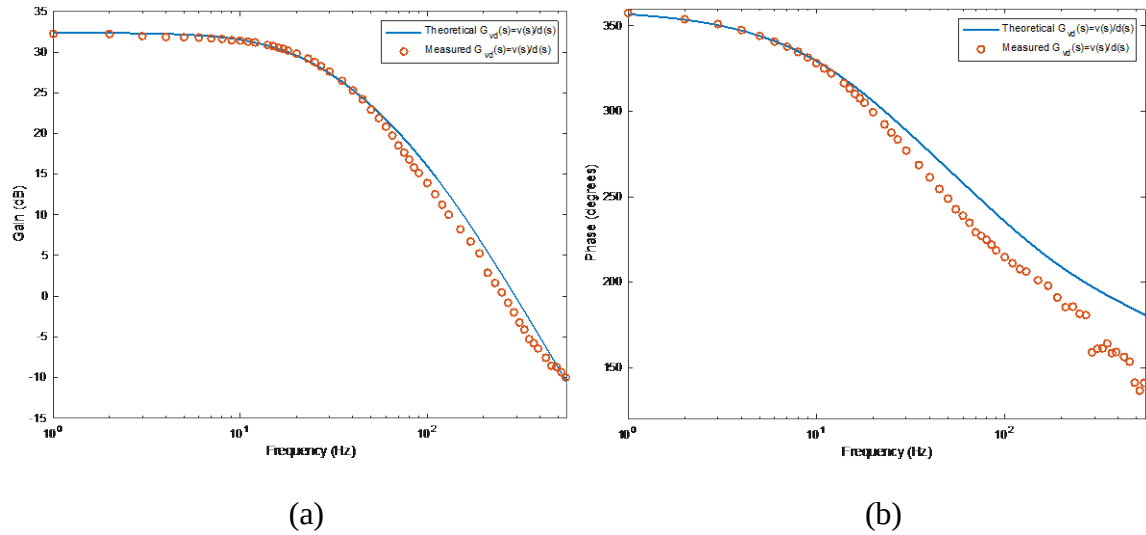


Figure 53. Comparison between Theoretical and Measured Open-Loop Transfer Function $G_{vd}(s)$ for a Duty Cycle of 55.9%: (a) Gain, (b) Phase

The results from using the mean gain and phase difference that were acquired from the hardware experiment for the case of a desired output of 17.7V and a duty cycle of 49.7% are depicted in Figure 54

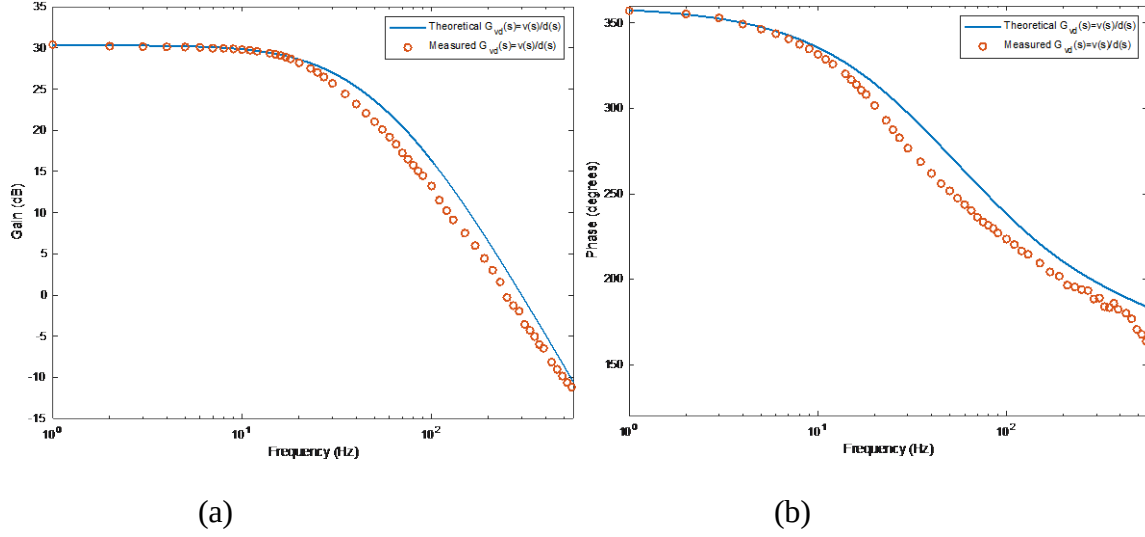


Figure 54. Comparison between Theoretical and Measured Open-Loop Transfer Function $G_{vd}(s)$ for a Duty Cycle of 49.7%: (a) Gain, (b) Phase

The results from using the mean gain and phase difference that were acquired from the hardware experiment for the case of a desired output of 22.9V and a duty cycle of 62.1% are depicted in Figure 55.

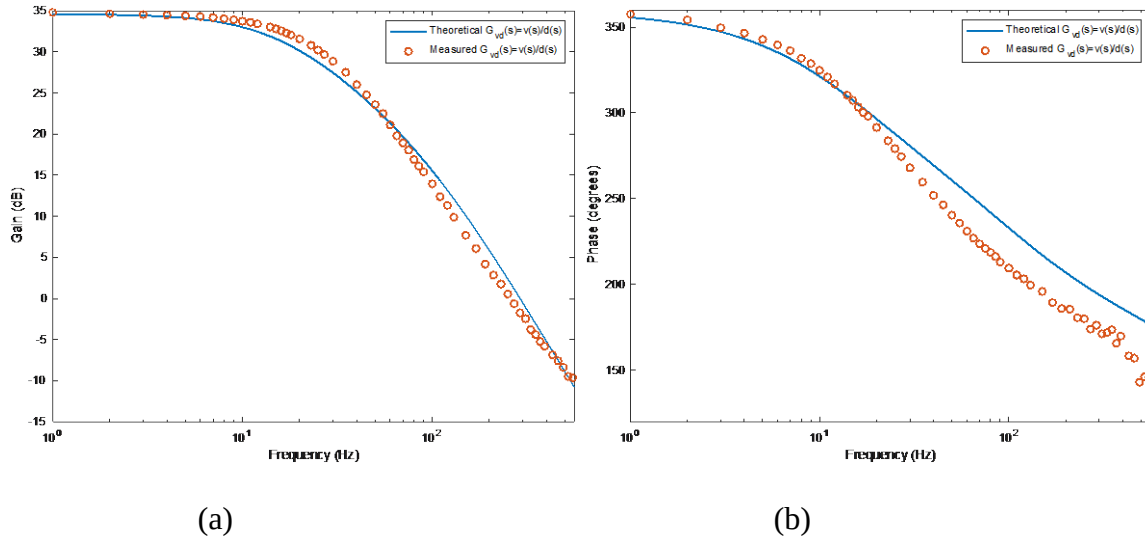


Figure 55. Comparison between Theoretical and Measured Open-Loop Transfer Function $G_{vd}(s)$ for a Duty Cycle of 62.1%: (a) Gain, (b) Phase

The results from using the gain and phase difference that were acquired from Simulink experiment for the case of a desired output of 20V and a duty cycle of 55.9% are depicted in Figure 56.

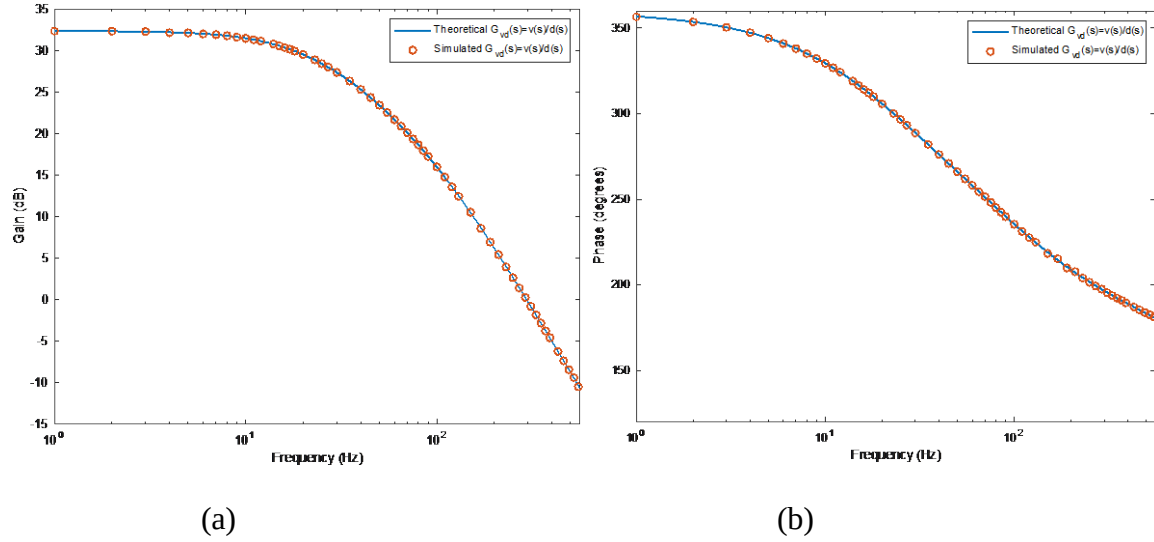


Figure 56. Comparison between Theoretical and Simulated Open-Loop Transfer Function $G_{vd}(s)$ for a Duty Cycle of 55.9%: (a) Gain, (b) Phase

The results prove that (37) successfully represents a non-ideal boost converter. Furthermore, the Simulink model matches an actual hardware model. Therefore, the model can be used to simulate the boost converter with different control strategies. Comparing different control strategies in Simulink/MATLAB is equivalent to comparing them in the laboratory using a hardware setup.

