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LINEAR AND NON LINEAR CONTROLLER DESIGN FOR DC-DC CONVERTER

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CERTIFICATE OF APPROVAL

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DEDICATION

To

Our Beloved Parents

And

Our Whole Family

ABSTRACT

DC-DC power converters play an important role in powering telecom and computing systems. With the speed improvement and cost reduction of digital control, digital controller is becoming a trend for DC-DC converters in addition to existed digital monitoring and management technology. In this thesis, digital control is investigated for DC-DC converters applications.

To deeply understand the whole control systems, DC-DC converter models are investigated based on averaged state-space modeling. Considering Buck DC-DC converter, the thesis takes it as an example for digital control modeling and implementation.

In Chapter 2, unified steady-state DC models and small-signal models are developed for DC-DC buck converters. Based on the models, digital controller design is implemented in Chapter 3. In Chapter 4, Fuzzy logic control is introduced and based on that knowledge Fuzzy controller is designed for the control of buck converter. It should be noted that Fuzzy logic control provides superior performance than conventional PI and PID controller especially during under load transient condition.

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CHAPTER 1: INTRODUCTION TO DC-DC CONVERTER

1.1 DC-DC Converter:

A DC-to-DC converter is an electronic circuit which converts a source of direct current (DC) from one voltage level to another. It is a class of power converter. DC to DC converters are important in portable electronic devices such as cellular phones and laptop computers, which are supplied with power from batteries primarily. Such electronic devices often contain several sub-circuits, each with its own voltage level requirement different from that supplied by the battery or an external supply (sometimes higher or lower than the supply voltage). Additionally, the battery voltage declines as its stored power is drained. Switched DC to DC converters offer a method to increase voltage from a partially lowered battery voltage thereby saving space instead of using multiple batteries to accomplish the same thing.

Most DC to DC converters also regulate the output voltage. Some exceptions include high-efficiency LED power sources, which are a kind of DC to DC converter that regulates the current through the LEDs, and simple charge pumps which double or triple the input voltage.

The functions of dc-dc converters are:

1. to convert a dc input voltage VS into a dc output voltage VO;

2. to regulate the dc output voltage against load and line variations;

3. to reduce the ac voltage ripple on the dc output voltage below the required level;

4. to provide isolation between the input source and the load (isolation is not always required);

5. to protect the supplied system and the input source from electromagnetic interference (EMI); and

6. to satisfy various international and national safety standards.

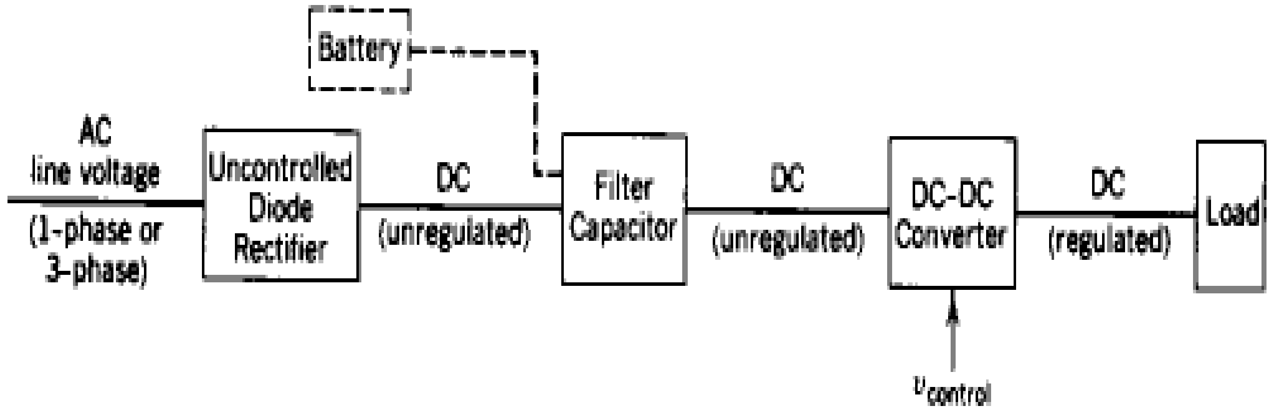

Fig 1.1: A DC-DC Converter System

1.2 Conversion methods:

a) Linear:

Linear regulators can only output at lower voltages from the input. They are very inefficient when the voltage drop is large and the current is high as they dissipate heat equal to the product of the output current and the voltage drop; consequently they are not normally used for large-drop high-current applications.

The inefficiency wastes power and requires higher-rated and consequently more expensive and larger components. The heat dissipated by high-power supplies is a problem in itself and it must be removed from the circuitry to prevent unacceptable temperature rises.

Linear regulators are practical if the current is low, the power dissipated being small, although it may still be a large fraction of the total power consumed. They are often used as part of a simple regulated power supply for higher currents: a transformer generates a voltage which, when rectified, is a little higher than that needed to bias the linear regulator. The linear regulator drops the excess voltage, reducing hum-generating ripple current and providing a constant output voltage independent of normal fluctuations of the unregulated input voltage from the transformer/bridge rectifier circuit and of the load current.

Linear regulators are inexpensive, reliable if good heat sinks are used and much simpler than switching regulators. As part of a power supply they may require a transformer, which is larger for a given power level than that required by a switch-mode power supply. Linear regulators can provide a very low-noise output voltage, and are very suitable for powering noise-sensitive low-power analog and radio frequency circuits. A popular design approach is to use an LDO, Low Drop-out Regulator, that provides a local "point of load" DC supply to a low power circuit.

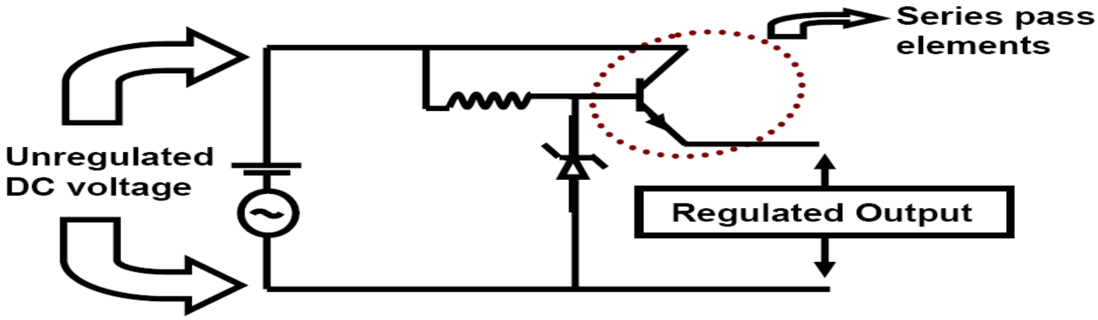

Fig 1.2: Schematic linear voltage regulator

b) Switched-mode conversion:

Electronic switch-mode DC to DC converters convert one DC voltage level to another, by storing the input energy temporarily and then releasing that energy to the output at a different voltage. The storage may be in either magnetic field storage components (inductors, transformers) or electric field storage components (capacitors). This conversion method is more power efficient (often 75% to 98%) than linear voltage regulation (which dissipates unwanted power as heat). This efficiency is beneficial to increasing the running time of battery operated devices. The efficiency has increased since the late 1980s due to the use of power FETs, which are able to switch at high frequency more efficiently than power bipolar transistors, which incur more switching losses and require a more complicated drive circuit. Another important innovation in DC-DC converters is the use of synchronous rectification replacing the flywheel diode with a power FET with low "on resistance", thereby reducing switching losses.

Most DC-to-DC converters are designed to move power in only one direction, from the input to the output. However, all switching regulator topologies can be made bi-directional by replacing all diodes with independently controlled active rectification. A bi-directional converter can move power in either direction, which is useful in applications requiring regenerative braking. Drawbacks of switching converters include complexity, electronic noise (EMI / RFI) and to some extent cost, although this has come down with advances in chip design.

DC-to-DC converters are now available as integrated circuits needing minimal additional components. They are also available as a complete hybrid circuit component, ready for use within an electronic assembly.

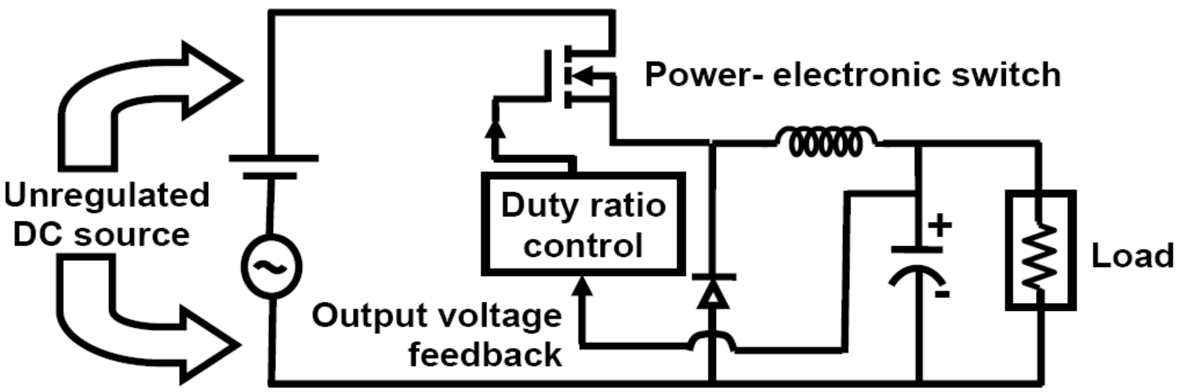

Fig 1.3: A schematic switched mode dc to dc chopper circuit

1.3 Types of Converters

Currently, dc/dc converters can be divided into two broad categories:

- Non-isolated dc/dc converters
- Isolated dc/dc converters

1.3.1 Non-Isolated DC/DC Converters

The non-isolated converter usually employs an inductor, and there is no dc voltage isolation between the input and the output. The vast majority of applications do not require dc isolation between input and output voltages. The non-isolated dc-dc converter has a dc path between its input and output. Battery-based systems that don't use the ac power line represent a major application for non-isolated dc-dc converters. Point-of-load dc-dc converters that draw input power from an isolated dc-dc converter, such as a bus converter, represent another widely used non-isolated application. Most of these dc-dc converter ICs use either an internal or external synchronous rectifier. Their only magnetic component is usually an output inductor and thus less susceptible to generating electromagnetic interference. For the same power and voltage levels, it usually has lower cost and fewer components while requiring less pc-board area than an isolated dc-dc converter. For lower voltages (12V) non-isolated buck converters can be used.

1.3.2 Isolated DC/DC Converters

For safety considerations, there must be isolation between an electronic system's ac input and dc output. Isolation requirements cover all systems operating from the ac power line, which can include an isolated front-end ac-dc power supply followed by an isolated "brick" dc-dc converter, followed by a non-isolated point-of-load converter. Typical isolation voltages for ac-dc and dc-dc power supplies run from 1500 to 4000V, depending on the application. An isolated converter employs a transformer to provide dc isolation between the input and output voltage which eliminates the dc path between the two. Isolated dc-dc converters use a switching transformer whose secondary is either diode-or synchronous-rectified to produce a dc output voltage using an inductorcapacitor output filter. This configuration has the advantage of producing multiple output voltages by adding secondary transformer windings. For higher input voltages (48V) transformer isolated converters are more viable [1].

1.4 Types of Switched Mode Power Supplies (SMPS):

There are four basic topologies of switching regulators:

1. **Buck regulator**
2. **Boost regulator**
3. **Buck-boost regulator**
4. **Cuk regulator**

1.5 PROBLEM STATEMENT

The output voltage (Vo) of buck alone usually is unstable. So criteria must concern is rise time, overshoot , settling time and steady state error , to get the desired output and to reduce the undesired output.

Problem statement:

· steady state error

- The output of buck alone is not reaching the desire value meaning it has error.

· rise time

- The rise time is too long

· settling time

- The output oscillating too long, it takes time to reach the stable state.

· Overshoot

-The over shoot is high.

CHAPTER 2: Buck Converter

2. BUCK CONVERTER

In this section we summarized the brief introduction of buck converter and purpose why we choose the buck converter as well the circuit topology and the brief explanation of the components used in the construction of buck converter.

Introduction

There has been an incredible development in the field of electrical components in recent years. The competition is to make things portable and flexible so that the usage will be more with less effort. As stated for electrical components to run, the power consumption is the major factor. For the optimum usage of electronic components, dc to dc converter plays a major role. The dc to dc converter can be used for many electronic components and it is widely used in telephone components and many other electronic devices. The purpose of dc to dc converter is to convert (i.e. to step down) the voltage from one value to the other and to perform regulation for the electronic circuit.

2.1 Why Buck Converter?

In general the simplest way to reduce the voltage of a DC supply is by using linear regulators. Consider the linear regulator as shown in Figure 2.1. Here, the source voltage is Vs which is to be step down to voltage VL across the resistor R1 which means the voltage across RL must be dropped which intern results in waste of power in the form of heat [2]. This problem can be overcome by using Buck Converter as it uses switch (Diode) to operate in ON and OFF states.

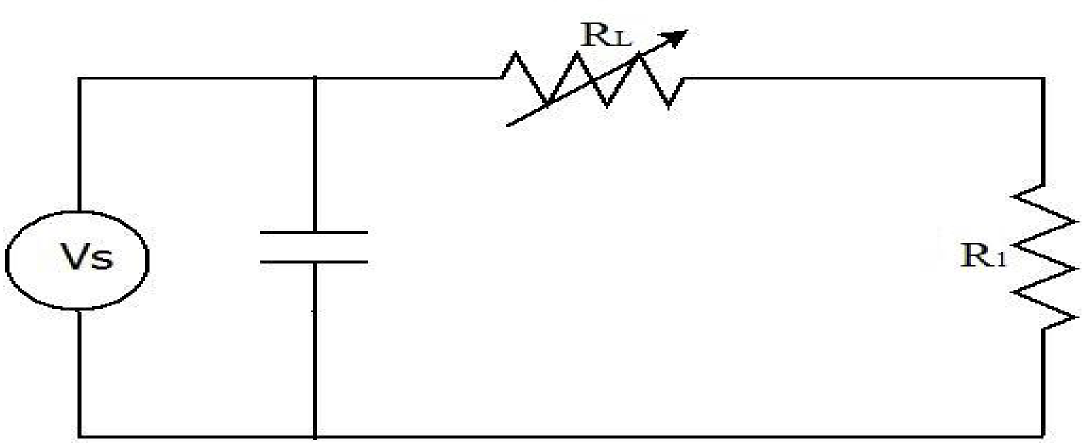

Fig 2.1: Circuit diagram of a Linear Regulator

The dc-dc buck converter topology is most widely used power management and microprocessor voltage-regulator applications. These applications require high frequency and transient response over a wide load current range. They can convert high voltage into low regulated voltage. Buck converter can be used in computers, where we need voltage to be stepped down. Buck converter provides long battery life for mobile phones which spend most of the time in stand-by state [3] .

When the switch is ON the inductor gets charged to its maximum level, because of its flexibility of ON and OFF states it can be switched to OFF state when inductor charges to its maximum capacity. With this feature the usage of heat sinks and cooling agents can be avoided. Hence, because of its advantage we opt for buck converter rather than a linear regulator.

2.2 Buck Converter Circuit topology

The name "Buck Converter" itself indicates that the input voltage is bucked or attenuated and low voltage appears at the output. A buck converter or step down voltage regulator provides non isolated, switch mode dc-dc conversion with the advantage of simplicity and low cost [3]. Figure 2.2, shows a simplified dc-dc buck converter that accepts a dc input and uses pulse width modulation of switching frequency to control the output voltage. The buck converter consists of Source Voltage 'V_s ', Diode, Inductor 'L', Inductor Resistance 'RL', Capacitor 'C', and Capacitive Resistance 'Rc' all connected to a Load.

Switch mode power supply is generally used to provide the output voltage which is less than the input voltage to the load from an intermediate DC input voltage bus or a battery source. A simplified buck converter point of load which has power supply from a switch mode buck converter is shown in Figure.2.3. The buck converter consists of main power switch, a diode, a low-pass filter (L and C) and a load [4]. The basic buck converter operates in ON and OFF states. In ON state i.e. when the switch is closed the current to load is supplied from source voltage through inductor, where inductor gets charged to its peak level. Where as in OFF state i.e. when switch is open the inductor acts as source to the load.