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V. CONTROL METHODOLOGIES COMPARISON— SIMULATION RESULTS

A physics-based model of the boost converter, presented in Figure 1, was implemented using Simulink/MATLAB. The values of the parameters used are shown in Table 3. Five different control models were implemented, as presented in Figure 17, Figure 18, Figure 28, Figure 29, and Figure 36, respectively. A time delay due to the A/D was added after each feedback loop. Each model was tuned for a load resistance of 50 Ω . The following scenario was applied to every control method. When the system operated at steady state, four step responses due to load variation were simulated, in order to examine the boost converter in all possible modes of operation. First, the load was changed from 50 Ω to 185 Ω , and after the system achieved steady state the load was changed back to 50 Ω . The next step response was changing the load resistance from 50 Ω to 500 Ω . The third step load change was changing from 500 Ω to 50 Ω . Then, after the system achieved steady state, the load changed to 400 Ω . Finally, the fourth step load change was changing the load resistance from 400 Ω to 1600 Ω . The simulation scenario is shown in Table 7. The performance for each controller was evaluated by the command voltage tracking capability, the robustness to the disturbances, and the time to reach the steady state for each step response.

Table 7. Simulation Scenario

State	Initial Values			Final Values		
Rating	R(Ω)	P(W)	Mode	R(Ω)	P(W)	Mode
Step 1	50	800	CCM	185	216.2	CCM
Step 2	50	800	CCM	500	80	DCM
Step 3	500	80	DCM	50	800	CCM
Step 4	400	100	DCM	1600	25	DCM

A. CONTROL METHODOLOGIES PARAMETERS AND EVALUATION

The first modification of K-factor control compensator was set up with parameters and results as shown in Table 8.

Table 8. Mod 1 K-Factor Compensator Parameters and Evaluation

Mod 1 K – Factor				
Gains	CCM		DCM	
ω_z (rad/sec)	320.98		1080.04	
ω_p (rad/sec)	27207.99		3902595.88	
K_c	20945.41		8342236824.4	
K	84.76		3613.37	
$Rload(\Omega)$	Step from 50 Ω to 185 Ω	Step from 50 Ω to 500 Ω	Step from 500 Ω to 50 Ω	Step from 400 Ω to 1600 Ω
Time to SS(msec)	10.7	8	12.1	4.6
Max Ripple (%)	0.45	0.55	0.55	0.05
Figure	57a	58a	59a	60a

The second modification of K-factor control compensator was set up with parameters and results as shown in Table 9.

Table 9. Mod 2 K-Factor Compensator Parameters and Evaluation

Mod 2 K – Factor				
Gains	CCM		DCM	
ω_z (rad/sec)	320.98		1080.04	
ω_p (rad/sec)	27207.99		3902595.88	
K_c	20945.41		8342236824.4	
K	84.76		3613.37	
$Rload(\Omega)$	Step from 50 Ω to 185 Ω	Step from 50 Ω to 500 Ω	Step from 500 Ω to 50 Ω	Step from 400 Ω to 1600 Ω
Time to SS(msec)	4.2	4.7	24	<1
Max Ripple (%)	0.55	0.7	1.5	0.1
Figure	57b	58b	59b	60b

The first modification of voltage mode control compensator was set up with parameters and results as shown in Table 10.

Table 10. Mod 1 Voltage Mode Control Compensator Parameters and Evaluation

Mod 1 Voltage Mode Control				
Gains	CCM		DCM	
	Inner loop	Outer loop	Inner loop	Outer loop
K_p	0.095	5.53	0.96	11.71
K_i	1351.95	10089.22	131740.74	6850.02
$Rload(\Omega)$	Step from 50 Ω to 185 Ω	Step from 50 Ω to 500 Ω	Step from 500 Ω to 50 Ω	Step from 400 Ω to 1600 Ω
Time to SS(msec)	3.3	11.5	3.2	2.2
Max Ripple (%)	0.35	0.4	0.5	<0.05
Figure	57c	58c	59c	60c

The second modification of voltage mode control compensator was set up with parameters and results as shown in Table 11.

Table 11. Mod 2 Voltage Mode Control Compensator Parameters and Evaluation

Mod 2 Voltage Mode Control				
Gains	CCM		DCM	
	Inner loop	Outer loop	Inner loop	Outer loop
K_p	0.095	5.53	0.96	11.71
K_i	1351.95	10089.22	131740.74	6850.02
$Rload(\Omega)$	Step from 50 Ω to 185 Ω	Step from 50 Ω to 500 Ω	Step from 500 Ω to 50 Ω	Step from 400 Ω to 1600 Ω
Time to SS(msec)	4.3	5.6	3.2	14.7
Max Ripple (%)	0.2	0.25	0.5	0.1
Figure	57d	58d	59d	60d

The current mode control compensator was set up with parameters and results as shown in Table 12.

Table 12. Current Mode Control Compensator Parameters and Evaluation

Current Mode Control				
K_p	9.12			
K_i	13920.36			
$\frac{G_{ld}}{G_{vd}} _d$	0.086			
Slope	53191.49			
$Rload(\Omega)$	Step from 50 Ω to 185 Ω	Step from 50 Ω to 500 Ω	Step from 500 Ω to 50 Ω	Step from 400 Ω to 1600 Ω
Time to SS(msec)	1.5	4.3	2.1	4.9
Max Ripple (%)	0.25	0.3	0.4	0.1
Figure	57e	58e	59e	60e

B. COMPARISON OF THE CONTROL METHODOLOGIES

The scenarios were simulated for each controller. The voltage output during the transient response, due to the occurrence of the step load change, was plotted, and comparison of the results was made. The step response of the boost converter was simulated and the results are shown in Figure 57, Figure 58, Figure 59, and Figure 60.

The first step to the load from 50 Ω to 185 Ω happened at 0.03 s, as is clearly visible in Figure 57. In this case the boost converter operated in CCM. All the controllers achieved similar performance, with the current mode control achieving the smaller ripple and the fastest transient response.

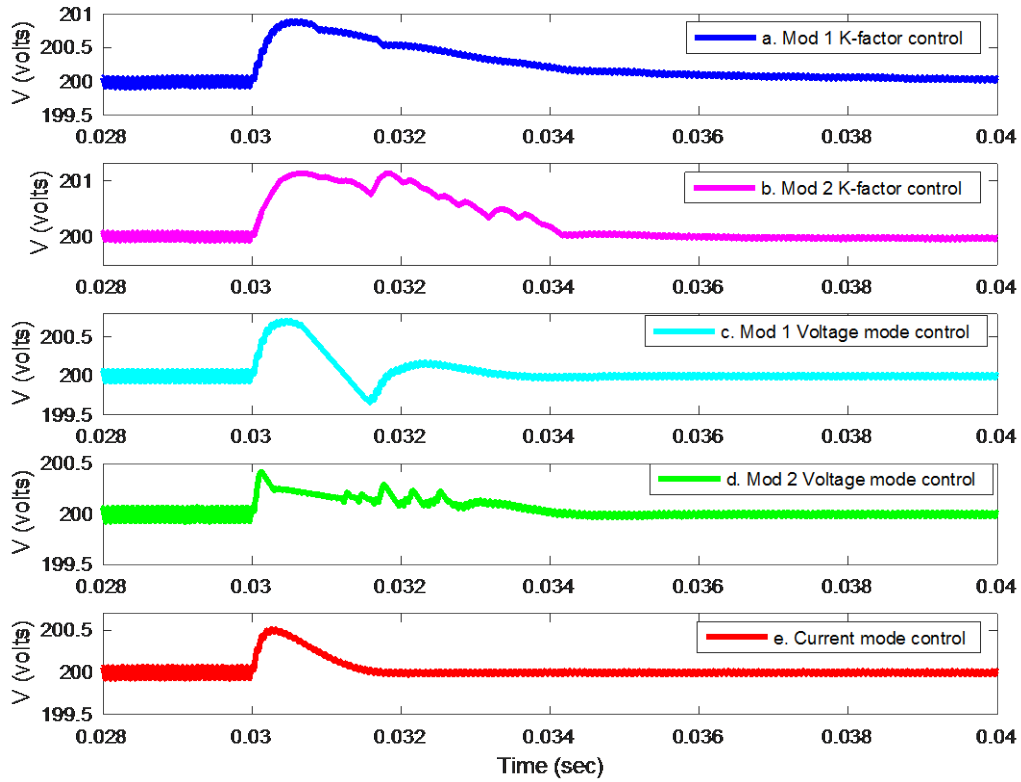


Figure 57. Closed-Loop Step Response for Load Change from 50 Ω to 185 Ω : (a) Mod 1 K-Factor Control, (b) Mod 2 K-Factor Control, (c) Mod 1 Voltage Mode Control, (d) Mod 2 Voltage Mode Control (e) Current Mode Control

The second step to the load from 50 Ω to 500 Ω happened at 0.07 s, as is clearly visible in Figure 58. In this case the boost converter initially operated in CCM and then changed to DCM. Mod 2 voltage mode control achieved the smaller voltage ripple and the fastest transient response.

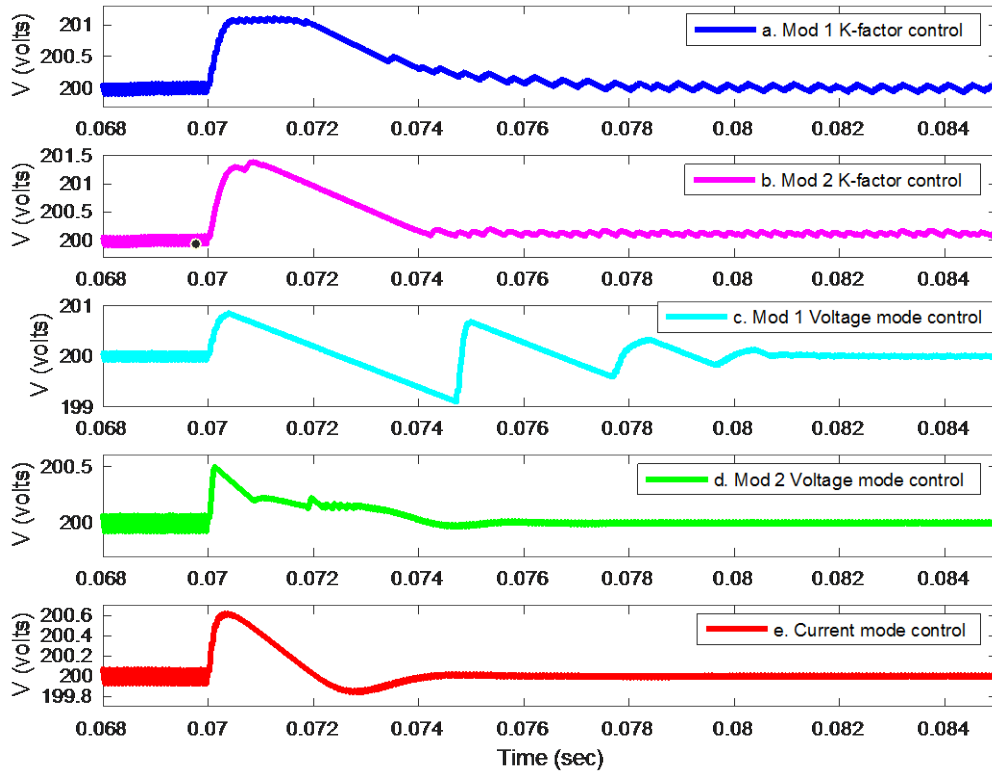


Figure 58. Closed-Loop Step Response for Load Change from 50 Ω to 500 Ω : (a) Mod 1 K-Factor Control, (b) Mod 2 K-Factor Control, (c) Mod 1 Voltage Mode Control, (d) Mod 2 Voltage Mode Control (e) Current Mode Control

The third step to the load from 500 Ω to 50 Ω happened at 0.30 s, as is clearly visible in Figure 59. In this case, the boost converter initially operated in DCM and then changed to CCM. Current mode control achieved the smaller voltage ripple and the fastest transient response.

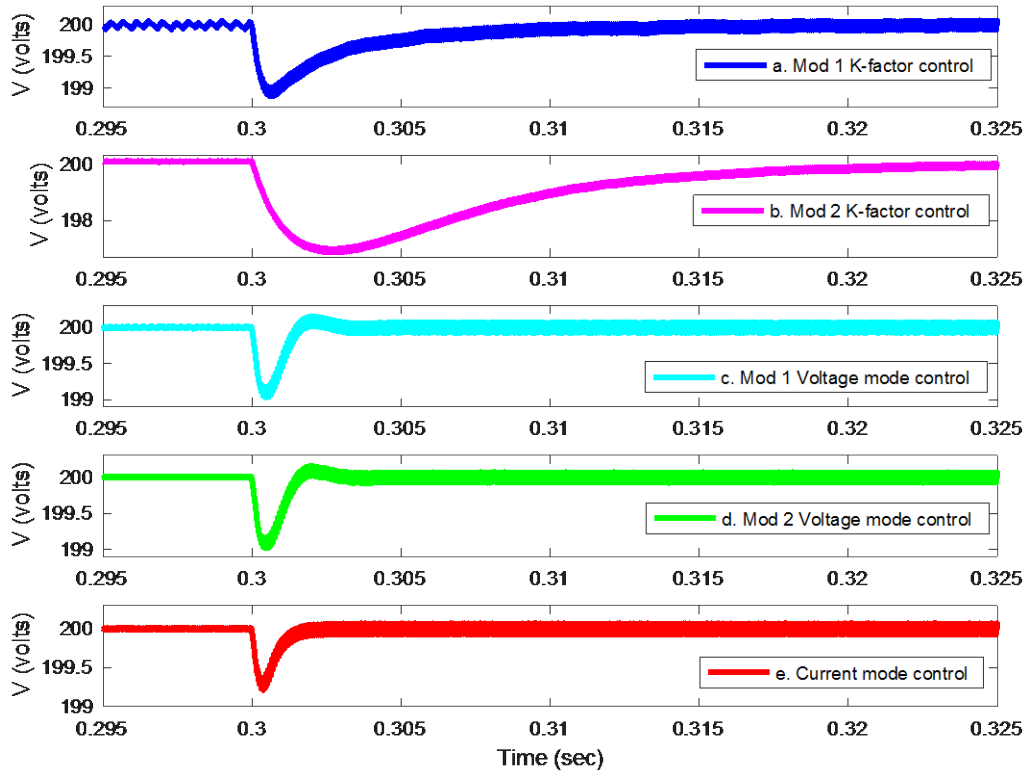


Figure 59. Closed-Loop Step Response for Load Change from 500 Ω to 50 Ω :
(a) Mod 1 K-Factor Control, (b) Mod 2 K-Factor Control,
(c) Mod 1 Voltage Mode Control, (d) Mod 2 Voltage Mode Control
(e) Current Mode Control

The fourth step to the load from 400 Ω to 1600 Ω happened at 0.45 s, as is clearly visible in Figure 60. In this case the boost converter operated in DCM, and all the controllers achieved similar performance, with the Mod 1 voltage mode control achieving the smaller ripple and the fastest transient response.

The simulated waveforms demonstrate that all the proposed controllers responded well to small and large changes of the load. Evaluating each of the controller's performance, we observed that all of them are stable. The maximum ripple was less than 1% in all the cases except the second modification of the K-factor, and in all the cases the time to reach steady state was sufficient. In a system without time delays all the controllers had almost the same performance, but with time delays it seems that the K-factor

compensator is the most sensitive to this nonlinearity. In particular, the current mode control and the second modification of the voltage mode control are operating almost with the same performance as in a system without time delays. Furthermore, observing the simulation during the DCM mode of operation, the ripple is smaller and the transient response is faster compared to the other modes simulated. Finally, from the above simulated cases we can see that the performance of the second modification of the voltage mode control method is close to the one of the current mode control methodology.