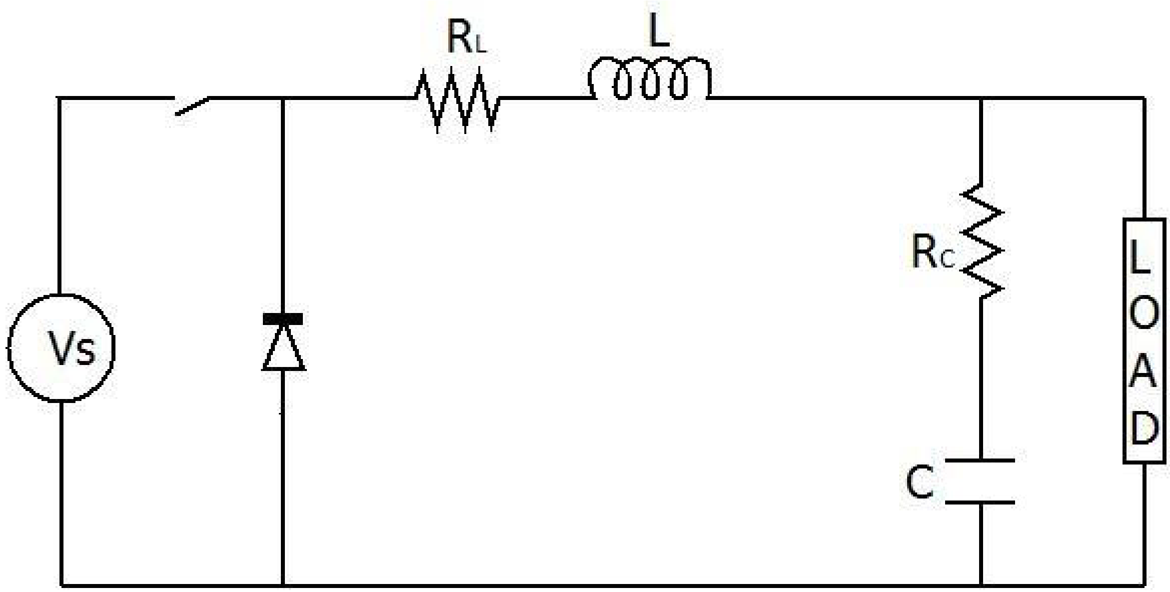

Figure 2.2. Buck Converter

2.3 Circuit components explanation:

2.3.1 Switch

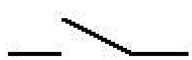

Figure 2.3. Switch

Consider a switch as shown in Figure 2.3. We use transistor as a switch in buck converter, the input to the transistor is a pulse width modulated (PWM) signal which is used to turn ON or turn OFF the transistor. When the switch is turned ON the input voltage equals the load voltage and the voltage across the inductor, when the switch is turned OFF the load voltage equals the voltage across the inductor. The average output voltage can be controlled by varying the PWM signal [5].

2.3.2 Inductor

Figure 2.4. Inductor

Consider an inductor as shown in Figure 2.4. An inductor supplies constant power to the load resistor when the switch is turned OFF. It helps to maintain a continuous current across the load resistor when there is no supply voltage. It also controls sudden changes in the current when the switch is ON [5].

2.3.3 Capacitor

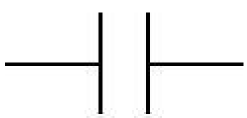

Figure 2.5. Capacitor

Consider a capacitor as shown in Figure 2.5. This acts as a low pass filter and removes harmonics in the output. It must be chosen large enough in accordance to control the voltage changes, overshoots, ripples during the time when the switch is changing on and off [5].

2.3.4 Resistor

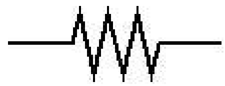

Figure 2.6. Resistor

Consider a resistor as shown in Figure 2.6. A resistor is a component of an electrical circuit which helps to oppose the flow of electrons into the component. The flow of current through the resistor is inversely proportional to the value of the resistance [5].

2.3.5 Diode

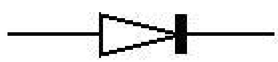

Figure 2.7. Diode

Consider a diode as shown in Figure 2.7. Diode is an electrical component which has two states of operation i.e. ON and OFF state. The ON and OFF states depends on the direction of flow of current through it. When the current flows from positive to negative terminal the diode acts as short circuit and allows the flow of current through it which is stated as ON state. When the current flows from negative to positive terminal the diode acts as open circuit and opposes the flow of current through it which is stated as OFF state [5].

2.4 Two States of operation of Buck Converter

2.4.1 On State

Figure 2.8, shows the buck converter operating in on state. In this state of operation the switch will be in closed state so that Vs will be the source voltage for the inductor. Obviously in this state the current through the diode flows from negative terminal to positive terminal which causes diode to be inactive. Hence there will be no backward current to the inductor.

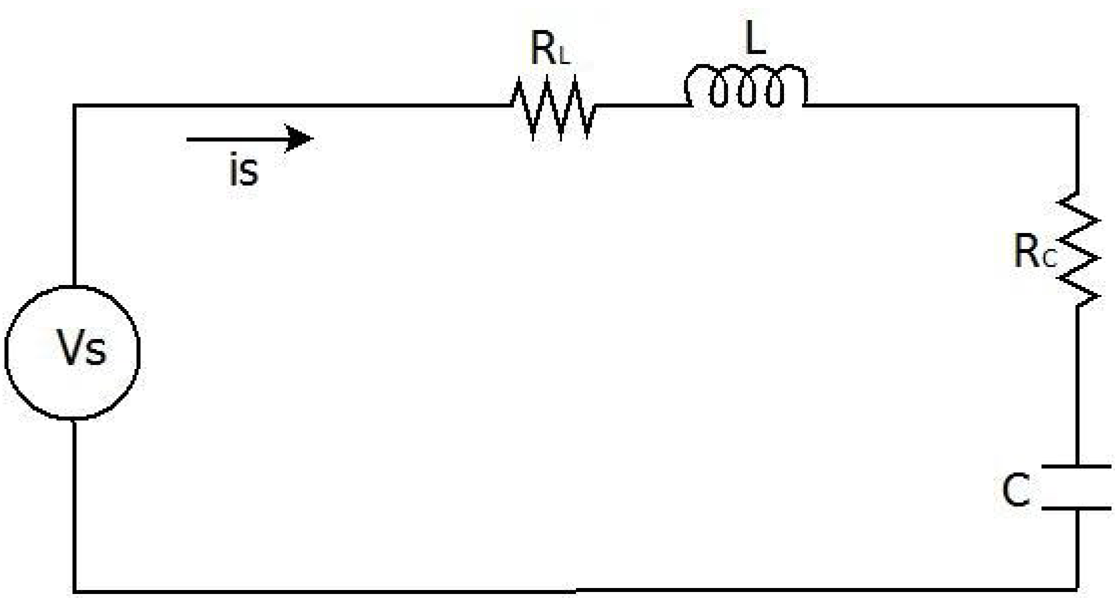

Figure 2.8. On State of Buck Converter

2.4.2 Off State

Figure 2.9, shows the buck converter operating in off state. In this state of operation the switch will be in open state so that there will be no path to current to flow from source voltage Vs to inductor. In this state inductor starts discharging, which cause current in diode to flow from positive to negative terminals. Hence there will be a backward current to the inductor.

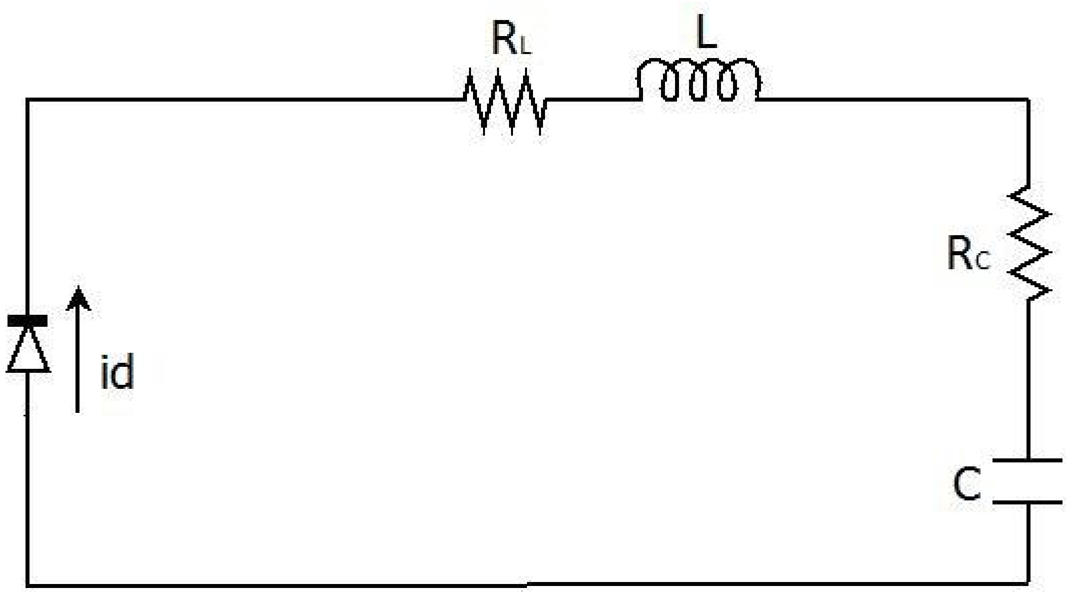

Figure 2.9. Off State of Buck Converter

Figure 2.10, shows the buck converter wave forms i.e. 'VL' shows the voltage across the inductor, 'iS' shows the switch modes during the time T and 'iL' shows the current flow during on and off states.

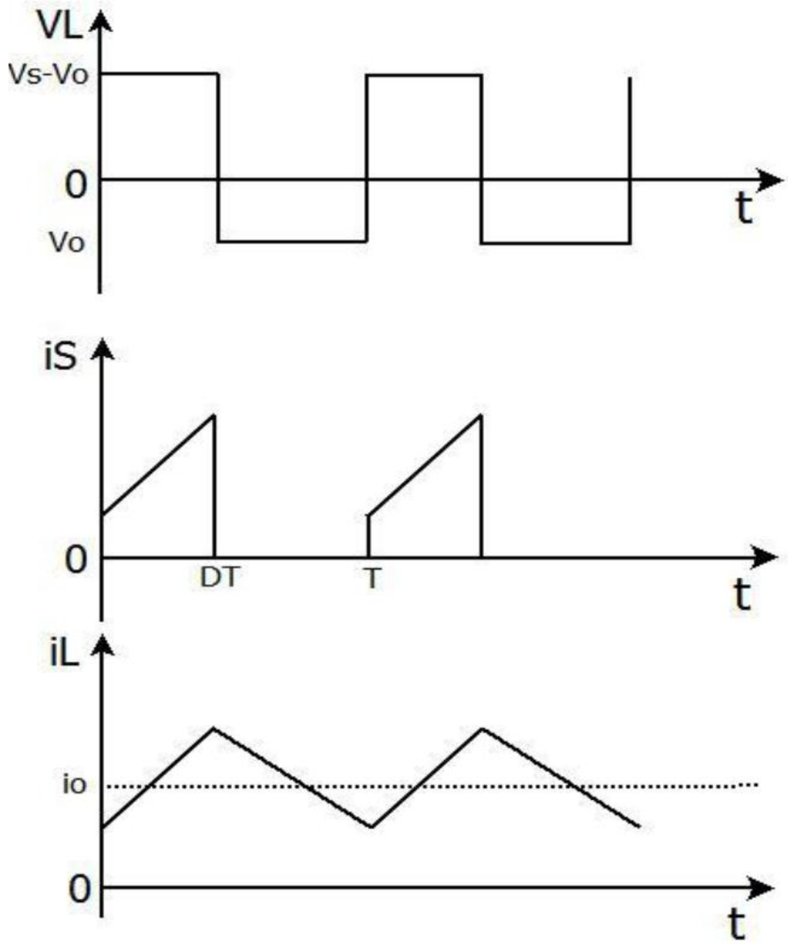

Figure 2.10. Buck converter wave forms

The relationship between input voltage, output voltage and the switch duty cycle 'D' can be derived from VL waveform. According to Faraday's law, the inductor volt second product over a period of steady state operation is zero [6].

For the buck converter:

$$(V_s - V_o)DT = -V_o(1 - D)T$$

$$\begin{aligned} \text{Where } & V_s \text{ : Source Voltage,} \\ & V_o \text{ : Output Vlotage,} \\ & T \text{ : Time period,} \\ \text{And } & D \text{ : Duty cycle.} \end{aligned}$$

Hence the dc voltage transfer function can be defined as the ratio of the output voltage to the input voltage,

$$D = \frac{V_o}{V_s}$$

2.5 Modes of operation

Buck converter can operate in two modes of operation, Continuous mode and Discontinuous mode. In continuous mode, current at inductor never falls to zero. Where as in discontinuous mode at one point of time the current in inductor falls to zero due to consumption of energy by the load.

2.6 Calculation for Duty Ratio

For calculation of the duty ratio we will first of all assume that the converter is in steady state.The switches are treated as being ideal, and the losses in the inductive and the capacitive elements are neglected. Also it is important to point out that the following analysis does not include any parasitic resistances (all ideal case). The analysis also has the assumption that the converter is operating in Continuous conduction mode only i.e. i (t) > 0 L .When the switch is on for time duration on t , the switch conducts the inductor current and the diode becomes reverse biased. This results in a positive voltage VL= Vd-Vo across the inductor in Figure 2.11(a). This voltage causes a linear increase in the inductor current iL. When the switch is turned off, because of the inductive energy storage, iL continues to flow. This current now flows through the diode, and VL = −Vo in Figure 2.11 (b).

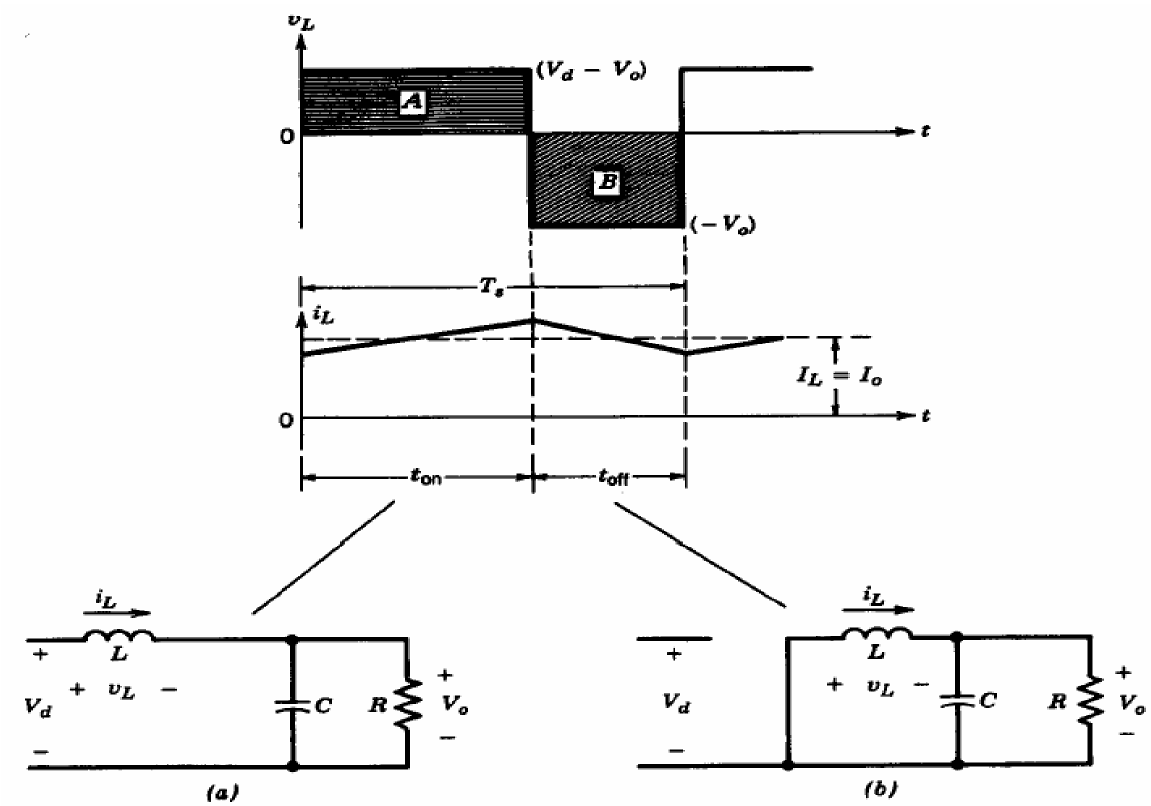

Figure 2.11: Step-down converter circuit states (assuming iL flows continuously): (a) switch on; (b) switch off [1]

Since in steady-state operation waveform must repeat from one time period to the next, the integral of the inductor voltage VL over one time period must be zero, where Ts = ton + toff:

$$\int_0^{Ts} VL\,\mathrm{dt}=\int_0^{ton} VL\,dt+\int_0^{Ts} VL\,dt=0 \qquad (2.3)$$

From Figure 2.11, it implies that areas A and B must be equal.Therefore,

$$(V_d-V_0)t_{on}=V_0(T_s-t_{on}) \qquad (2.4)$$

Or

$$V_0/V_d=t_{on}/T_s = D \text{ (duty ratio)} \qquad (2.5)$$

Hence in this mode, the voltage output varies linearly with the duty ratio of the switch for a given input voltage and does not depend on any other circuit parameter.