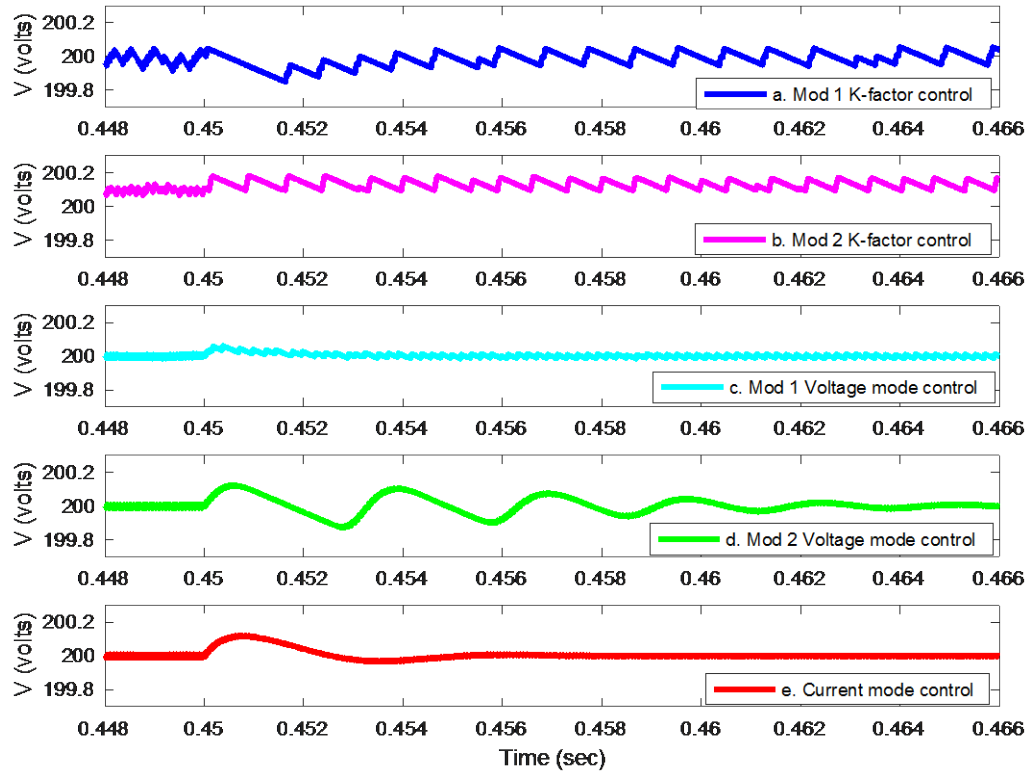


Figure 60. Closed-Loop Step Response for Load Change from 400 Ω to 1600 Ω :
(a) Mod 1 K-Factor Control, (b) Mod 2 K-Factor Control, (c) Mod 1
Voltage Mode Control, (d) Mod 2 Voltage Mode Control (e) Current
Mode Control

VI. CONCLUSIONS AND FUTURE WORK

The models, simulations, and results described in this thesis lead to the conclusions described in this section. There are several opportunities for follow-on work.

A. CONCLUSIONS

Effectively controlling the power converters is essential in increasing the energy efficiency and reducing the energy cost. Therefore, various methods to compensate the output of a DC-DC boost converter have been developed and improved over time.

To study the control of a non-ideal boost converter, a mathematical model of the converter needs to be constructed. A hardware or a simulation experiment can be used to verify the mathematical model. Mod 1 K-factor and Mod 2 K-factor successfully improved the performance of the regular K-factor, having a faster transient response and a much smaller ripple. A significant benefit of the Mod 2 K-factor is that it uses only one voltage sensor, so implementing a Mod 2 K-factor mode compensator is less complicated and more cost effective, since the extra cost of installing a Hall Effect current sensor in the system is avoided. Mod 1 voltage mode control and Mod 2 voltage mode control overcome the limitations of the regular voltage mode control. These modified voltage mode compensators have faster transient response and smaller ripple compared to the regular controller.

In a system without time delays, the performance of all controllers is similar, but when a more realistic system with time delays is simulated, the performance of each compensator varies. From the results of all the simulations, we concluded that during the CCM and DCM modes of operation, all control methodologies resulted in similar performance. During the transition from CCM to DCM, however, the performance of each controller varied. In particular, during the change from DCM to CCM the Mod 2 K-factor had a bigger ripple and slower transient response compared to the other control methodologies. During the DCM all the compensators responded with a much lower ripple and faster transient response compared with the other modes of operation. Finally, a converter designed to operate with light loads operated in DCM. In that mode of operation,

the converter can be regulated more effectively, with a very small ripple and very fast transient response due to step load changes.

B. RECOMMENDATIONS FOR FUTURE WORK

There are several opportunities to continue the work accomplished in this thesis with follow-on modeling, experimental implementation, and simulations.

First of all, more parasitic components could be added to the system. A parasitic capacitor could be added to the switch. This parasitic capacitor results in an additional time delay from charging and discharging [12]. Furthermore, parasitic resistances and inductors could be added due to the current and voltage sensors.

Different control strategies could be implementing using only voltage sensors. First, a linear quadratic regulator (LQR) could be implemented to control the boost converter using only one voltage sensor; this control methodology gives the optimal solution for the given criteria and cost function. Second, the Mod 3 K-factor control could be examined. The Mod 3 K-factor control could use different control parameters depending on the load. An extended Kalman filter (EKF) could be implemented to predict the load by measuring the output voltage and the inductor voltage. In this way, appropriate parameters depending on the load could be selected. Reference [13] suggests the implementation of an algorithm to predict the inductor current for the current mode control; this algorithm could be tested to replace the current sensor of Mod 1 K-factor, Mod 1 voltage mode control, and Mod 2 voltage mode control compensator. Finally, an EKF could be implemented and evaluated on its ability to predict the inductor current, in order to replace the current sensor.

Furthermore, a second DC-DC boost converter as a source with supercapacitors (SCs) could be added to the system to provide the necessary power to the load during the transient response, to extend the battery lifetime [25]. This suggestion could be implemented by treating the two switches as one switch with four possible states. After deriving the system transfer function, a LQR compensator could be set up. The cost function of the LQR is the summation of the cost function for the battery anticipating power, the SCs anticipating power, the final state reach, and the control effort. The battery cost function could have a high cost if the delivering power from the battery changes fast,

and low cost if the delivering power from the battery changes slowly. The opposite could be implemented for the SCs' cost function. Therefore, the system will mainly provide power from the battery to the load during the steady state and power from the SCs during transient response. Furthermore, the addition of a DC-DC buck converter is necessary to have the ability to charge the SCs.

C. IMPLICATIONS FOR THE NAVY

This research contributes to more effectively controlling DC-DC boost converters, which are key components to ships and other naval platforms. Two modifications for each regular K-factor and regular voltage mode control were implemented. The performance of those modifications was compared to the current mode control compensator. Knowing the performance of each compensator contributed to selecting the suitable controller for each application.

More effectively controlling the DC-DC boost converters of a ship has several implications for any navy. Controlling the electrical energy in a ship more effectively results in increased fuel efficiency. Furthermore, the operational cost of a ship is reduced for two reasons. First, less power is consumed to charge the batteries, which leads to lower fuel consumption. Second, since the batteries are used more effectively, their lifetime is increased, resulting in less frequent replacement. Since less fuel is consumed, the environmental impact of the ship is reduced and its endurance is increased. Finally, a naval ship performs different missions and at any time the maximum available power might be required to overcome an enemy. Having more effective power converters might be vital for the survival of the ship and the success of its mission.

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APPENDIX. MATLAB SCRIPTS

A. MODEL CALL-BACK FUNCTION

```
% Simulate the models
clear all;
close all;

%% Voltage Mode

My_analysis_microcrd_Boost_2PI_Current
% regular
sim boost_regular_2PI_Current
save_simulink_data_regular_V
% mod 1
sim boost_2PI_Current
save_simulink_data_V
%mod 2
sim boost_mod2_2PI_Current
save_simulink_data_mod2_V

%% K factor

My_analysis_microcrd_Boost
% regular
sim boost_regular_K_factor
save_simulink_data_reg_K
% mod 1
sim boost_new_K_factor
save_simulink_data_K
% mod 2
sim boost_new_K_factor_experiment
save_simulink_data_K_mod2

%% Current Mode

My_analysis_microcrd_Boost_Current_Control
% Regular
sim boost_forbattery_iff
save_simulink_data_F
```

B. K-FACTOR PARAMETERS

```
% K-factor parameters
clear all;
close all;
%% Parameters
Rload=50; %load
sw_freq=15000; %fs
fs=sw_freq;
Fc=fs
```

```

Ts=1/fs;
Vbattery=100;
Lboost=940e-6;
Rbattery=.4;
Cbus=990e-6;
Vdc_ref=200;
Vload=Vdc_ref;
Vd=0.7;
tstep=5e-7;
%Ts=tstep;
%fs=1/tstep;
%% Steady State Values
% D=(Vdc_ref-Vbattery)/Vdc_ref
Dp=(Vbattery-Vd+sqrt((Vbattery-Vd)^2-4*Rbattery*(Vload^2)/Rload))/2/Vload
D=1-Dp
Rcrit_min=(27/2)*Lboost/Ts %Tab 5-1
Rcrit=2*Lboost/D/Ts/(1-D)^2
%Iload_ideal=Vbattery/Rload/Dp^2
IL_dc=(Vbattery-Vd)/(Rbattery+Rload*Dp^2);

%% Small signal analysis
A=[-Rbattery/Lboost,-Dp/Lboost;Dp/Cbus,-1/Rload/Cbus];
Bv=[1/Lboost;0];
Bd=[Vload/Lboost;-IL_dc/Cbus]
C=[0,1];
Ed=0;
Ev=0;
Cl=[1,0]

%% tranfer function Gvd
[num1 den1]=ss2tf(A,Bd,C,Ed,1);
[num2 den2]=ss2tf(A,Bv,C,Ev,1);
[num3 den3]=ss2tf(A,Bd,Cl,Ev,1); %Gid
[num4 den4]=ss2tf(A,Bv,Cl,Ev,1); %Gig

%% Body plot
Gid=tf(num3,den3);
Gig=tf(num4,den4);
Gvd=tf(num1,den1)
figure(1)
bode(Gvd)
grid on;
Gvg=tf([Dp/Lboost/Cbus],[1,Rbattery/Lboost+1/Rload/Cbus,(Dp^2)/Lboost/Cbus+Rbattery/Rload/Cbus/Lboost])
figure(2)
bode(Gvg)
grid on;
num=num1
den=den1
%% K factor design
wSW=2*pi*sw_freq
w1=sqrt((Dp^2)/Lboost/Cbus+Rbattery/Rload/Cbus/Lboost)
wRHP=(Rload*Dp^2-Rbattery)/Lboost
wGC=wRHP/4

```

```

TF_wGC=polyval(num, j*wGC)/polyval(den, j*wGC);
phase_wGC=angle(TF_wGC)*180/pi; % phase at wGC, in degrees
gain_wGC=abs(TF_wGC); % gain at wGC
%
PhBoost=(150+180-phase_wGC)*pi/180; % desired phase Boost from compensator in radians
K=(tan(PhBoost/4+pi/4))^2;
wz=wGC/sqrt(K); wp=wGC*sqrt(K); % zero and pole of type 3 compensator
Kc=K*wGC/gain_wGC;
numc=Kc*[1, 2*wz, wz^2];
denc=[1, 2*wp, wp^2, 0];
C=tf(numc,denc)
figure(3)
bode(Gvd*C)
grid on;
%% K factor for DCM
Rload=500;
Dp=(Vbattery-Vd+sqrt((Vbattery-Vd)^2-4*Rbattery*(Vload^2)/Rload))/2/Vload
D=1-Dp;
IL_dc=(Vbattery-Vd)/(Rbattery+Rload*Dp^2);
A=[-Rbattery/Lboost,-Dp/Lboost;Dp/Cbus,-1/Rload/Cbus];
Bv=[1/Lboost;0];
Bd=[Vload/Lboost;-IL_dc/Cbus]
C=[0,1];
Ed=0;
Ev=0;
[num1 den1]=ss2tf(A,Bd,C,Ed,1);
[num2 den2]=ss2tf(A,Bv,C,Ev,1);
num=num1;
den=den1;

wSW=2*pi*sw_freq
w1=sqrt((Dp^2)/Lboost/Cbus+Rbattery/Rload/Cbus/Lboost)
wRHP=(Rload*Dp^2-Rbattery)/Lboost
wGC=wRHP/2
TF_wGC=polyval(num, j*wGC)/polyval(den, j*wGC);
phase_wGC=angle(TF_wGC)*180/pi; % phase at wGC, in degrees
gain_wGC=abs(TF_wGC); % gain at wGC
%
PhBoost=(150+180-phase_wGC)*pi/180; % desired phase Boost from compensator in radians
K=(tan(PhBoost/4+pi/4))^2;
wz=wGC/sqrt(K); wp=wGC*sqrt(K); % zero and pole of type 3 compensator
Kc=K*wGC/gain_wGC;
numc2=Kc*[1, 2*wz, wz^2];
denc2=[1, 2*wp, wp^2, 0];

```

C. VOLTAGE MODE CONTROL PARAMETERS

```

% Voltage mode control
clear all;
close all;
%% Parameters
Rload=50;
sw_freq=15000;

```

```

fs=sw_freq;
Fc=fs
Ts=1/fs;
Vbattery=100;
Lboost=940e-6;
Rbattery=0.4;
Cbus=990e-6;
Vdc_ref=200;
Vload=Vdc_ref;
Vd=0.7;
tstep=5e-7;
%Ts=tstep;
%fs=1/tstep;
%% Steady State Values
% D=(Vdc_ref-Vbattery)/Vdc_ref
Dp=(Vbattery-Vd+sqrt((Vbattery-Vd)^2-4*Rbattery*(Vload^2)/Rload))/2/Vload
D=1-Dp
Rcrit_min=(27/2)*Lboost/Ts %Tabe 5-1
Rcrit=2*Lboost/D/Ts/(1-D)^2
%Iload_ideal=Vbattery/Rload/Dp^2
IL_dc=(Vbattery-Vd)/(Rbattery+Rload*Dp^2)e

%% Small signal analysis
A=[-Rbattery/Lboost,-Dp/Lboost;Dp/Cbus,-1/Rload/Cbus];
Bv=[1/Lboost;0];
Bd=[Vload/Lboost;-IL_dc/Cbus]
C=[0,1];
Ed=0;
Ev=0;
CI=[1,0]
%% tranfer function Gvd
[num1 den1]=ss2tf(A,Bd,C,Ed,1)
[num2 den2]=ss2tf(A,Bv,C,Ev,1)
[num3 den3]=ss2tf(A,Bd,CI,Ev,1) %Gid
[num4 den4]=ss2tf(A,Bv,CI,Ev,1) %Gig

%% Body plot
Gvd=tf(num1,den1)

figure(1)
bode(Gvd)
grid on;
Gvg=tf(num2,den2)
figure(2)
bode(Gvg)
grid on;
num=num3
den=den3
%% PI Current design

wSW=2*pi*sw_freq
w1=sqrt((Dp^2)/Lboost/Cbus+Rbattery/Rload/Cbus/Lboost)
wRHP=(Rload*Dp^2-Rbattery)/Lboost
wGC1=wRHP/0.5

```

```

TF_wGC=polyval(num, j*wGC1)/polyval(den, j*wGC1);
phase_wGC=angle(TF_wGC)*180/pi; % phase at wGC, in degrees
gain_wGC=abs(TF_wGC); % gain at wGC

Kp=abs(cos((60-phase_wGC)*pi/180))/gain_wGC
Ti=-1/wGC1/tan((60-phase_wGC)*pi/180)
Ki=abs(Kp/Ti)

%% Voltage PI
% start with tf
Gvi=tf([-Lboost,Dp*Rload-Rbattery],[(Dp^2)*Rload*Cbus,2*Dp])
numVI=[-Lboost,Dp*Rload-Rbattery]
denVI=[Dp*Rload*Cbus,2*Dp]
wGC2=wRHP/2
TF_wGC_VI=polyval(numVI, j*wGC2)/polyval(denVI, j*wGC2);
phase_wGC_VI=angle(TF_wGC_VI)*180/pi; % phase at wGC, in degrees
gain_wGC_VI=abs(TF_wGC_VI); % gain at wGC

KpV=abs(cos((60-phase_wGC_VI)*pi/180))/gain_wGC_VI
TiV=-1/wGC2/tan((60-phase_wGC_VI)*pi/180)
KiV=abs(KpV/TiV)

%% DCM
Rload2=50%50;
Dp2=(Vbattery-Vd+sqrt((Vbattery-Vd)^2-4*Rbattery*(Vload^2)/Rload2))/2/Vload
D2=1-Dp2;
IL_dc2=(Vbattery-Vd)/(Rbattery+Rload2*Dp2^2);
A=[-Rbattery/Lboost,-Dp2/Lboost;Dp2/Cbus,-1/Rload2/Cbus];
Bv=[1/Lboost;0];
Bd=[Vload/Lboost;-IL_dc2/Cbus]
C=[0,1];
Ed=0;
Ev=0;
[num1 den1]=ss2tf(A,Bd,C,Ed,1);
[num2 den2]=ss2tf(A,Bv,C,Ev,1);
[num3 den3]=ss2tf(A,Bd,[1,0],Ed,1);%Gid
wRHP2=(Rload2*Dp2^2-Rbattery)/Lboost;

Gvi2=tf([-Lboost,Dp2*Rload2-Rbattery],[(Dp2^2)*Rload2*Cbus,2*Dp2]);
numVI=[-Lboost,Dp2*Rload2-Rbattery];
denVI=[Dp2*Rload2*Cbus,2*Dp2];

num=num3;
den=den3;

wSW=2*pi*sw_freq
w1=sqrt((Dp2^2)/Lboost/Cbus+Rbattery/Rload2/Cbus/Lboost)

wGC3=wRHP2/0.05
TF_wGC=polyval(num, j*wGC3)/polyval(den, j*wGC3);
phase_wGC=angle(TF_wGC)*180/pi; % phase at wGC, in degrees
gain_wGC=abs(TF_wGC); % gain at wGC