2.7 Calculation for Inductor

From Figure 2.11(a) we can derive a simplified differential equation based on the assumption that the voltage across the load, and thereby across the capacitor, is fairly constant. The differential equation in terms of the current through the inductor, when the switch is closed, may now be written as

$$L di_L(t)/dt= V_d-V0 \qquad (2.6)$$

Assuming that the circuit has assumed steady state hence there may already be some current in the inductor, I_{Lmin}, just prior to the closing of switch S. Hence for a time interval $0 \le t \le Ton = DT$, gives:

$$i_L(t)=(V_d-V_0)t/L + I_{L,min} \qquad (2.7)$$

The inductor current increases linearly with time and attains its maximum value $I_{L,max}$ as $t \rightarrow Ton = DT$ such that

$$I_{L,\max} = \frac{V_d - V_o}{L} DT + I_{L,\min} \qquad (2.8)$$

Defining the change in the current from its minimum to maximum value as the peak-to-peak current ripple ΔI_L, the equation 2.8 yields an expression for ΔI_L ,as

$$\Delta I_L = I_{L,\max} - I_{L,\min} = \frac{V_d - V_o}{L} DT \tag{2.9}$$

Note that the current ripple is directly proportional to D, the duty cycle, upon which we may not have any control because of the output voltage requirement. However, it is inversely proportional to the inductance L upon which we can exert some control. Thus, the current ripple can be controlled by a proper selection of the inductor. Let us now analyze the circuit when the switch is in its open position. The inductor current completes its path through the lower side MOSFET and the corresponding differential equation, for OFF $0 \le t \le T_{off}$, is

$$L\frac{di_L(t)}{dt} = -V_o \tag{2.10}$$

From the solution of the above first-order differential equation, we obtain

$$i_L(t) = -\frac{V_o}{L}t + I_{L,\max} \tag{2.11}$$

Where $I_{L,max}$ is the maximum value of the current in the inductor at the opening of the switch or the beginning of the off period. As $t \rightarrow T_{off} = (1-D)T$, the inductor current decreases to its minimum value $I_{L,min}$ such that

$$I_{L,\min} = -\frac{V_o}{L}(1-D)T + I_{L,\max}$$

(2.12)

The Eq.2.12 yields another expression for the peak-to-peak current ripple as

$$\Delta I_L = I_{L,\max} - I_{L,\min} = \frac{V_o}{L}(1-D)T$$

(2.13)

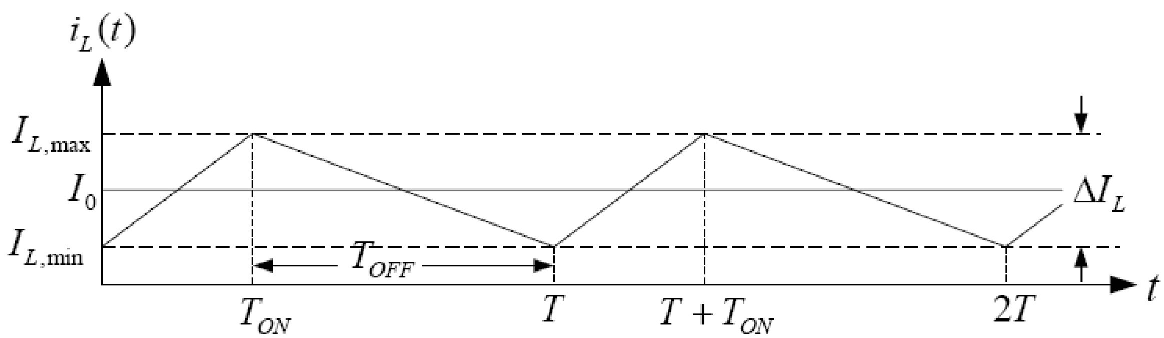

Figure 2.12 : Inductor current

The current through the inductor as given by Eq. 2.7 during the on time and by Eq. 2.11 during the off time is sketched in the Figure 2.12. The average current in the inductor must be equal to the dc current through the load. That is,

$$I_{L,avg} = I_0 = V_0/R \qquad (2.14)$$

The expressions for the maximum and minimum currents through the inductor may now be written as

$$I_{L,\max} = I_{L,avg} + \frac{\Delta I_L}{2} = \frac{V_o}{R} + \frac{V_o}{2L}(1-D)T$$

$$I_{L,\min} = I_{L,avg} - \frac{\Delta I_L}{2} = \frac{V_o}{R} - \frac{V_o}{2L}(1-D)T$$

(2.15)

(2.16)

The current supplied by the source varies from $I_{L,min}$ to $I_{L,max}$ during the time the switch is closed and is zero otherwise as shown in Figure 2.13.

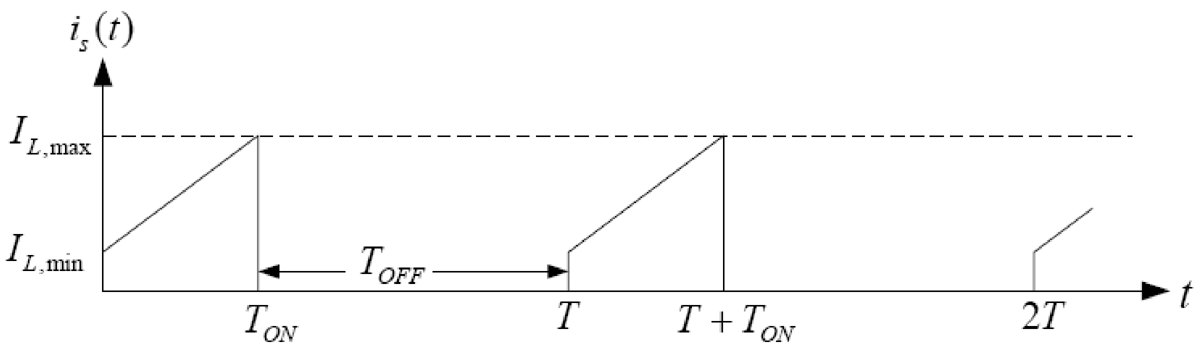

Figure 2.13 The source current

When the switch, the inductor, and the capacitor are treated as ideal elements, the average power dissipated by them is zero. Consequently, the average power supplied by the source must be equal to the average power delivered to the load. That is,

$$V_d I_d = V_o I_o = D V_S I_o \tag{2.17}$$

This equation helps us express the average source current in terms of the average load current as

$$Is= DI0 \tag{2.18}$$

The current through the lower side MOSFET is shown in Figure 2.14. Its average value is

ILS= (1-D)I0 (2.19)

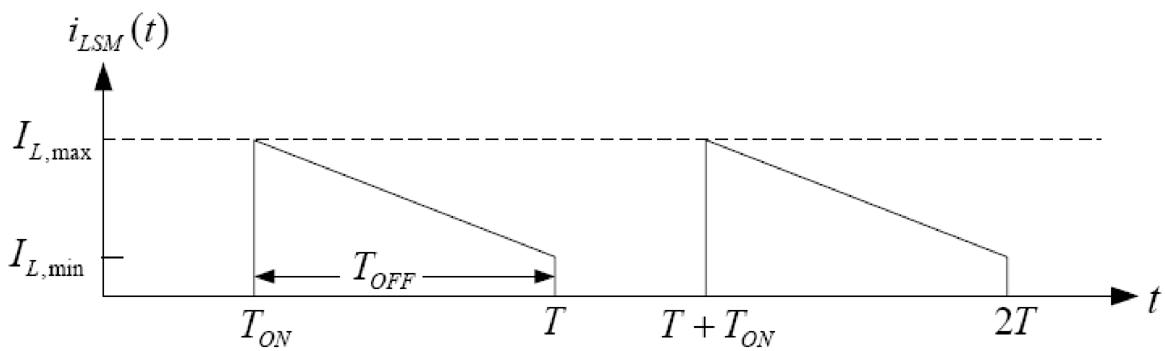

Figure 2.14 : current through the low side MOSFET

We know the fact that the buck converter can either operate in its continuous conduction mode or discontinuous mode. When it operates in the continuous conduction mode, there is always a current in the inductor. The minimum current in the continuous conduction mode can be zero. Consequently, there is a minimum value of the inductor that ensures its continuous conduction mode. It can be obtained from Eq. 2.16 by setting $I_{L,min}$ to zero as

$$\frac{V_o}{R} - \frac{V_o}{2L_{\min}}(1-D)T = 0 \qquad (2.20)$$

$$L_{\min} = \frac{1-D}{2}RT = \frac{1-D}{2f}R \qquad (2.21)$$

2.8 Calculation for Capacitor

The output capacitor is assumed to be so large as to yield $v_0(t)= V_0$. However, the ripple in the output voltage with a practical value of capacitance can be calculated by considering the waveforms shown in Figure 2.15 for a continuous conduction mode of operation. Assuming that all of the ripple component in iL flows through the capacitor and its average component flows through the load resistor, the shaded area in Figure2.15 represents an additional charge ΔQ.

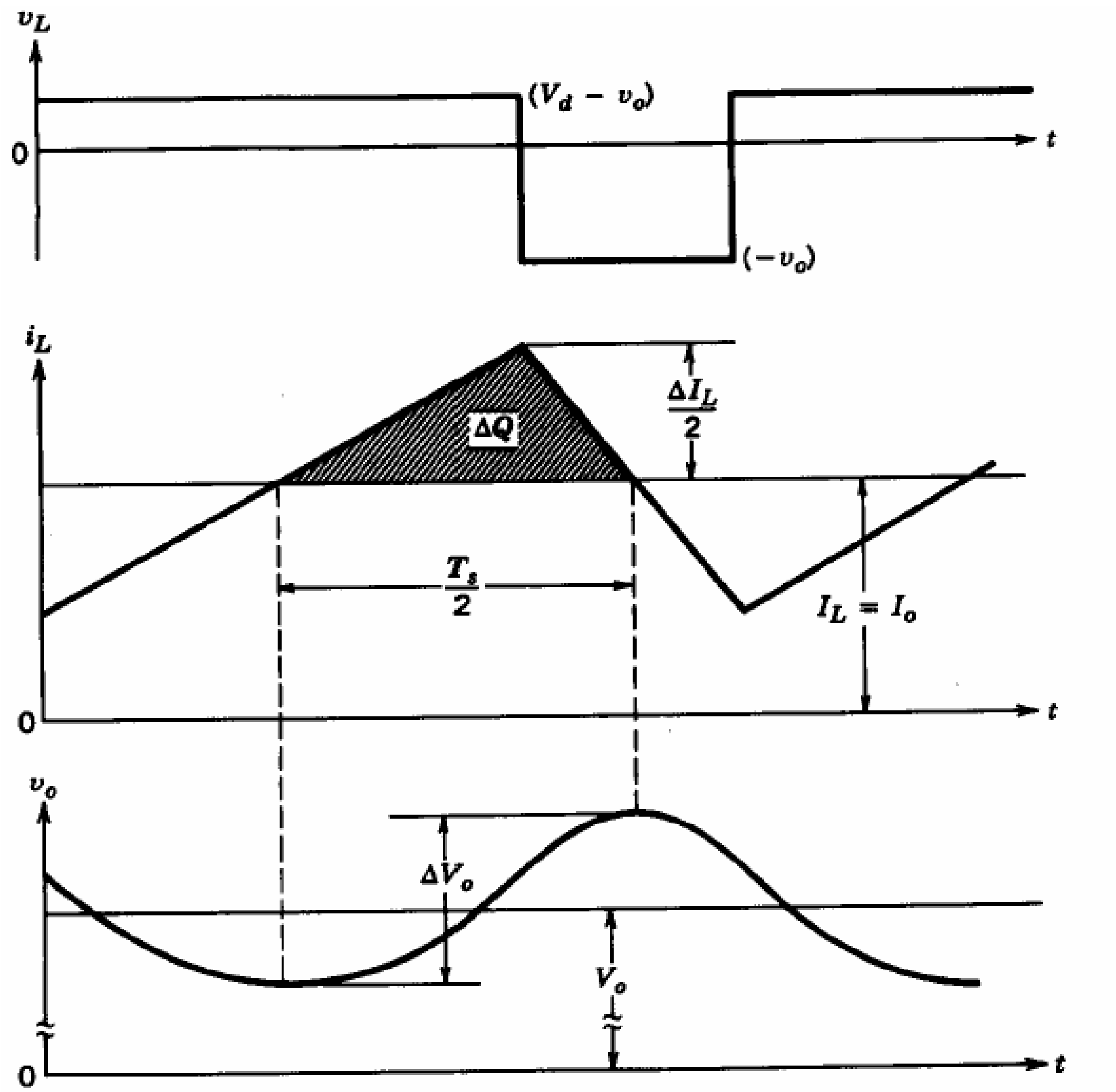

Figure 2.15: Output Voltage ripple in a step - down converter

Therefore, the peak-to-peak voltage ripple o ΔV can be written as

$$\Delta V_o = \frac{\Delta Q}{C} = \frac{1}{C}\frac{1}{2}\frac{\Delta I_L}{2}\frac{T_s}{2} \qquad (2.22)$$

From Figure 2.15 during t_{off}

$$\Delta I_L = \frac{V_o}{L}(1-D)T_s \tag{2.23}$$

Therefore, substituting ΔI_L from Eq. 2.23 into the Eq. 2.22 gives

$$\Delta V_o = \frac{T_s}{8C}\frac{V_o}{L}(1-D)T_s \tag{2.24}$$

$$\therefore \frac{\Delta V_o}{V_o} = \frac{1}{8}\frac{T_s^2(1-D)}{LC} = \frac{\pi^2}{2}(1-D)\left(\frac{f_c}{f_s}\right)^2 \tag{2.25}$$

Where switching frequency $f_s=1/T_s$ and

$$f_c = \frac{1}{2\pi\sqrt{LC}} \tag{2.26}$$

Equation 2-25 shows that the voltage ripple can be minimized by selecting a corner frequency fc of the low pass filter at the output such that fc <<fs. Also, the ripple is independent of the output load power, so long as the converter operates in the continuous-conduction mode. We should note that in switch-mode dc power supplies, the percentage ripple in the output voltage is usually specified to be less than, for instance, 1%. The analysis carried out above assumes ideal components and if we were to make the analysis using all the non-ideal components it would make the derivation a bit more complex with a lot of other parameters included in the final equation. But for the calculation of initial values of the components the above approximations does result in reasonable values. It is also important to realize here that the ESR

and ESL are also important and can even dominate. More about how the non-ideality can affect the overall system can be found on [7].