```

```

Kp2=abs(cos((60-phase_wGC)*pi/180))/gain_wGC
Ti=-1/wGC3/tan((60-phase_wGC)*pi/180)
Ki2=abs(Kp2/Ti)

```

```

wGC4=wRHP2/0.9;
TF_wGC_VI=polyval(numVI, j*wGC4)/polyval(denVI, j*wGC4);
phase_wGC_VI=angle(TF_wGC_VI)*180/pi; % phase at wGC, in degrees
gain_wGC_VI=abs(TF_wGC_VI); % gain at wGC

```

```

KpV2=abs(cos((60-phase_wGC_VI)*pi/180))/gain_wGC_VI
TiV=-1/wGC4/tan((60-phase_wGC_VI)*pi/180)
KiV2=abs(KpV2/TiV)

```

D. CURRENT MODE PARAMETERS

```

% Current mode control
clear all;
close all;
%% Parameters
slope_compensation=1
Rload=50;
sw_freq=15000;
fs=sw_freq;
Fc=fs;
Ts=1/fs;
Vbattery=100;
Lboost=940e-6;
Rbattery=0.4;
Cbus=990e-6;
Vdc_ref=200;
Vload=Vdc_ref;
Vd=0.7;
tstep=5e-7;
%Ts=tstep;
%fs=1/tstep;
%% Steady State Values
% D=(Vdc_ref-Vbattery)/Vdc_ref
Dp=(Vbattery-Vd+sqrt((Vbattery-Vd)^2-4*Rbattery*(Vload^2)/Rload))/2/Vload
D=1-Dp
Rcrit_min=(27/2)*Lboost/Ts %Tabe 5-1
Rcrit=2*Lboost/D/Ts/(1-D)^2
%Iload_ideal=Vbattery/Rload/Dp^2
IL_dc=(Vbattery-Vd)/(Rbattery+Rload*Dp^2)

%% Small signal analysis
A=[-Rbattery/Lboost,-Dp/Lboost,Dp/Cbus,-1/Rload/Cbus];
Bv=[1/Lboost;0];
Bd=[Vload/Lboost;-IL_dc/Cbus]
C=[0,1];
Ed=0;
Ev=0;

```

```

%% tranfer function Gvd
[num1 den1]=ss2tf(A,Bd,C,Ed,1)
[num2 den2]=ss2tf(A,Bv,C,Ev,1)
[num3 den3]=ss2tf(A,Bd,[1,0],Ed,1)%Gid

%% Body plot
Gvd=tf(num1,den1)
figure(1)
bode(Gvd)
grid on;
Gvg=tf(num2,den2)
figure(2)
bode(Gvg)
grid on;
num=num2
den=den2
%% PI design

wRHP=(Rload*Dp^2-Rbattery)/Lboost
% Kp -Ki selection
%start with tf
wGC=wRHP/1.2
Gvi=tf([-Lboost,Dp*Rload-Rbattery],[(Dp^2)*Rload*Cbus,2*Dp])
numVI=[-Lboost,Dp*Rload-Rbattery]
denVI=[Dp*Rload*Cbus,2*Dp]
TF_wGC_VI=polyval(numVI, j*wGC)/polyval(denVI, j*wGC);
phase_wGC_VI=angle(TF_wGC_VI)*180/pi; % phase at wGC, in degrees
gain_wGC_VI=abs(TF_wGC_VI); % gain at wGC

KpV=abs(cos((60-phase_wGC_VI)*pi/180))/gain_wGC_VI
TiV=-1/wGC/tan((60-phase_wGC_VI)*pi/180)
KiV=abs(KpV/TiV)
Kp=KpV
Ki=KiV
% iFF
iFF=2*Dp/(Rload*Dp^2-Rbattery);
wfilter=2*pi*100*10^3;

%% DCM gains
Rload2=2000;
Dp2=(Vbattery-Vd+sqrt((Vbattery-Vd)^2-4*Rbattery*(Vload^2)/Rload2))/2/Vload
D2=1-Dp2;
IL_dc2=(Vbattery-Vd)/(Rbattery+Rload2*Dp2^2);
A=[-Rbattery/Lboost,-Dp2/Lboost;Dp2/Cbus,-1/Rload2/Cbus];
Bv=[1/Lboost;0];
Bd=[Vload/Lboost;-IL_dc2/Cbus]
C=[0,1];
Ed=0;
Ev=0;
[num1 den1]=ss2tf(A,Bd,C,Ed,1);
[num2 den2]=ss2tf(A,Bv,C,Ev,1);
[num3 den3]=ss2tf(A,Bd,[1,0],Ed,1);%Gid
wRHP2=(Rload2*Dp2^2-Rbattery)/Lboost;

```

```

wGC2=wRHP2/1.2
Gvi=tf([-Lboost,Dp2*Rload2-Rbattery],[(Dp2^2)*Rload2*Cbus,2*Dp2])
numVI=[-Lboost,Dp2*Rload2-Rbattery]
denVI=[Dp2*Rload2*Cbus,2*Dp2]
TF_wGC_VI=polyval(numVI, j*wGC2)/polyval(denVI, j*wGC2);
phase_wGC_VI=angle(TF_wGC_VI)*180/pi; % phase at wGC, in degrees
gain_wGC_VI=abs(TF_wGC_VI); % gain at wGC

KpV=abs(cos((60-phase_wGC_VI)*pi/180))/gain_wGC_VI
TiV=-1/wGC2/tan((60-phase_wGC_VI)*pi/180)
KiV=abs(KpV/TiV)
Kp2=KpV
Ki2=KiV
iFF2=2*Dp2/(Rload2*Dp2^2-Rbattery);

```

E. PLOT THE RESULTS

```

% Plotting the results
%% K factor compare

```

```

figure (1)
subplot(3,1,1)
plot(time_K_reg,Vout_K_reg,'b','linewidth',3)
ylabel ('V (volts)')
ylim([199.5 201.3])
xlim([0.028 0.04])
legend('a. Regular K- factor')
%title(' Step response: 50\Omega to 185\Omega')
grid on;

```

```

subplot(3,1,2)
plot(time_K,Vout_K,'g','linewidth',3)
ylim([199.5 201.3])
xlim([0.028 0.04])
grid on;
ylabel ('V (volts)')
legend('b. Mod 1 K-factor')

```

```

subplot(3,1,3)
plot(time_K_m2,Vout_K_m2,'r','linewidth',3)
ylim([199.5 201.3])
xlim([0.028 0.04])
grid on;
ylabel ('V (volts)')
legend('c. Mod 2 K-factor')
xlabel('Time (sec)')

```

```

figure (2)
subplot(3,1,1)
plot(time_K_reg,Vout_K_reg,'b','linewidth',3)
ylabel ('V (volts)')

```

```

ylim([198 204])
xlim([0.06 0.27])
legend('a. Regular K- factor')
%title(' Step response: 50\Omega to 500\Omega')
grid on;

```

```

subplot(3,1,2)
plot(time_K,Vout_K,'g','linewidth',3)
ylim([199 201.5])
xlim([0.068 0.085])
grid on;
ylabel('V (volts)')
legend('b. Mod 1 K-factor')

```

```

subplot(3,1,3)
plot(time_K_m2,Vout_K_m2,'r','linewidth',3)
ylim([199 201.5])
xlim([0.068 0.085])
grid on;
ylabel('V (volts)')
legend('c. Mod 2 K-factor')
xlabel('Time (sec)')

```

```

figure (3)
subplot(3,1,1)
plot(time_K_reg,Vout_K_reg,'b','linewidth',3)
ylabel('V (volts)')
ylim([192.5 200.5])
xlim([0.295 0.35])
legend('a. Regular K- factor')
%title(' Step response: 500\Omega to 50\Omega')
grid on;

```

```

subplot(3,1,2)
plot(time_K,Vout_K,'g','linewidth',3)
ylim([198.5 200.5])
xlim([0.295 0.35])
grid on;
ylabel('V (volts)')
legend('b. Mod 1 K-factor')

```

```

subplot(3,1,3)
plot(time_K_m2,Vout_K_m2,'r','linewidth',3)
ylim([196 200.5])
xlim([0.295 0.35])
grid on;
ylabel('V (volts)')
legend('c. Mod 2 K-factor')
xlabel('Time (sec)')

```

```

%% Voltage Mode control
figure (4)
subplot(3,1,1)

```

```

plot(time_V_reg,Vout_V_reg,'b','linewidth',3)
ylabel('V (volts)')
ylim([199.5 200.8])
xlim([0.029 0.038])
legend('a. Regular Voltage mode control')
%title(' Step responce: 50\Omega to 185\Omega')
grid on;

```

```

subplot(3,1,2)
plot(time_V,Vout_V,'g','linewidth',3)
ylim([199.5 200.8])
xlim([0.029 0.038])
grid on;
ylabel('V (volts)')
legend('b. Mod 1 Voltage mode control')

```

```

subplot(3,1,3)
plot(time_V_mod2,Vout_V_mod2,'r','linewidth',3)
ylim([199.5 200.8])
xlim([0.029 0.038])
grid on;
ylabel('V (volts)')
legend('c. Mod 2 Voltage mode control')
xlabel('Time (sec)')

```

```

figure (5)
subplot(2,1,1)
plot(time_V,Vout_V,'g','linewidth',3)
ylim([199 201])
xlim([0.068 0.085])
grid on;
ylabel('V (volts)')
legend('a. Mod 1 Voltage mode control')
%title(' Step responce: 50\Omega to 500\Omega')

```

```

subplot(2,1,2)
plot(time_V_mod2,Vout_V_mod2,'r','linewidth',3)
ylim([199.5 200.6])
xlim([0.068 0.085])
grid on;
ylabel('V (volts)')
legend('b. Mod 2 Voltage mode control')
xlabel('Time (sec)')

```

```

figure (6)
subplot(2,1,1)
plot(time_V,Vout_V,'g','linewidth',3)
ylim([198.8 200.2])
xlim([0.295 0.33])
grid on;
ylabel('V (volts)')
legend('a. Mod 1 Voltage mode control')
%title(' Step responce: 500\Omega to 50\Omega')

```

```

subplot(2,1,2)
plot(time_V_mod2,Vout_V_mod2,'r','linewidth',3)
ylim([198.8 200.2])
xlim([0.295 0.33])
grid on;
ylabel('V (volts)')
legend('b. Mod 2 Voltage mode control')
xlabel('Time (sec)')

%% Final results compare
%% 50->185
figure (7)
subplot(5,1,1)
plot(time_K,Vout_K,'b','linewidth',3)
ylim([199.5 201])
xlim([0.028 0.04])
grid on;
ylabel('V (volts)')
legend('a. Mod 1 K-factor control')
%title(' Step responce: 50\Omega to 185\Omega')

subplot(5,1,2)
plot(time_K_m2,Vout_K_m2,'m','linewidth',3)
ylim([199.5 201.3])
xlim([0.028 0.04])
grid on;
ylabel('V (volts)')
legend('b. Mod 2 K-factor control')
%title(' Step responce: 50\Omega to 185\Omega')

subplot(5,1,3)
plot(time_V,Vout_V,'c','linewidth',3)
ylabel('V (volts)')
ylim([199.5 200.8])
xlim([0.028 0.04])
legend('c. Mod 1 Voltage mode control')
grid on;

subplot(5,1,4)
plot(time_V_mod2,Vout_V_mod2,'g','linewidth',3)
ylabel('V (volts)')
ylim([199.5 200.5])
xlim([0.028 0.04])
legend('d. Mod 2 Voltage mode control')
grid on;

subplot(5,1,5)
plot(time_F,Vout_F,'r','linewidth',3)
ylim([199.5 200.7])
xlim([0.028 0.04])
grid on;
legend('e. Current mode control')

```

```

ylabel ('V (volts)')
xlabel('Time (sec)')

%50->500
figure (8)
subplot(5,1,1)
plot(time_K,Vout_K,'b','linewidth',3)
ylim([199.7 201.3])
xlim([0.068 0.085])
grid on;
ylabel ('V (volts)')
legend('a. Mod 1 K-factor control')
%title(' Step responce: 50\Omega to 185\Omega')

subplot(5,1,2)
plot(time_K_m2,Vout_K_m2,'m','linewidth',3)
ylim([199.7 201.5])
xlim([0.068 0.085])
grid on;
ylabel ('V (volts)')
legend('b. Mod 2 K-factor control')
%title(' Step responce: 50\Omega to 185\Omega')

subplot(5,1,3)
plot(time_V,Vout_V,'c','linewidth',3)
ylabel ('V (volts)')
ylim([199 201])
xlim([0.068 0.085])
legend('c. Mod 1 Voltage mode control')
grid on;

subplot(5,1,4)
plot(time_V_mod2,Vout_V_mod2,'g','linewidth',3)
ylabel ('V (volts)')
ylim([199.7 200.6])
xlim([0.068 0.085])
legend('d. Mod 2 Voltage mode control')
grid on;

subplot(5,1,5)
plot(time_F,Vout_F,'r','linewidth',3)
ylim([199.7 200.7])
xlim([0.068 0.085])
grid on;
legend('e. Current mode control')
ylabel ('V (volts)')
xlabel('Time (sec)')

%500->50
figure (9)
subplot(5,1,1)
plot(time_K,Vout_K,'b','linewidth',3)
ylim([198.7 200.3])

```

```

xlim([0.295 0.325])
grid on;
ylabel('V (volts)')
legend('a. Mod 1 K-factor control')
%title(' Step response: 50\Omega to 185\Omega')

subplot(5,1,2)
plot(time_K_m2,Vout_K_m2,'m','linewidth',3)
ylim([196.7 200.3])
xlim([0.295 0.325])
grid on;
ylabel('V (volts)')
legend('b. Mod 2 K-factor control')
%title(' Step response: 50\Omega to 185\Omega')

subplot(5,1,3)
plot(time_V,Vout_V,'c','linewidth',3)
ylabel('V (volts)')
ylim([198.9 200.3])
xlim([0.295 0.325])
legend('c. Mod 1 Voltage mode control')
grid on;

subplot(5,1,4)
plot(time_V_mod2,Vout_V_mod2,'g','linewidth',3)
ylabel('V (volts)')
ylim([198.9 200.3])
xlim([0.295 0.325])
legend('d. Mod 2 Voltage mode control')
grid on;

subplot(5,1,5)
plot(time_F,Vout_F,'r','linewidth',3)
ylim([199 200.3])
xlim([0.295 0.325])
grid on;
legend('e. Current mode control')
ylabel('V (volts)')
xlabel('Time (sec)')

% DCM-DCM 400->1600
figure(10)
subplot(5,1,1)
plot(time_K,Vout_K,'b','linewidth',3)
ylim([199.7 200.3])
xlim([0.448 0.466])
grid on;
ylabel('V (volts)')
legend('a. Mod 1 K-factor control')
%title(' Step response: 50\Omega to 185\Omega')

subplot(5,1,2)
plot(time_K_m2,Vout_K_m2,'m','linewidth',3)

```

```

ylim([199.7 200.3])
xlim([0.448 0.466])
grid on;
ylabel ('V (volts)')
legend('b. Mod 2 K-factor control')
%title(' Step responce: 50\Omega to 185\Omega')

subplot(5,1,3)
plot(time_V,Vout_V,'c','linewidth',3)
ylabel ('V (volts)')
ylim([199.7 200.3])
xlim([0.448 0.466])
legend('c. Mod 1 Voltage mode control')
grid on;

subplot(5,1,4)
plot(time_V_mod2,Vout_V_mod2,'g','linewidth',3)
ylabel ('V (volts)')
ylim([199.7 200.3])
xlim([0.448 0.466])
legend('d. Mod 2 Voltage mode control')
grid on;

subplot(5,1,5)
plot(time_F,Vout_F,'r','linewidth',3)
ylim([199.7 200.3])
xlim([0.448 0.466])
grid on;
legend('e. Current mode control')
ylabel ('V (volts)')
xlabel('Time (sec)')
% %print(gcf,'-dtiff','-r350','CCM to DCM');