2.10 PWM Controller

The heart of a switching power supply is its switch control circuit (controller). One of the key objectives in designing a controller for the power converter is to obtain tight output voltage regulation under different line and load conditions [8]. Often, the control circuit is a negative-feedback control loop connected to the switch through a comparator and a Pulse Width Modulator (PWM). The switch control signal (PWM), controls the state (on or off) of the switch. This control circuit regulates the output voltage against changes in the load and the input voltage.

2.10.1 PWM

PWM is the method of choice to control modern power electronics circuits. The basic idea is to control the duty cycle of a switch such that a load sees a controllable average voltage. To achieve this, the switching frequency (repetition frequency for the PWM signal) is chosen high enough that the load cannot follow the individual switching events and they appear just a "blur" to the load, which reacts only to the average state of the switch.

With pulse-width modulation control, the regulation of output voltage is achieved by varying the duty cycle of the switch, keeping the frequency of operation constant. Duty cycle refers to the ratio of the period for which the power semiconductor is kept ON to the cycle period. A clearer understanding can be acquired by the Figure 2.16.

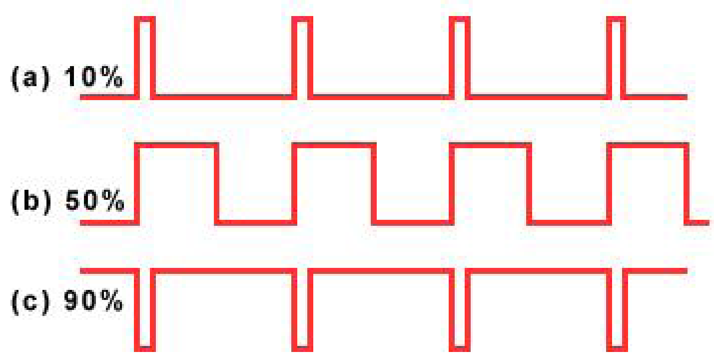

Figure 2.16: PWM Signal

The Figure 2.16 shows PWM signals for 10% (a), 50% (b), and 90% (c) duty cycles. Usually control by PWM is the preferred method since constant frequency operation leads to optimization of LC filter and the ripple content in output voltage can be controlled within the set limits.

2.10.2 Comparator and Voltage to PWM Converter

Switching power supplies rely on negative feedback to maintain the output voltages at their specified value. To accomplish this, a differential amplifier is used to sense the difference between an ideal voltage (the reference voltage) and the actual output voltage to establish a small error signal ($v_{control}$).

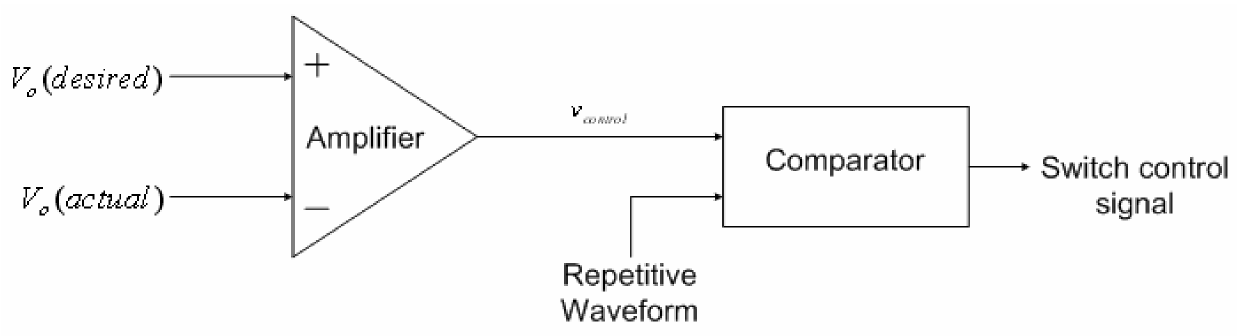

Figure 2.17: Voltage Reference Comparator [9]

The PWM switching at a constant switching frequency is generated by comparing a signal-level control voltage $v_{control}$ with a repetitive waveform as shown in Figure 2.17.

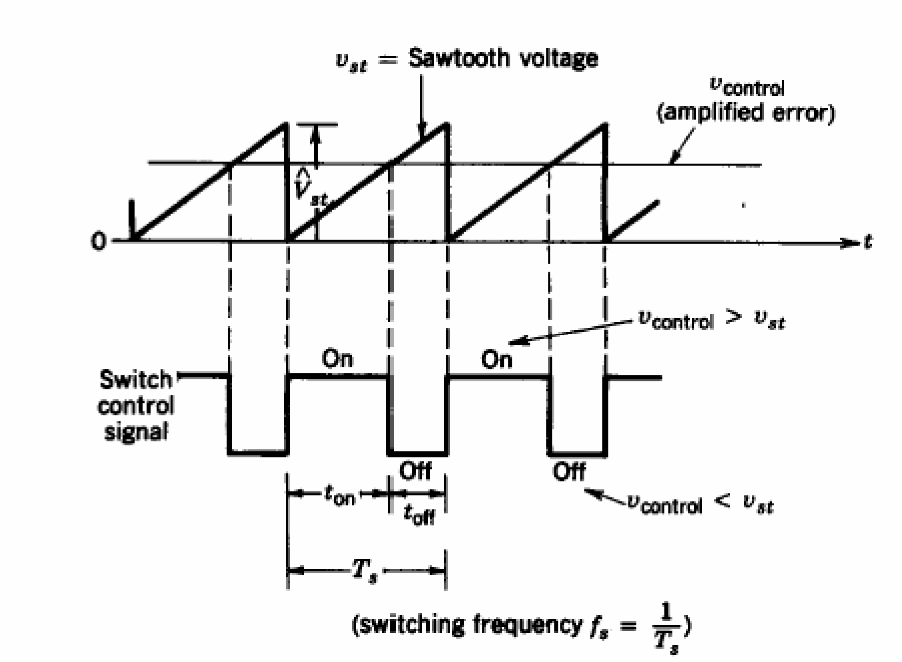

Figure 2.18: PWM Comparator Signals [9]

The frequency of the repetitive waveform with a constant peak, which is shown to be a sawtooth, establishes the switching frequency. This frequency is kept constant in a PWM control and is chosen to be in a few hundred kilohertz range. When the amplified error signal, which varies very slowly with time relative to the switching frequency, is greater than the sawtooth waveform, the switch control signal becomes HIGH, causing the switch to turn on. Otherwise, the switch is off. So when the circuit output voltage changes, $v_{control}$ also changes causing the comparator threshold to change. Consequently, the output pulse width also changes. This duty cycle change then moves the output voltage to reduce to error signal to zero, thus completing the control loop. In terms of $v_{control}$ and the peak of the sawtooth waveform V_{st} in Figure 2.18, the switch duty ratio can be expressed as

$$D = \frac{t_{on}}{T_s} = \frac{v_{control}}{\hat{V}_{st}} \qquad (2.27)$$

2.11 Feedback Control System

As has been mentioned earlier as well that the output voltages of dc power supplies are regulated to be within a specified tolerance band (e.g., ±1% around its nominal value) in response to changes in the output load and the input voltage lines. This process is accomplished by employing a negative feedback system which can be seen in Figure 2.19.

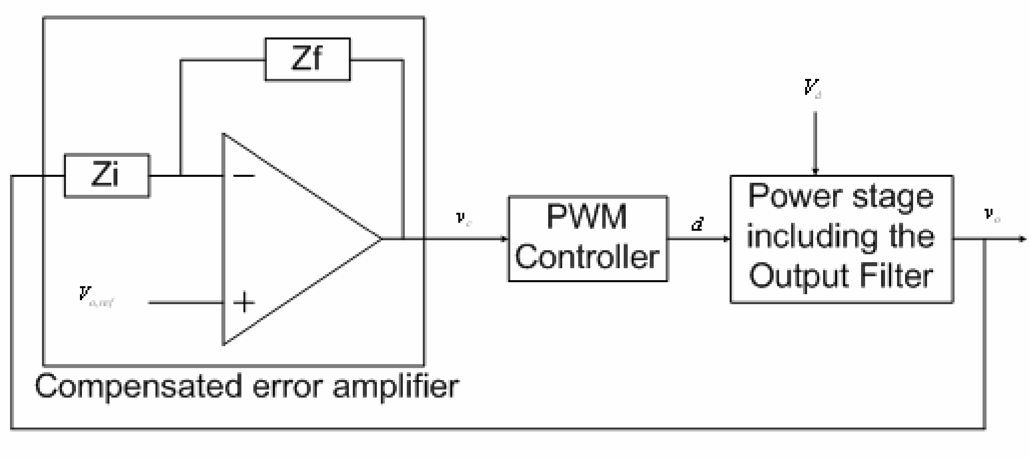

Figure 2.19: Feedback Control System [9]

The Power stage of the switch converter is not linearized. Since nonlinear systems are not equal to the sum of their parts, they are often difficult (or impossible) to model, and their behaviour with respect to a given variable (for example, time) is extremely difficult to predict. When modelling non-linear systems, therefore, it is common to approximate them as linear, where possible. With the Linear model, it will make possible certain mathematical assumptions and approximations, allowing for simple computation of results. In nonlinear systems these assumptions cannot be made. If the power stage of the switch-mode converter in Figure 2.19 can be linearized, then the Nyquist stability criterion and the Bode plots can be used to determine the appropriate compensation in the feedback loop for the desired steady-state and transient response. Each block in Figure 2.19 can be linearized around a steady-state operating point as using the state-space averaging technique [10] which allowed the theoretical prediction of a converters frequency response, and therefore a better understanding of a switched-mode regulator's feedback loop and stability criteria. This technique was developed by R. D. Middlebrook at Power Electronics Group - California Institute of Technology, USA. Therefore, each block in Figure 2.19 can be represented by a transfer function as shown in Figure 2.20, where the small ac signals are represented by "~."

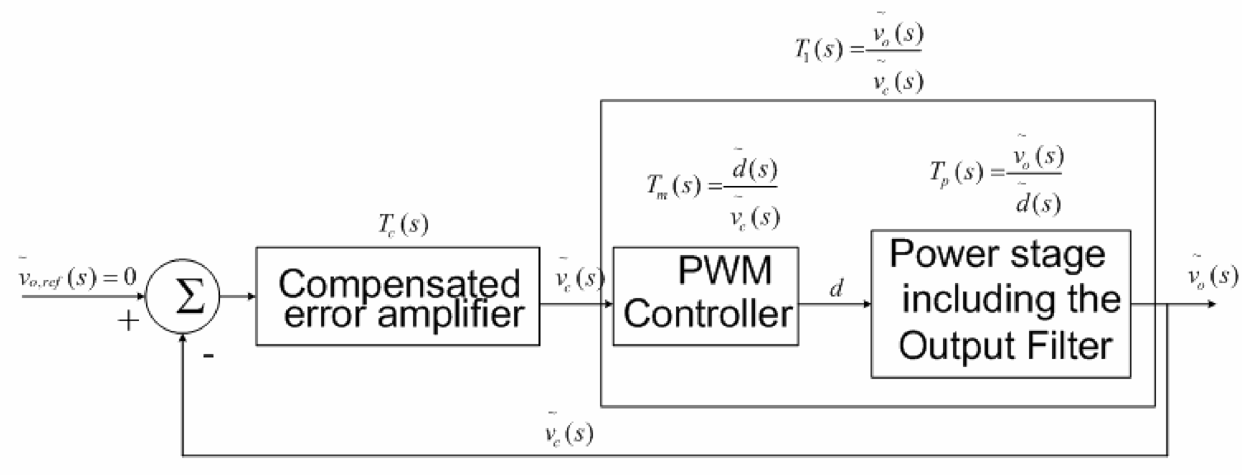

Figure 2.20: Linearized Feedback Control System [9]

2.12 Linearization using State-Space Averaging

The goal of the following analysis is to obtain a small signal transfer function $\tilde{v}(s)/\tilde{d}(s)$, Where $\tilde{v}o$ and $\tilde{d}$ are small perturbations in the output voltage vo and the switch duty ratio d, respectively, around their steady-state dc operating values Vo and D [9].

2.12.1 Power Stage & Output Filter

Step 1 State-Variable Description for Each Circuit State. In a converter operating in a continuous-conduction mode, there are two circuit states: one state corresponds to when the switch is on and the other to when the switch is off. A third circuit state exists during the discontinuous interval, which is not considered in the following analysis because of the assumption of a continuous conduction mode of operation.

During each circuit state, the linear circuit is described by means of the statevariable vector x consisting of the inductor current and the capacitor voltage. In the circuit description, the parasitic elements such as the resistance of the filter inductor and the equivalent series resistance (ESR) of the filter capacitor should also be included. Here Vd is the input voltage. A lowercase letter is used to represent a variable, which includes its steady-state dc value plus a small ac perturbation, for example, v0= V0 + $\tilde{v}0$. Therefore, during each circuit state, we can write the following state equations:

$$\dot{x} = A1x + B1\ vd \qquad \text{during d.Ts} \qquad (2.28)$$

And

$$\dot{x} = A2x + B2\ vd \qquad \text{during (1-d).Ts} \qquad (2.29)$$

where A1 and A2 are state matrices and B1 and B2 are vectors. The output v0 in all converters can be described in terms of their state variables alone as

$$v_o = C_1 x \qquad \text{during d.Ts} \qquad (2.30)$$

$$v_o = C_2\, x \quad \text{during (1-d).Ts} \quad (2.31)$$

where C1 and C2 are transposed vectors.

Step 2 Averaging the State-Variable Description Using the Duty Ratio d.

To produce an average description of the circuit over a switching period, the equations corresponding to the two foregoing states are time weighted and averaged, resulting in the following equations:

$$\dot{\mathbf{x}} = \left[\mathbf{A}_1 d + \mathbf{A}_2 (1-d)\right]\mathbf{x} + \left[\mathbf{B}_1 d + \mathbf{B}_2 (1-d)\right] v_d \qquad (2.32)$$

and

$$v_o = \left[\mathbf{C}_1 d + \mathbf{C}_2 (1-d)\right]\mathbf{x} \qquad (2.33)$$

Step 3: Introducing Small ac Perturbations and Separation into ac and dc Components.

Small ac perturbations, represented by "~", are introduced in the dc steady-state quantities (which are represented by the upper case letters). Therefore,

$$\mathbf{x} = \mathbf{X} + \tilde{\mathbf{x}} \qquad (2.34)$$

$$v_o = V_0 + \tilde{v}_o \qquad (2.35)$$

And

$$d = D + \tilde{d} \qquad (2.36)$$

In general, $v_d = V_d + \tilde{v}_o$. However, in view of our goal to obtain the transfer function between voltage $\tilde{v}_0$ and the duty ratio $\tilde{d}$ the perturbation $\tilde{v}_d$ is assumed to be zero in the input voltage to simplify our analysis. Therefore

$$v_d = V_d \qquad (2.37)$$

Using Eq. 2.34 through 2-37 in Eq 2-32 and recognizing that in steady state, $\dot{X}$=0

$$\dot{\tilde{\mathbf{x}}} = \mathbf{A}\mathbf{X} + \mathbf{B}V_d + \mathbf{A}\tilde{\mathbf{x}} + \left[(\mathbf{A}_1 - \mathbf{A}_2)\mathbf{X} + (\mathbf{B}_1 - \mathbf{B}_2)V_d\right]\tilde{d} + \text{terms containing products of } \tilde{\mathbf{x}} \text{ and } \tilde{d} \text{ (to be neglected)} \qquad (2.38)$$

Where

$$\mathbf{A} = \mathbf{A}_1 D + \mathbf{A}_2 (1 - D) \qquad (2.39)$$

And

$$\mathbf{B} = \mathbf{B}_1 D + \mathbf{B}_2 (1 - D) \qquad (2.40)$$

The steady-state equation can be obtained from Eq. 2-38 by setting all the perturbation terms and their derivatives to zero. Therefore, the steady-state equation is

$$\mathbf{AX} + \mathbf{B} V_d = 0 \qquad (2.41)$$

and therefore in Eq. 2-38

$$\dot{\tilde{\mathbf{x}}} = \mathbf{A} \tilde{\mathbf{x}} + \left[(\mathbf{A}_1 - \mathbf{A}_2) \mathbf{X} + (\mathbf{B}_1 - \mathbf{B}_2) V_d \right] \tilde{d} \qquad (2.42)$$

Similarly, usings Eqs. 2-34 to 2-36 in Eq. 2-33 results in

$$V_o + \tilde{v}_o = \mathbf{CX} + \mathbf{C} \tilde{\mathbf{x}} + \left[(\mathbf{C}_1 - \mathbf{C}_2) \mathbf{X} \right] \tilde{d} \qquad (2.43)$$

Where

$$\mathbf{C} = \mathbf{C}_1 D + \mathbf{C}_2 (1 - D) \qquad (2.44)$$

In Eq. 2-43, the steady-state output voltage is given as

$$V_o = \mathbf{CX} \qquad (2.45)$$

and therefore,

$$\tilde{v}_o = \mathbf{C}\tilde{\mathbf{x}} + \left[(\mathbf{C}_1 - \mathbf{C}_2)\mathbf{X}\right]\tilde{d} \tag{2.46}$$

Using Eqs. 2-41 and 2-45, the steady-state dc voltage transfer function is

$$\frac{V_o}{V_d} = -\mathbf{C}\mathbf{A}^{-1}\mathbf{B} \tag{2.47}$$

Step 4: Transformation of the ac Equations in to s-Domain to Solve for the Transfer Function.

Equations 2-42 and 2-46 consist of the ac perturbations. Using Laplace transformation in

Eq 2-42,

$$s\tilde{\mathbf{x}}(s) = \mathbf{A}\tilde{\mathbf{x}}(s) + \left[(\mathbf{A}_1 - \mathbf{A}_2)\mathbf{X} + (\mathbf{B}_1 - \mathbf{B}_2)V_d\right]\tilde{d}(s) \tag{2.48}$$

$$\tilde{\mathbf{x}}(s) = \left[s\mathbf{I} - \mathbf{A}\right]^{-1}\left[(\mathbf{A}_1 - \mathbf{A}_2)\mathbf{X} + (\mathbf{B}_1 - \mathbf{B}_2)V_d\right]\tilde{d}(s) \tag{2.49}$$

Where I is a unity matrix. Using a Laplace transformation in Eq. 2-46 and expressing in terms $\tilde{x}(s)$ in terms of $\tilde{d}(s)$ from Eq. 2-49 results in the desired transfer function Tp(s) of the powerstages.

$$T_p(s) = \frac{\tilde{v}_o(s)}{\tilde{d}(s)} = \mathbf{C}\left[s\mathbf{I} - \mathbf{A}\right]^{-1}\left[(\mathbf{A}_1 - \mathbf{A}_2)\mathbf{X} + (\mathbf{B}_1 - \mathbf{B}_2)V_d\right] + (\mathbf{C}_1 - \mathbf{C}_2)\mathbf{X} \tag{2.50}$$

Buck Converter

Now we will linearize the power stage and the output filter of the Buck Converter given in Figure 2.21. The two switches are represented by diodes.

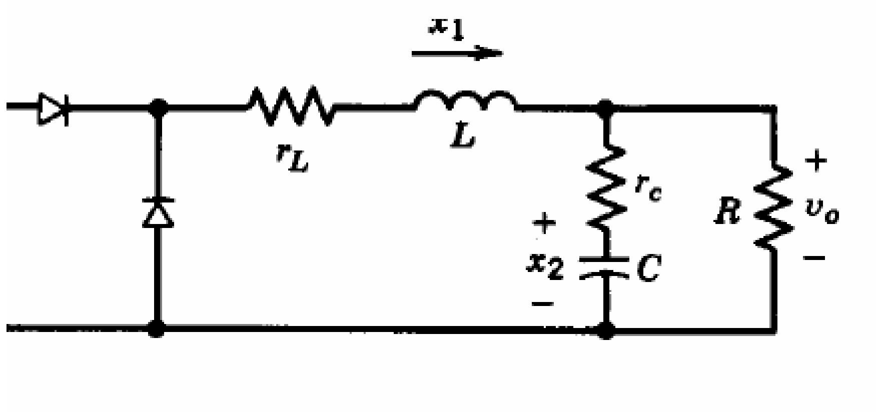

Figure 2.21: Buck Converter Circuit [9]

rL is inductor resistance, rc is the equivalent series resistance of the capacitor, and R is the load resistance.

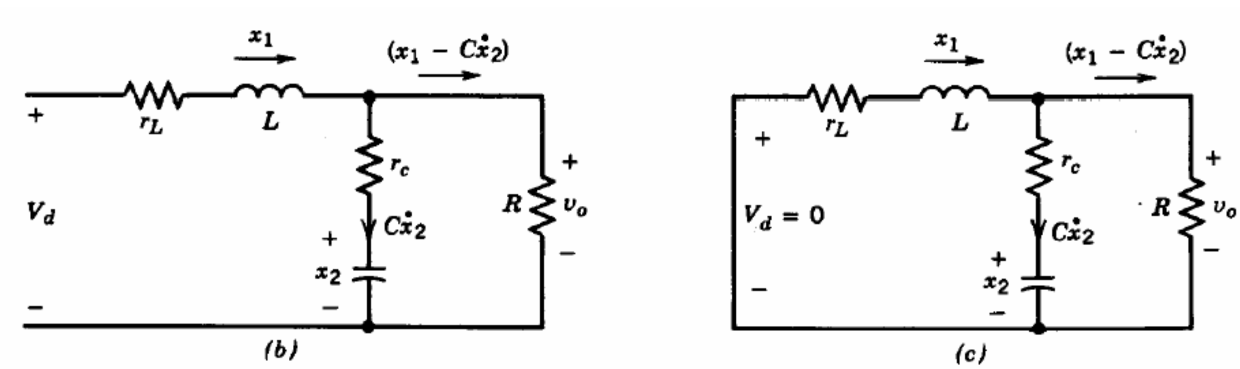

Figure 2.22: Buck Converter (a) switch on; (b) switch off [9]

From Figure 2.22 the following equations can be derived

$$-V_d + L\dot{x}_1 + r_L x_1 + R(x_1 - C\dot{x}_2) = 0$$

(2.51)

$$-x_2 - Cr_c\dot{x}_2 + R(x_1 - C\dot{x}_2) = 0$$

(2.52)

In matrix form, these two equations can be written as

$$\begin{bmatrix} \dot{x}_1 \\ \dot{x}_2 \end{bmatrix} = \begin{bmatrix} -\dfrac{Rr_c + Rr_L + r_c r_L}{L(R+r_c)} & -\dfrac{R}{L(R+r_c)} \\ \dfrac{R}{C(R+r_c)} & -\dfrac{1}{C(R+r_c)} \end{bmatrix} \begin{bmatrix} x_1 \\ x_2 \end{bmatrix} + \begin{bmatrix} \dfrac{1}{L} \\ 0 \end{bmatrix} V_d$$

(2.53)

Comparing the equations with Eq. 2-28 yields

$$\mathbf{A}_1 = \begin{bmatrix} -\dfrac{Rr_c + Rr_L + r_c r_L}{L(R+r_c)} & -\dfrac{R}{L(R+r_c)} \\ \dfrac{R}{C(R+r_c)} & -\dfrac{1}{C(R+r_c)} \end{bmatrix}$$

(2.54)

And

$$\mathbf{B}_1 = \begin{bmatrix} \frac{1}{L} \\ 0 \end{bmatrix} \tag{2.55}$$

$$\mathbf{A}_2 = \mathbf{A}_1 \tag{2.56}$$

$$\mathbf{B}_2 = 0 \tag{2.57}$$

The output voltage in both the circuit states is given as

$$v_o = R(x_1 - C\dot{x}_2) \tag{2.58}$$

$$= \frac{Rr_c}{R + r_c} x_1 + \frac{R}{R + r_c} x_2$$

Using $\dot{x}_2$ from Eq. 2-52

$$= \begin{bmatrix} \frac{Rr_c}{R + r_c} & \frac{R}{R + r_c} \end{bmatrix} \begin{bmatrix} x_1 \\ x_2 \end{bmatrix}$$

Therefore, in Eq. 2-30 and 2-31

$$\mathbf{C}_1 = \mathbf{C}_2 = \begin{bmatrix} \frac{Rr_c}{R+r_c} & \frac{R}{R+r_c} \end{bmatrix} \qquad (2.59)$$

$$\mathbf{A} = \mathbf{A}_1$$ (from Eq. 2-39 and Eq. 2-56) (Eq. 2-60)

$$\mathbf{B} = \mathbf{B}_1 D$$ (from Eq. 2-40 and Eq. 2-57) (Eq. 2-61)

$$\mathbf{C} = \mathbf{C}_1$$ (from Eq. 2-44 and Eq. 2-59) (Eq. 2-62)

Model Simplification

In all practical circuits,

$$R >> (r_c + r_L) \qquad (2.63)$$

Therefore, A and C are simplified as

$$\mathbf{A} = \mathbf{A}_1 = \mathbf{A}_2 = \begin{bmatrix} -\frac{r_c + r_L}{L} & -\frac{1}{L} \\ \frac{1}{C} & -\frac{1}{CR} \end{bmatrix} \qquad (2.64)$$

$$\mathbf{C} = \mathbf{C}_1 = \mathbf{C}_2 \cong \begin{bmatrix} r_c & 1 \end{bmatrix} \qquad (2.65)$$

and B remains unaffected as

$$\mathbf{B} = \mathbf{B}_1 D = \begin{bmatrix} 1/L \\ 0 \end{bmatrix} D \tag{2.66}$$

Where B2=0. From Eq. ,

$$\mathbf{A}^{-1} = \frac{LC}{1 + (r_c + r_L)/R} \begin{bmatrix} -\frac{1}{CR} & \frac{1}{L} \\ -\frac{1}{C} & -\frac{r_c + r_L}{L} \end{bmatrix} \tag{2.67}$$

$$\frac{V_o}{V_d} = D \frac{R + r_c}{R + (r_c + r_L)} \cong D \tag{2.68}$$

$$T_p(s) = \frac{\tilde{v}_o(s)}{\tilde{d}(s)} \cong V_d \frac{1 + s r_c C}{LC\{s^2 + s[1/CR + (r_c + r_L)/L] + 1/LC\}} \tag{2.69}$$

Eq. 2-69 is the Open Loop Transfer Function of the circuit represented in Figure 2.22. The term in the curly brackets in the denominator of Eq. 2-69 are of the form

$s^2 + 2\xi\omega_o s + \omega_0^2$, where

$$\omega_o = \frac{1}{\sqrt{LC}} \qquad (2.69)$$

$$\xi = \frac{1/CR + (r_c + r_L)/L}{2\omega_o} \qquad (2.70)$$

where $\omega_0 = F_{LC}$

Therefore, from Eq. 2-69 the transfer function

$$T_p(s) = \frac{\tilde{v}_o(s)}{\tilde{d}(s)} = V_d \frac{\omega_o^2}{\omega_z} \frac{s + \omega_z}{s^2 + 2\xi\omega_o s + \omega_o^2} \qquad (2.71)$$

where a zero is introduced due to the equivalent series resistance of the output capacitor at the frequency.

$$\omega_z = F_{ESR} = \frac{1}{r_c C} \qquad (2.72)$$

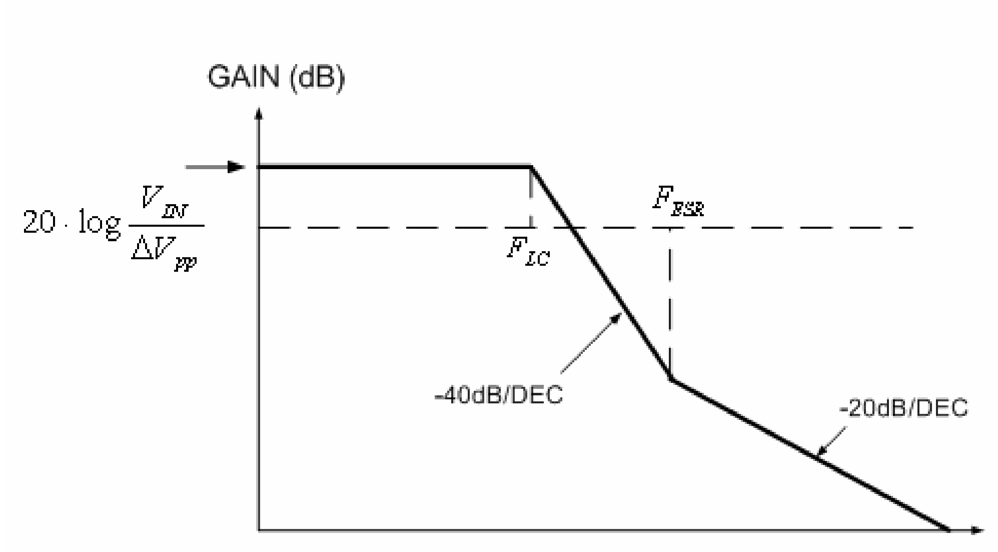

Figure 2.23: Open Loop System Gain [12]

Figure 2.23 shows the Bode plot for the transfer function in Eq. 2-71 using the numerical values given in the Fig. It shows the transfer function has a fixed gain and a minimal phase shift at low frequencies. Beyond the resonant frequency $w_0 = \sqrt{1/LC}$ of the LC output filter, the gain begins to fall with a slope of – 40dB/decade and the phase tends toward -180°. At frequencies beyond w_z The gain falls with a slope of -20 dB/decade and the phase tends toward -90°. The gain plots shifts vertically with V_d but the phase plot is not affected.

2.13 Stability Criteria

It is the desire of all designers of power supplies, whether they are switching or not, for accurate and tight regulation of the output voltage(s). To accomplish regulation we need to add a feedback loop. The feedback loop can cause an otherwise stable system to become unstable. Even though the transfer function of the original converter might not contain any right hand poles but after feedback it is possible that right hand poles may be introduced. Also we need to introduce a high DC gain. But with high gain again comes the possibility of instability. These two issues determine the need to have stability criteria for a power supply. Hence, feedback compensation design involves selection of a suitable compensation circuit configuration and positioning of its poles and zeros to yield an open loop transfer function. Certain very important parameters need to be taken in to account when calculating the stability of the power supply.

• Variations in input voltage do not cause instability.

• Allow for variations in the peak-to-peak oscillator voltage.

• Error amplifier (which we will discuss in the next section) has sufficient attenuation at the switching frequency so that it does not amplify the output voltage ripple and cause sub harmonic oscillations.

• Mid-frequency gain is greater than zero to prevent a large overshoot at turn-on and during transient conditions.

• Error amplifier has the drive capability to drive the feedback network properly.

• High gain at low frequency region to provide tight output voltage regulation and minimize the steady-state error in the power supply output.

• The phase margin determines the transient response of the output voltage in response to sudden changes in the load and the input voltage. The difference between 180° and the actual phase when the gain reaches unity gain. (In this case it is approaching zero.) Phase margins of 45º to 60º (360º degree minus the total closed-loop phase lag) are considered safe values that yield well-damped transient load responses. The recommended value is 45° to 60° [9].

• Gain Margin is the difference between unity gain (zero dB) and the actual gain when the phase reaches 180°. (In this case it is a positive number.) The recommended value is -6dB to -12 dB.

• A crossover frequency (or bandwidth), c f , of between one tenth and one fourth of a switching frequency for a system to respond sufficiently fast to transients, such as a sudden change of load. A commonly used derivative from the above definitions is that if the slope of the gain response as it crosses the unity-gain axis is not more than -20 dB / decade, the phase margin will be greater than 45° and the system will be stable.