```

F. BODE AND ROOT LOCUS PLOT

```

% Bode and root locus
bode(Gvd)
grid on
opts = bodeoptions('cstprefs')
opts.Title.String = '';
set(findall(gcf,'type','line'),'linewidth',3)

h = rlocusplot(Gvd);
p = getoptions(h);
p.Title.String = '';
setoptions(h,p);
set(findall(gcf,'type','line'),'linewidth',3)

```

G. SAVE SIMULINK VARIABLES

```

%save variables
%filename='PI_data.mat'

```

```

iL_F=iL;
Vout_F=Vout;
time_F=time;
save('F_data.mat','iL_F','Vout_F','time_F')

%save variables
%filename='PI_data.mat'
iL_K=iL;
Vout_K=Vout;
time_K=time;
save('K_data.mat','iL_K','Vout_K','time_K')

%save variables
%filename='PI_data.mat'
iL_K_m2=iL;
Vout_K_m2=Vout;
time_K_m2=time;
save('K_data_m2.mat','iL_K_m2','Vout_K_m2','time_K_m2')

%save variables
%filename='PI_data.mat'
iL_V_mod2=iL;
Vout_V_mod2=Vout;
time_V_mod2=time;
save('V_data_mod2.mat','iL_V_mod2','Vout_V_mod2','time_V_mod2')

%save variables
%filename='PI_data.mat'
iL_K_reg=iL;
Vout_K_reg=Vout;
time_K_reg=time;
save('K_data_reg.mat','iL_K_reg','Vout_K_reg','time_K_reg')

%save variables
%filename='PI_data.mat'
iL_V_reg=iL;
Vout_V_reg=Vout;
time_V_reg=time;
save('V_data_reg.mat','iL_V_reg','Vout_V_reg','time_V_reg')

%save variables
%filename='PI_data.mat'
iL_V=iL;
Vout_V=Vout;
time_V=time;
save('V_data.mat','iL_V','Vout_V','time_V')

```

H. SIMULINK EXPERIMENT INITIAL CONDITIONS

```

% Simulink verification initial conditions
clear all;
clc;

```

```

fa=550;
sw_freq=15000;
Vout_ref=20;
Vbatt=10;
Lboost=1114*10^-6;
Cap=2096*10^-6;
Rload1=101.45;
Rload2=Rload1;

tstep = 3e-8;
tstop=0.1

```

I. SIMULINK EXPERIMENT GAIN AND PHASE CALCULATION

```

%Simulink verification gain and phase calculation
sample=length(vo);
gain2=vo_amp(end)/vin_amp(end)
phase_sh=phaseshift(end)

```

J. SIMULINK EXPERIMENT PWM PLOT

```

%Simulink verification PWM plot

```

```

samples=length(time);
samples=round(samples/5);
cycles=round(1/sw_freq/tstep);
cycles2=round(1/fa/tstep);
time_new=time(samples:samples+4*cycles);
PWM_new=PWM(samples:samples+4*cycles);
vin_new=vin(samples:samples+4*cycles);
Vswt_new=Vswt(samples:samples+4*cycles);

```

```

figure (1)
plot(time_new,vin_new)
hold on;
plot(time_new,Vswt_new)
plot(time_new,PWM_new)
xlabel('Time (sec)')
ylabel('Amplitude (V)')
title('PWM with injected perturbations at 250Hz')
legend('Control signal','Sawtooth','PWM','location','southeast')
% ylim([0 1.1])
% xlim([0.0024 0.00265])
vo_new=vo(samples:samples+2*cycles2);
vo_new=vo_new-mean(vo_new);
vin_new2=vin(samples:samples+2*cycles2);
vin_new2=vin_new2-mean(vin_new2);
time_new2=time(samples:samples+2*cycles2);
figure (2)

```

```

plot(time_new2,5*vin_new2)
hold on;
plot(time_new2,vo_new)
xlabel('Time (sec)')
ylabel('Amplitude (V)')
title('Injected perturbubtions at 250Hz vs Output perturbubtions')
legend('Injected perturbubtions','Output perturbubtions','location','southeast')
% ylim([0 1.1])
% xlim([0.12 0.12025])

```

K. SIMULINK AND HARDWARE EXPERIMENT PLOTS

% This file has been created plot hardware and software verification
 %% Parameters

```

Rload=101.45;
sw_freq=15000;
fs=sw_freq;
Fc=fs
Ts=1/fs;
Vbattery=10;
Lboost=1114e-6;
Rbattery=0.355+0.5;
Cbus=2094e-6;
Vdc_ref=20;
Vload=Vdc_ref;
Vd=0.8;
tstep=6e-6;
%Ts=tstep;
%fs=1/tstep;
%% Steady State Values
% D=(Vdc_ref-Vbattery)/Vdc_ref
Dp=(Vbattery-Vd+sqrt((Vbattery-Vd)^2-4*Rbattery*(Vload^2)/Rload))/2/Vload
D=1-Dp
Rcrit_min=(27/2)*Lboost/Ts %Tabe 5-1
Rcrit=2*Lboost/D/Ts/(1-D)^2
%Iload_ideal=Vbattery/Rload/Dp^2
IL_dc=(Vbattery-Vd)/(Rbattery+Rload*Dp^2)

%% Small signal analysis
A=[-Rbattery/Lboost,-Dp/Lboost,Dp/Cbus,-1/Rload/Cbus];
Bv=[1/Lboost;0];
Bd=[Vload/Lboost;-IL_dc/Cbus]
C=[0,1];
Ed=0;
Ev=0;
Ci=[1,0];
%% Matrix form
[num1 den1]=ss2tf(A,Bd,C,Ed,1);
% [num2 den2]=ss2tf(A,Bv,C,Ev,1);
% [num3 den3]=ss2tf(A,Bd,Ci,Ev,1) ;%Gid
% [num4 den4]=ss2tf(A,Bv,Ci,Ev,1);

```

```

%% Tranfer Function
Gvd1=tf(num1,den1)

%% Lab Measurements Gain Plot
figure(1)
[mag,ph,w]=bode(Gvd1);
bode(Gvd1)
grid on;
mag=squeeze(mag);
w=squeeze(w);
ph=squeeze(ph);

%f_test=simulinktest(:,1); % Simulink data

f_test=hardtest1(:,1); %hardware data
a1=hardtest1(:,3);
b1=hardtest1(:,2)/10;
gain1=a1./b1;
a2=hardtest1(:,7);
b2=hardtest1(:,6)/10;
gain2=a2./b2;
a3=hardtest1(:,11);
b3=hardtest1(:,10)/10;
gain3=a3./b3;
gain_mean=(gain1+gain2+gain3)/3;

%gain_mean=simulinktest(:,3);
figure (2)
semilogx(w/2/pi,20*log10(mag),'linewidth',2)
hold on;
semilogx(f_test,20*log10(gain_mean),'o','linewidth',2)
grid on
xlabel('Frequency (Hz)')
ylabel('Gain (dB)')
xlim([1 560])
legend('Theoretical G_{vd}(s)=v(s)/d(s)', 'Measured G_{vd}(s)=v(s)/d(s)', 'Simulated G_{vd}(s)=v(s)/d(s)')
%title('Vin=10V, Vout=20V, Duty cycle=55.9%')
%legend('Theoretical G_{vd}(s)=v(s)/d(s)',
%legend('Simulated G_{vd}(s)=v(s)/d(s)')
%% Lab measurementents Frequency plot

ph1=hardtest1(:,4);
ph1=360-ph1;
ph2=hardtest1(:,8);
ph2=360-ph2;
ph3=hardtest1(:,12);
ph3=360-ph3;
ph_mean=(ph1+ph2+ph3)/3;

%ph_mean=360-simulinktest(:,4);
figure (3)
semilogx(w/2/pi,ph,'linewidth',2)

```

```

hold on;
semilogx(f_test,ph_mean,'o','linewidth',2)
ylim([120 360])
grid on
xlabel('Frequency (Hz)')
ylabel('Phase (degrees)')
xlim([1 560])
legend('Theoretical  $G_{vd}(s)=v(s)/d(s)$ ','Measured  $G_{vd}(s)=v(s)/d(s)$ ')
%legend('Theoretical  $G_{vd}(s)=v(s)/d(s)$ ','Simulated  $G_{vd}(s)=v(s)/d(s)$ ')
%title('Vin=10V, Vout=20V, Duty cycle=55.9%')

```

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