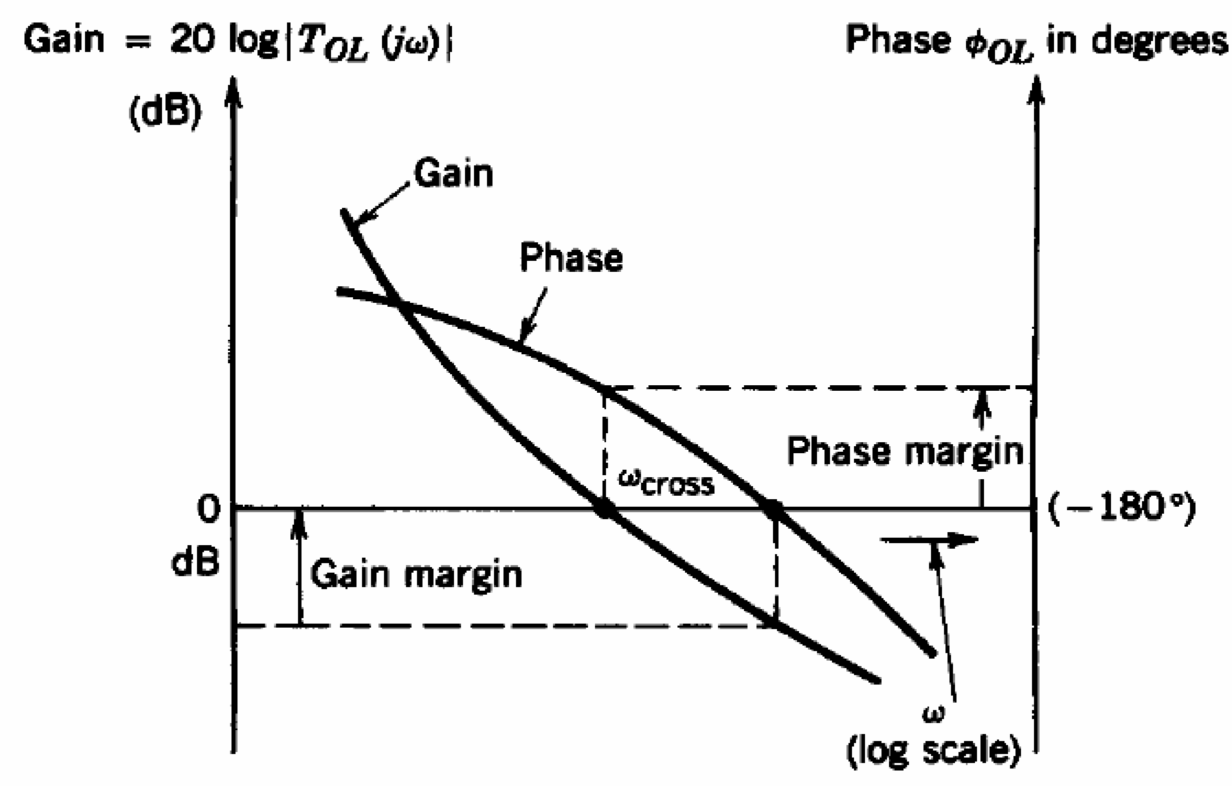

Figure2 .24: Definitions of the crossover frequency, phase and gain margins [9]

CHAPTER 3
LINEAR CONTROL DESIGN FOR DC-DC CONVERTERS

Presented in this chapter is the control design for DC-DC converters using linear control methods. An accurate model is essential to design linear controllers. Small signal models for buck converter were obtained using the standard state-space averaging techniques. The actual frequency response was also measured to compare with the small signal model. For the buck converter, the actual frequency response matches the small signal model. Therefore, for the buck converter, the control design was based on the small signal model. Frequency response and root locus methods [11] may be utilized to design linear controllers. In the frequency response method, analog PID and PI controllers were designed based on the converters' small signal models [12]. The system was compensated to achieve high loop gain, wide bandwidth and sufficient phase margin. The PID and PI controllers were then transformed into digital controllers using the backward integration method.

3.1 Buck Converters

The buck converter's small signal control-to-output transfer function, derived by the standard state-space averaging technique, is given by (3.1).

$$\frac{\hat{v}_O(s)}{\hat{d}(s)} = \left(\frac{V_O}{D}\right)\left[\frac{1+sR_CC}{1+s\left(R_CC+\left[R//R_L\right]C+\dfrac{L}{R+R_L}\right)+s^2LC\left(\dfrac{R+R_C}{R+R_L}\right)}\right] \tag{3.1}$$

The small signal input-to-output transfer function is given by (3.2):

$$\frac{\hat{v}_O(s)}{\hat{v}_{in}(s)} = \left[\frac{DR}{R+R_L}\right]\left[\frac{1+sR_CC}{1+s\left(R_CC+\left[R//R_L\right]C+\dfrac{L}{R+R_L}\right)+s^2LC\left(\dfrac{R+R_C}{R+R_L}\right)}\right] \tag{3.2}$$

In the transfer functions, Vin and Vo are the input and output voltages respectively.$\widehat{v_o}$(s),$\widehat{v_{in}}$(s) and $\widehat{d_s}$ are the small variations of the output voltage, input voltage and duty cycle, respectively. D is the duty cycle, C is the output capacitance, L is the inductance, and R is the load resistance. RC and RL are the ESR of C and L.

The control-to-output transfer function is utilized to design the controller. It is a common two-pole low pass filter, with a left half plane zero introduced by the ESR of the filter capacitance [13]. The cutoff frequency of the low pass filter is $w_c = {}^{1}/_{\sqrt{LC}}$. The magnitude falls with a slope of –40 dB/decade at the cutoff frequency. The phase associated with it is a –180 degree phase delay. The zero is at $-\frac{1}{R_c C}$. There is a 20 dB/decade magnitude rise at that frequency and the phase shift is 90 degrees. The magnitude of the transfer function depends on the duty cycle D. When D increases, the magnitude decreases; when D decreases, the magnitude increases. However, variations of D don't change the shape of the magnitude plot of the transfer function. It only shifts the magnitude upward or downward.

The buck converter's nominal operating point is as follows: Vin = 20 V, Vo = 12 V, and D = 0.6. The capacitance C is 1000 µF, L is 150 µH, and R is 10 Ω. The parasitic elements RC and RL are estimated to be 30 mΩ and 10 mΩ, respectively [14, 15]. This buck converter was used as a prototype buck converter in this dissertation. The controlto- output transfer function at the nominal operating point is given by (3.3):

$$\frac{\hat{v}_o(s)}{\hat{d}(s)} = \frac{6*10^{-4}s+20}{1.503*10^{-7}s^2+5.4975*10^{-5}s+1} \qquad (3.3)$$

The Bode plot of the transfer function is shown in Figure 3.1. The model has complex conjugate poles at –615.9 ± j2481.5, which causes a 180 degrees phase delay at the approximate frequency of 2500 radians/s. The model also has a zero at 33,333 radians/s. Frequency response data for the prototype buck converter was measured using a Model 102B analog network analyzer by AP Instruments. Figure 3.2 shows the frequency response of the buck converter near the nominal operating point. It compares favorably with the theoretical model; thus linear controllers can be designed based on the theoretical model.

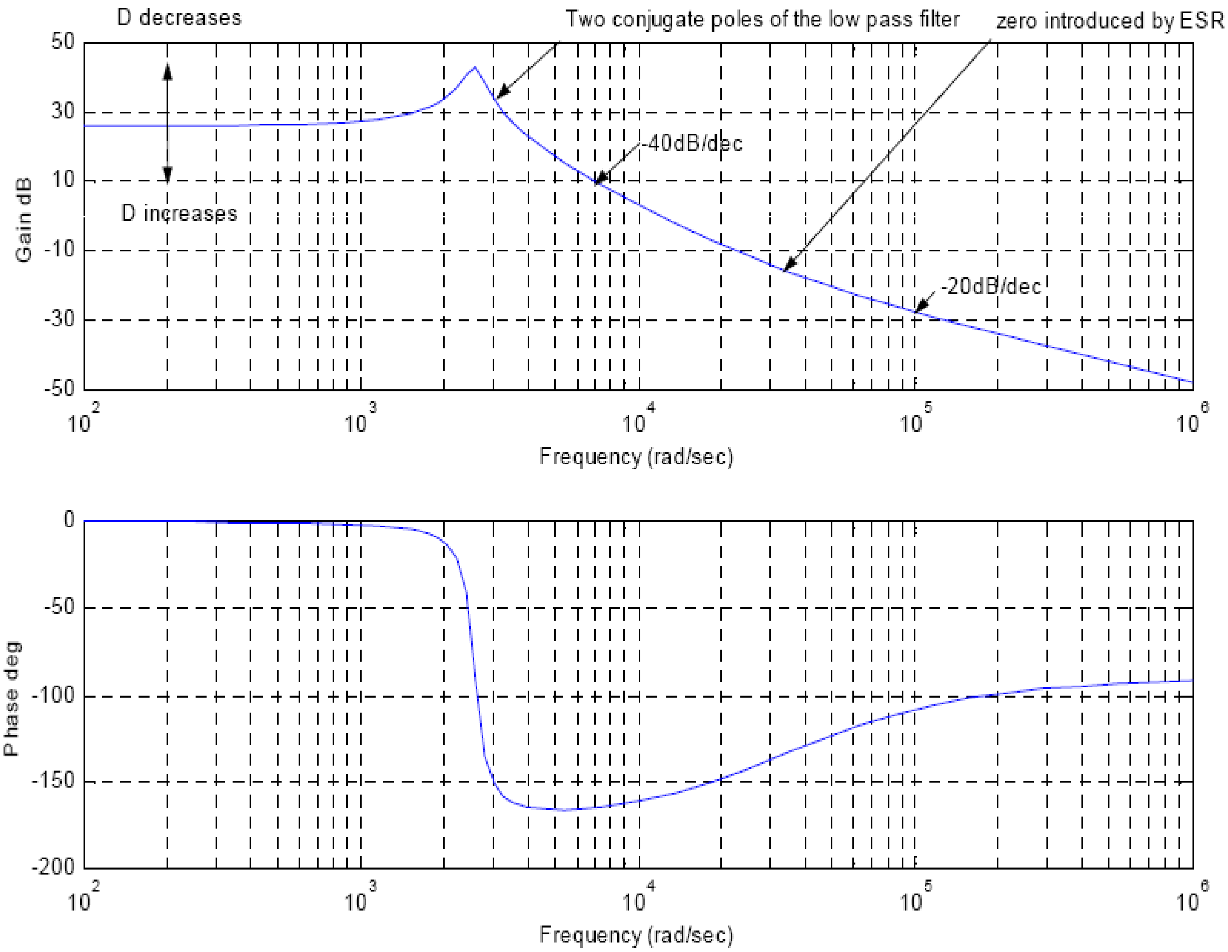

Figure 3.1 Bode plot of the state-space averaged model of the buck converter

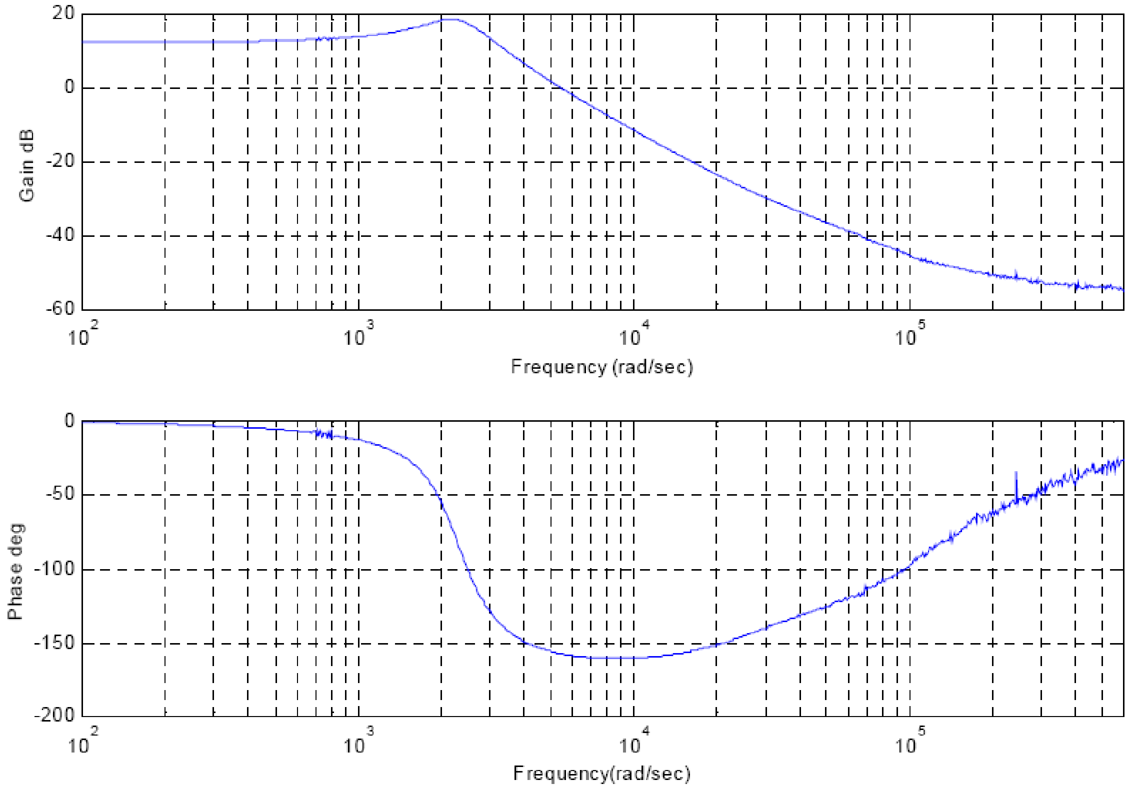

Figure 3.2 Frequency response of the buck converter obtained by the analog analyzer

3.2 Digital Controller Design for DC-DC Converters Using Frequency Response Techniques

In DC-DC converters, the output voltage is a function of the input line voltage, the duty cycle and the load current. It is desirable to have a constant output voltage in the event of disturbances such as a sudden change of input voltage or load current. Negative feedback control is applied to DC-DC converters to automatically adjust the duty cycle to obtain the desired output voltage with high accuracy in spite of disturbance [16].

In this section, frequency response techniques are used to design digital controllers for DC-DC converters. The compensated system is expected to have the following characteristics [17]. Firstly, the loop gain should be high at lower frequencies to minimize steady-state error and increase rejection to disturbances of input voltage and load current variations. Secondly, the crossover frequency should be as high as possible, but about an order of magnitude below the switching frequency to allow the DC-DC converter to respond quickly to the transients. Thirdly, the phase margin should be sufficient to ensure the system's stability. When the phase margin of the loop gain is positive, the system is stable. A phase margin of 45o to 60o is desirable. Phase margin determines the transient response of the DC-DC converter. An increase of the phase margin makes the system more stable with less ringing and oscillation. There is a qualitative relationship between the phase margin and the closedloop damping factor Q. To obtain Q = 1, a phase margin of 52o is required, and to obtain Q = 0.5, a phase margin of at least 76o is needed. The damping factor Q determines the shape of the transient response. When Q is equal to 0.5, the closed-loop system has two real poles at the same frequency, and the system is critically damped. The transient response will be fast without overshoot. When Q is larger than 0.5, there are two complex conjugate poles, and the system is underdamped. The transient response will have an oscillatory-type waveform with decaying magnitude. The higher Q, the higher overshoot the transient response will have. When Q is less than 0.5, the closed-loop system has two real poles at two different frequencies, and the system is overdamped. The transient response is a decaying exponential function of time with the time constant determined by the pole at the lower frequency. When Q is very low, the low-frequency pole results in a slow transient response. To design a controller using the frequency response method, phase-lead, phaselag or lead-lag compensation is usually used. A proportional-derivative (PD) controller is phase-lead compensation. PD controllers are used to increase the phase margin and improve the cross-over frequency. A zero is placed at frequency ωZ far below the cross over frequency to improve the phase margin. The transfer function of a PD controller is shown in (3.4)

$$G_C(s) = K_C \frac{(1+\frac{s}{\omega_z})}{(1+\frac{s}{\omega_P})} \quad (3.4)$$

The pole at ωP is placed well below the switching frequency to avoid amplification of the switching noise. The maximum phase shift occurs at the geometric mean of the pole ωP and the zero ωZ. To obtain maximum phase margin improvement, the maximum phase shift should be placed at the cross-over frequency.

A proportional-integral (PI) controller is a phase-lag controller. A PI controller is used to increase the low frequency loop gain, thus reducing steady-state error. The transfer function of a PI controller is shown in (3.5).

$$G_C(s) = K_P + \frac{K_I}{s} = \frac{K_P s + K_I}{s}$$

(3.5)

The PI controller has a pole at the origin. Both PD and PI controllers are first-order controllers. By using a lead-lag compensator, the advantages of lead compensation and lag compensation can be combined to obtain sufficient phase margin, high loop gain and wide control bandwidth. A proportional-integral-derivative (PID) controller is a lead-lag compensator. It is the most widely used compensator in feedback control systems. The PID controller is defined by (3.6), where e(t) is the compensator input and m(t) is the compensator output.

$$m(t) = K_P e(t) + K_I \int_0^t e(\tau) d\tau + K_D \frac{de(t)}{dt}$$

(3.6)

The Laplace transform of (3.6) yields the transfer function in (3.7).

$$G_C(s) = \frac{M(s)}{E(s)} = K_P + \frac{K_I}{s} + K_D s$$

(3.7)

The integral term is phase-lag and the derivative term is phase-lead. The low frequency gain is improved by the integral term, and the low-frequency components of the output voltage are accurately regulated. At high frequency, the phase margin and cross-over frequency are improved by the derivative term, which improves the system's stability and the speed of the transient response. An increase in the proportional term will increase the speed of system response; however, too much proportional gain will make the system unstable. A PID controller and a PI controller were designed for the buck converter in the following sections.

3.2.1 Buck Converters

A PID and a PI controller were designed for the buck converter for operation during a startup transient and steady state, respectively. The derivative term in a PID controller is susceptible to noise and measurement error of the system, which could result in oscillation of the duty cycle during steady state. However, during a transient, the derivative term is needed to reduce the settling time by predicting the changes in error. Therefore, the system switches between PID and PI controllers during transient and steady state to obtain the desired response. The PID controller is applied during start up to obtain a fast transient response. The PI controller is applied during steady state to reduce oscillation of the duty cycle and improve the system's stability.

3.2.1.1 PID Controller Design for Buck Converters

A PID controller was designed for the buck converter to improve the loop gain, cross-over frequency and phase margin. One zero was placed an octave below the cutoff frequency (approximately 260 radians/s) and the other one at 4600 radians/s. The transfer function of the PID controller is given by (3.8):

$$G_c(s) = 0.5786 + \frac{142.4}{s} + 0.000119s \tag{3.8}$$

The Bode plot for the compensated system is shown in Figure 3.3. As can be seen in this plot, the gain at low frequency is high, the phase margin is 107 degrees and the bandwidth is 19100 radians/s.

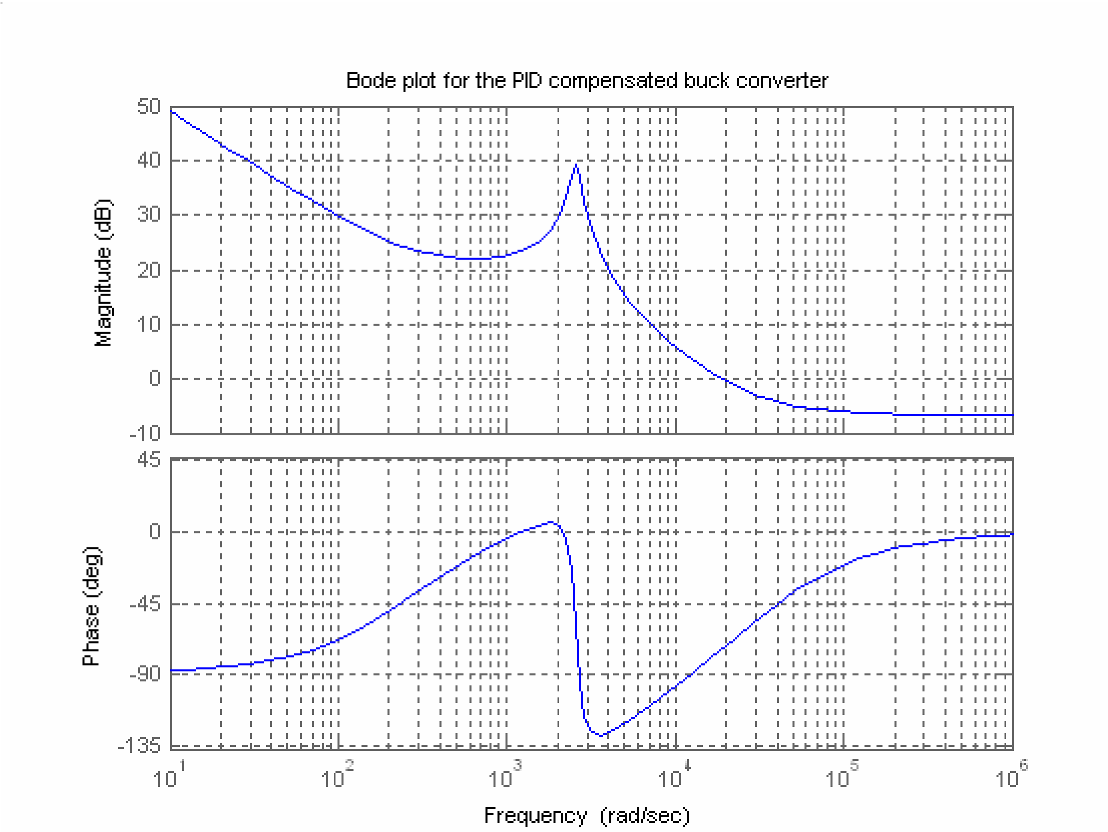

Figure 3.3 Bode plot of PID controller compensated buck converter

3.2.1.2 PI Controller Design for Buck Converters

A PI controller was also designed for the control of the buck converter at steady state to reduce steady-state oscillation. One pole was placed at the origin, and one zero was placed at 800 radians/s. The DC gain of the controller was adjusted to obtain sufficient phase margin and high cross-over frequency. The transfer function of the PI controller is given by (3.9):

$$G_c(s) = 0.75 + \frac{600}{s} \qquad (3.9)$$

The Bode plot for the PI compensated system is shown in Figure 3.4. The Bode plot shows that the phase margin is 15.4 degrees and the bandwidth is 10600 radians/s.

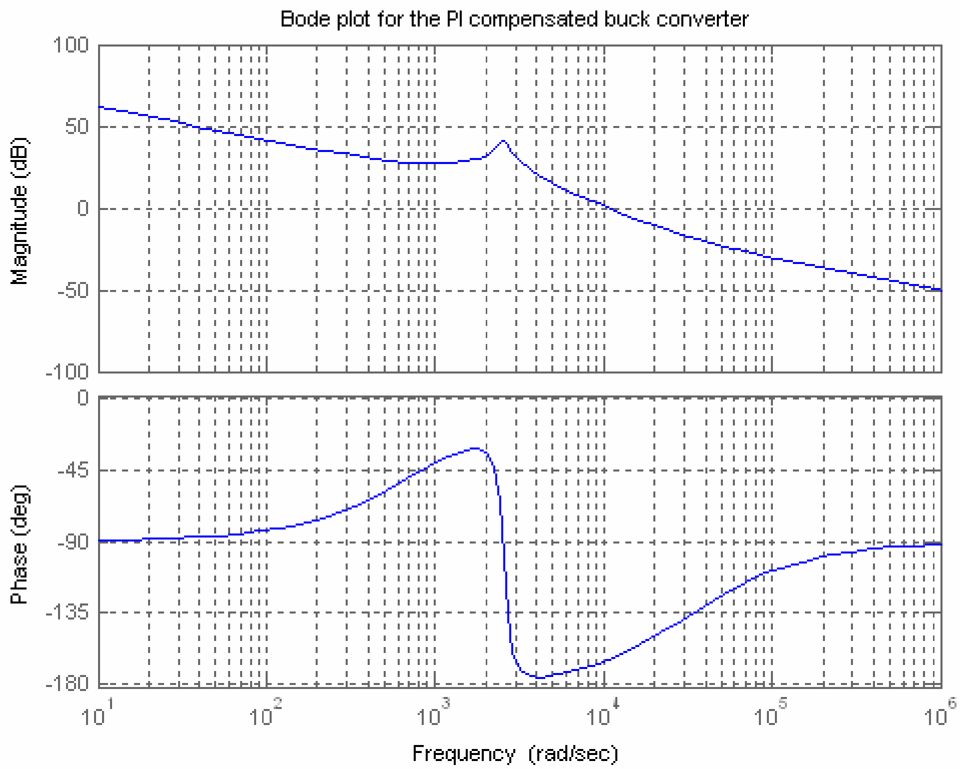

Figure 3.4 Bode plot of PI controller compensated buck converter

3.2.3 Transformation from an Analog Controller to a Digital Controller

The design in the continuous-time domain was transformed into the discrete-time domain using the backward integration method (Euler Method) [17, 18]. Using the Euler method, the transfer function of a numerical integrator is shown in (3.10).

$$M(z) = \frac{Tz}{z-1} E(z) \qquad (3.10)$$

The transfer function of a numerical differentiator is the reciprocal of the transfer function of the numerical integrator shown in (3.10). Therefore, by substituting (3.10) and its reciprocal into the PID controller's s-domain transfer function in (3.7), the digital PID controller's transfer function is shown in (3.11).

$$G_C(z) = K_P + \frac{K_I T z}{z-1} + \frac{K_D(z-1)}{Tz} \quad (3.11)$$

Similarly, by substituting (3.10) and its reciprocal into the PI controller's s-domain transfer function in (3.5), the digital PI controller's transfer function is shown in (3.12).

$$G_C(z) = K_P + \frac{K_I T z}{z-1} \quad (3.12)$$

CHAPTER 4
FUZZY CONTROLLER DESIGN FOR DC-DC CONVERTERS

Linear controllers for DC-DC converters are usually designed based on mathematical models. To obtain a certain performance objective, an accurate model is essential. In the previous chapter, linear controllers were designed for buck converter based on each converter's small signal model using frequency response and root locus design methods. The small signal model changes due to variations in operating point. Changes in the duty cycle only affect the magnitude of the buck converter's small signal model.

To achieve a stable and fast response, two solutions are possible. One is to develop a more accurate model for the converter. However, the model may become too complex to use in controller development. A second solution is to use a nonlinear controller [16]. Since fuzzy controllers don't require a precise mathematical model, they are well suited to nonlinear, time-variant systems. The design of fuzzy controllers is presented in this chapter. The first section introduces the concept of fuzzy control. The second section is mainly focused on the design of a fuzzy controller for the buck converter. Implementation of fuzzy controllers is presented in the third section of this chapter.

4.1 Introduction to Fuzzy Control

Fuzzy control is an artificial intelligence technique that is widely used in control systems. It provides a convenient method for constructing nonlinear controllers from heuristic information.

Conventional controllers are designed based on a mathematical model. Closed loop control specifications include disturbance rejection properties, insenstivity to plant parameter variations, stability, rise time, overshoot and settling time and steady-state error. Based on these specifications, conventional controllers are designed. Major conventional control methods include classical control methods (frequency response and root locus techniques), state-space methods, optimal control, robust control, adaptive control, sliding mode control and other nonlinear control methods such as feedback linearization and backstepping. These conventional control methods provide a variety of ways to utilize information from mathematical models on how to obtain good control.

Different from conventional control, fuzzy control is based on the expert knowledge of the system. Fuzzy control provides a formal methodology to represent and implement a human's heuristic knowledge about how to control the system. A block diagram of a fuzzy control system is shown in Figure. 4.1. A fuzzy controller contains four main components: (1) the fuzzification interface that converts its inputs into information that the inference mechanism can use to activate and apply rules, (2) the rule base which contains the expert's linguistic description of how to achieve good control, (3) the inference mechanism that evaluates which control rules are relevant in the current situation, and (4) the defuzzification interface which converts the conclusion from the inference mechanism into the control input to the plant [19].

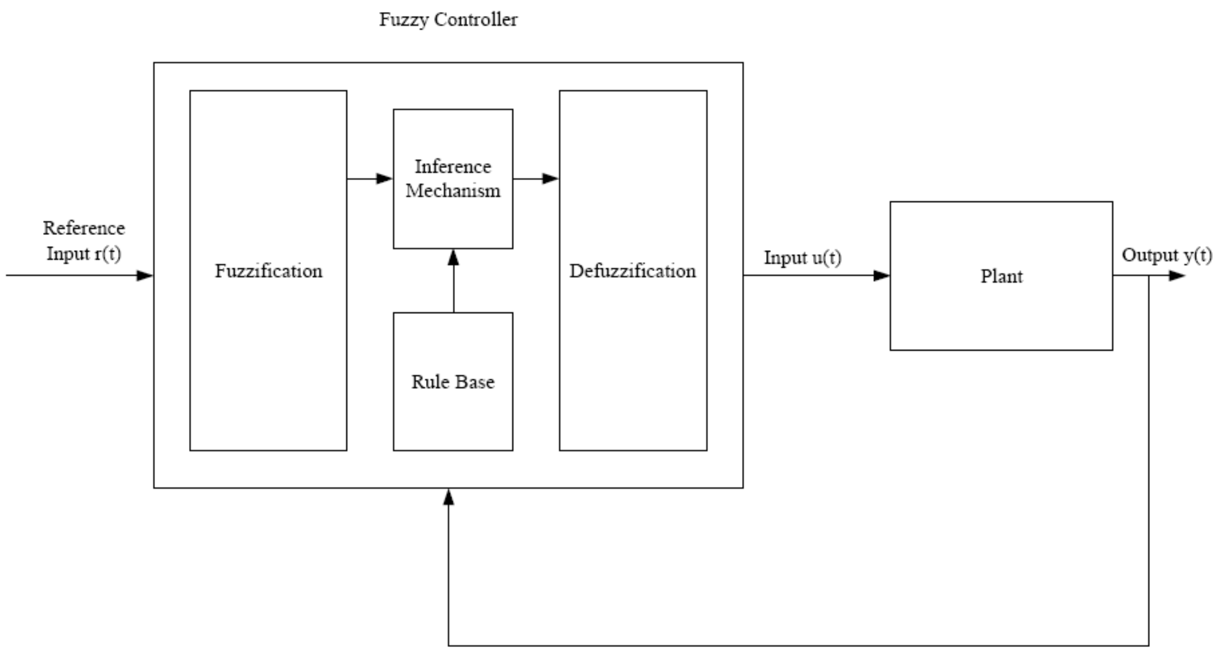

Figure 4.1 Block diagram of fuzzy control system

The performance objectives and design constraints are the same as those for conventional control. Design of fuzzy controllers involves the following procedures: (1) choose the fuzzy controller's inputs and outputs, (2) choose the preprocessing for the controller inputs and postprocessing for the controller outputs, and (3) design each of the four components of the fuzzy controller shown in Figure 4.1.

4.2 Fuzzy Control Design for DC-DC Converters

A fuzzy controller for a DC-DC converter has two inputs. The first input is the error in the output voltage e[k]=Ref-ADC[k], where ADC[k] is the converted digital value of the kth sample, and Ref is the digital value corresponding to the desired output voltage. The second input, ce[k]=e[k]-e[k-1], is the difference between the error of the kth sample and the error of the (k-1)th sample. The two inputs are multiplied by the scaling factors g0 and g1, respectively, and then fed into the fuzzy controller. The output of the fuzzy controller is the change in duty cycle δd[k]. It is scaled by a linear gain h. The scaling factors g0, g1 and h can be tuned to obtain a satisfactory response. There are two methods to calculate the new duty cycle from the fuzzy controller's output δd[k]. In the first method, the output of the fuzzy controller, scaled by the output gain h, is added to the previous sampling period's duty cycle d[k-1], which is written in (4.1).

$$\mathrm{d}[k] = \mathrm{d}[k\text{-}1]+\mathrm{h}*\delta\mathrm{d}[k] \qquad (4.1)$$

The integration of the fuzzy controller's output increases the system type and improves steady-state error. The Simulink model of the fuzzy controller using (4.1) to calculate the duty cycle d[k] is shown in Figure 4.2. The disadvantage of this method is that the output gain h has to be tuned to be very small to avoid oscillation in steady state. Since the change in duty cycle is accumulated every sampling period, the duty cycle varies around its nominal value during steady state, which could lead to oscillation. Quantization errors in digital controllers increase the magnitude of the oscillation, because digital controllers are restricted to a finite set of values. Oscillation between the maximum and minimum values of the duty cycle may even occur if h is relatively large compared to the duty cycle range. A very small output gain h tends to increase the transient response time because more sampling periods are necessary to arrive at the desired duty cycle [20].

A second method to calculate the new duty cycle is to add the output of the fuzzy controller scaled by h to the output of a linear integrator, which is shown by (4.2),

$$\mathrm{d}[k] = \mathrm{Ki}*\mathrm{I}[k]+\mathrm{h}*\delta\mathrm{d}[k] \qquad (4.2)$$

where I[k] is the output of the linear integrator of the error e[k], and KI is the gain of the integrator. The linear integrator is applied to eliminate steady-state error. The Simulink model of the fuzzy controller using (4.2) to generate the duty cycle d[k] is shown in Figure 4.3. In this method, the output gain h can be increased because the fuzzy controller's output is not accumulated every sampling period.

In order to prevent the MOSFET from being turned on or off for a full switching period, the duty cycle d[k] is limited to be between 10% and 90% for the buck converter. From the literature, the first method in Figure 4.2 is more prevalently used than the second method in Figure 4.3 [16, 21]. In this dissertation, only the second method was applied to the buck converter to obtain satisfactory response.

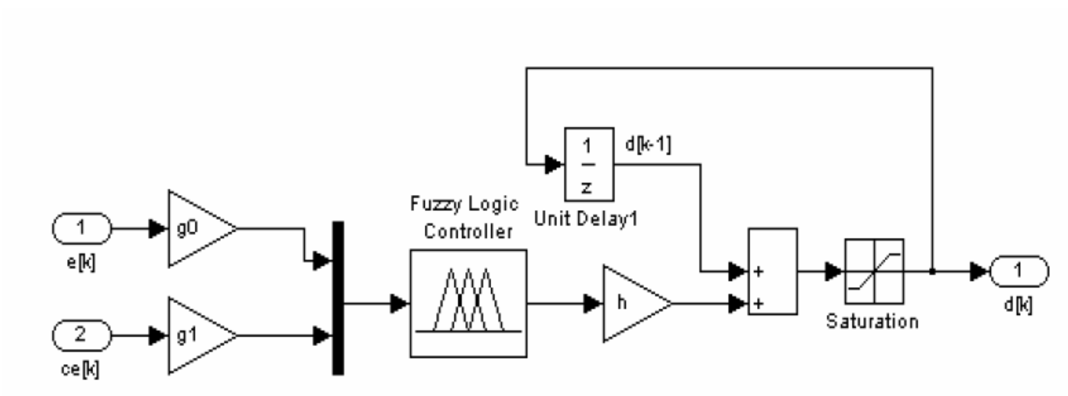

Figure 4.2 Simulink model of the fuzzy controller for the DC-DC converters Method 1: (d[k] = d[k-1]+h*δd[k].)

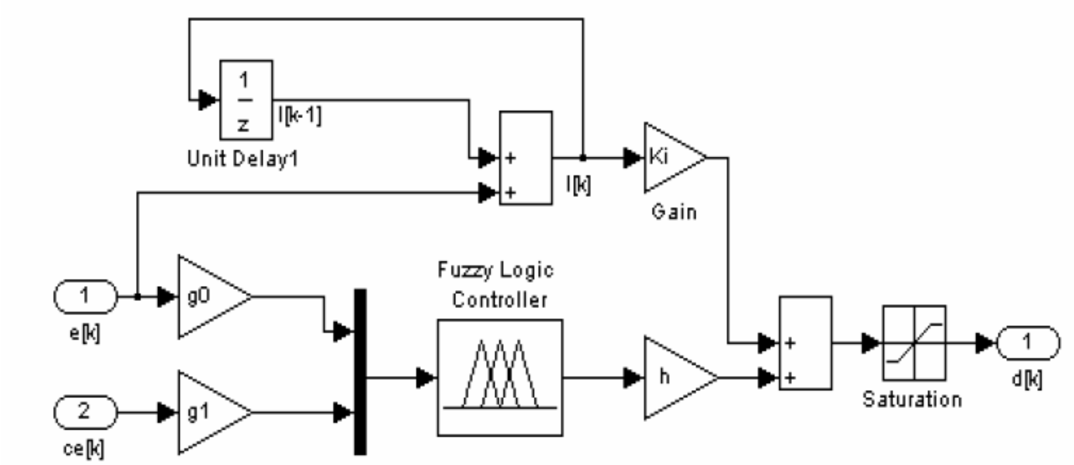

Figure 4.3 Simulink model of the fuzzy controller with a linear integrator for the DC-DC converters Method 2:(d[k] = K_I*I[k]+h*δd[k])

4.2.1 Fuzzification

First, the linguistic values are quantified using membership functions. Each universe of discourse is divided into fuzzy subsets. There were 17 fuzzy subsets in the fuzzy controller for the buck converter: N8, N7, N6, N5, N4, N3, N2, N1, Z, P1, P2, P3, P4, P5, P6, P7 and P8, where N indicates negative, Z represents zero, and P indicates positive.We have used 7 fuzzy subsets which are given as NB(negativebig),NM(negativemedium),NS(negativesmall),Z(zero),PS(positive small),PM(positive medium),PB(positive small). The membership functions for e[k] and ce[k] are shown in Figure 4.4.

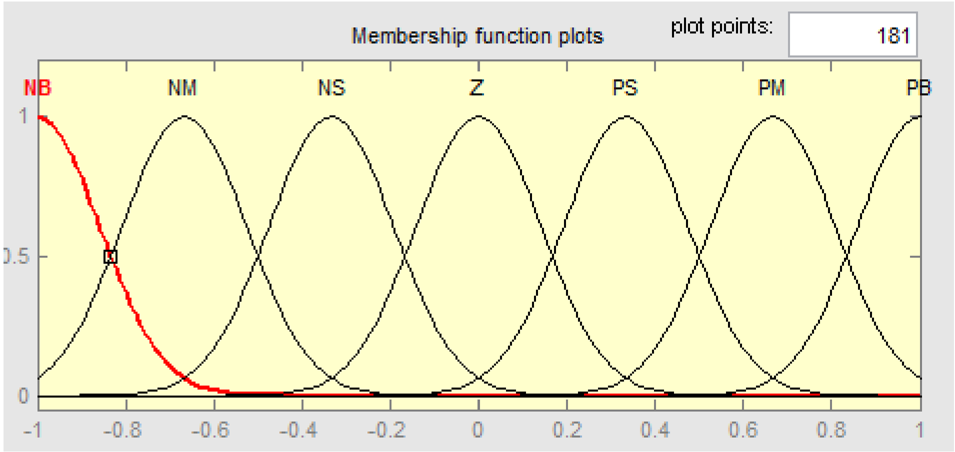

Figure 4.4 Membership functions of the inputs e[k] and ce[k] for the buck converter

A gaussian-shaped membership function was used for this controller design for the ease of implementation.

4.2.2 Rule Base

The rule base is derived from general knowledge of DC-DC converters, and adjusted based on experimental results. There is a tradeoff between the size of the rule base and the performance of the controller. A 7*7 rule base was designed and implemented for the buck converter.

		Change in error (CE)						
		NB	NM	NS	Z	PS	PM	PB
Error (E)	PB	Z	NS	NM	NM	NB	NB	NB
	PM	PS	Z	NS	NM	NM	NM	NB
	PS	PM	PS	Z	NS	NS	NM	NB
	Z	PM	PM	PS	Z	NS	NM	NM
	NS	PB	PM	PS	PS	Z	NS	NM
	NM	PB	PM	PM	PM	PS	Z	NS
	NB	PB	PB	PB	PM	PM	PS	Z

Table 4.1: 7×7 Rule base of the sliding mode fuzzy controller

4.2.3 Inference Mechanism

The results of the inference mechanism include the weighing factor wi and the change in duty cycle ci of the individual rule [22]. The weighing factor wi is obtained by Mamdani's min fuzzy implication of µe(e[k]) and µce(ce[k]), where wi = min{µe(e[k]), µce(ce[k])} and µe(e[k]), µce(ce[k]) are the membership degrees [21]. ci is taken from the rule table. The change in duty cycle inferred by the ith rule, zi is written in (4.3).

$$z_i = w_i \times c_i = \min\{\mu_e(e[k]), \mu_{ce}(ce[k])\} \times c_i \quad (4.3)$$

4.2.4 Defuzzification

The center of average method is used to obtain the fuzzy controller's output δd[k], which is given in (4.4).

$$\delta d[k] = \frac{\sum_{i=1}^{N} w_i \times c_i}{\sum_{i=1}^{N} w_i} \quad (4.4)$$

CHAPTER 5 SIMULATION RESULTS

5 Design Parameters

The converter, in this thesis, is designed for CCM operation, and needs to operate from a 20 ± 3 V DC source. The output voltage, V_o, from the converter must be 12 ± 0.1 V with a steady-state ripple of less than 2.5 percent or 0.05 V. The converter is required to maintain output voltage while the output current, I_o, varies between 1 A and 10 A.

5.1 Results With PI Compensated Buck Converter

5.1.1 Response to 20v DC Power Source with PI Compensated Buck Converter

The buck circuit is required to reach a steady-state voltage of 2 V when connected to a 20 V power source. The converter is connected to the source operating under minimum loading conditions: 1 A. The duty cycle is set to 0.6. From the response shown in Figure , it can be seen that the output voltage meets the steady-state op-rating conditions; the steady-state error is 0.4 percent. The transient characteristics are a maximum overshoot of 69.4 percent, a rise time of 88 sec, and a settling time of 2.75 msec.

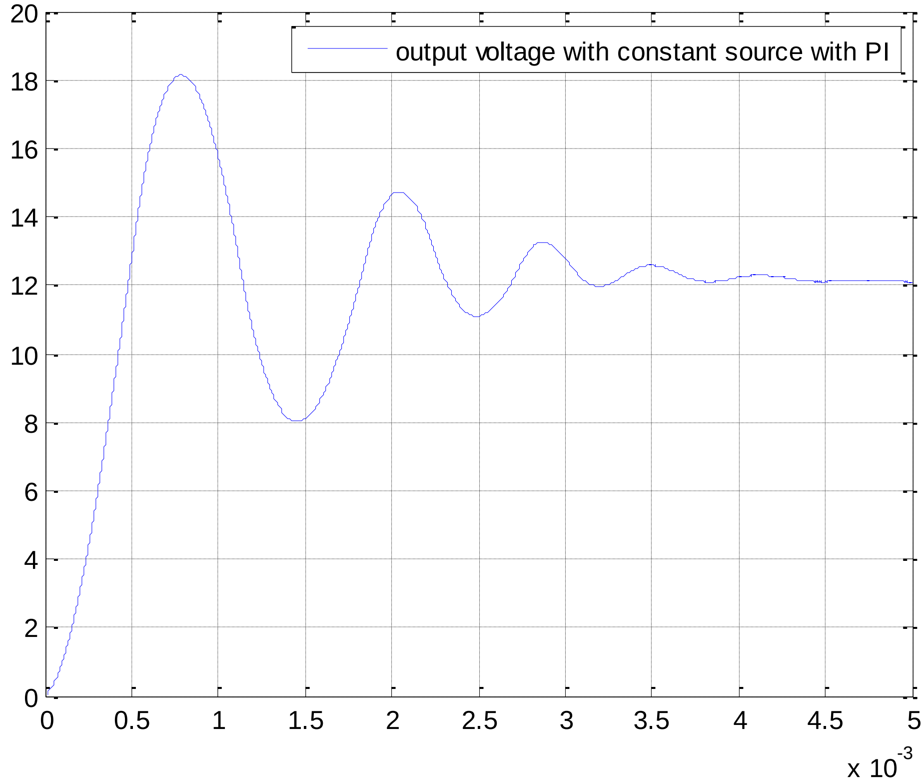

Figure 5.1 : Response to constant source with PI compensated Buck converter

5.1.2 Response to 1V change in Power Source with PI Compensated Buck Converter

The load is capable of varying between 1 A and 10 A after the system has reached steady-state. The changes in the load alter the operating condition of the system and therefore changes in the load are analyzed to see if the system still remains within operating parameters after a change in the load has occurred. To test the system, a 1 A step change in the load is applied. The response can be seen in Figure 5.2. The output voltage response still meets steady-state operating conditions but the transient operating parameters are not satisfied. The steady-state error is 1 percent. The maximum overshoot is 7.5 percent, and the maximum undershoot is 13.5 percent, which exceed outside the 5 percent requirement.

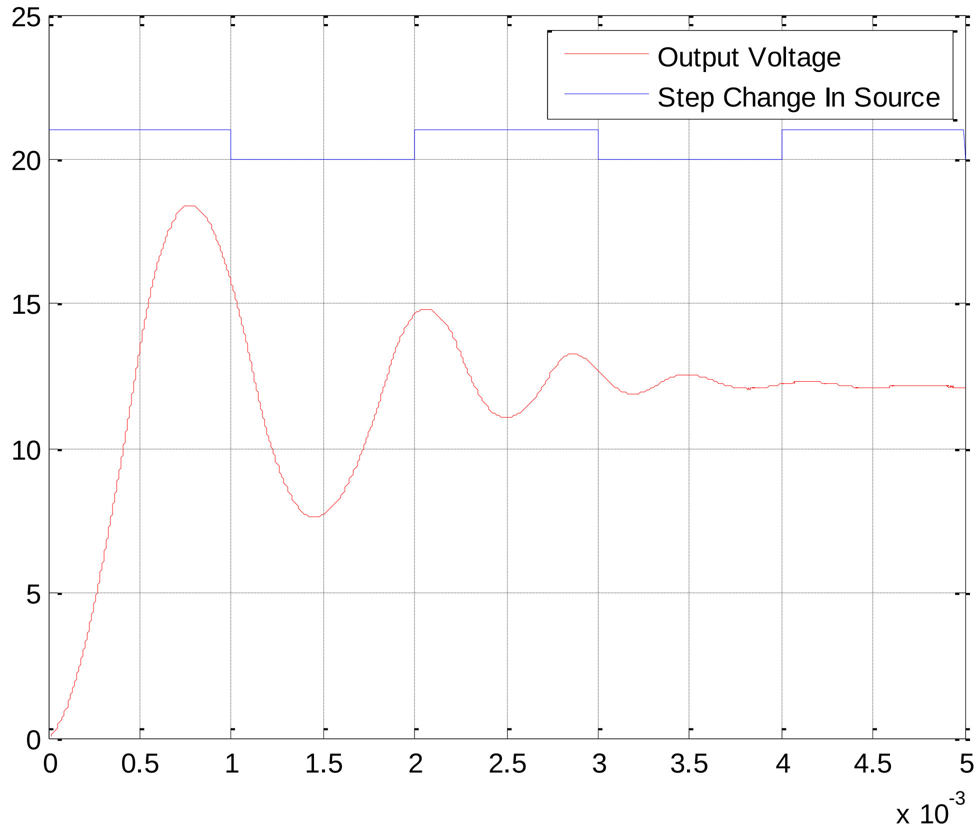

Figure 5.2 : Response to 1V change in power source with PI compensated Buck converter

5.1.3 Response to 1A change in Load current with PI Compensated Buck Converter

The source is capable of varying ±3 V, and changes in the source voltage affect the operating condition of the system, and therefore can impact the output voltage. To test if the system remains within operating parameters, a 1 V step change is applied to the converter after response from the power source has reached steady-state. The response of the output voltage can be seen in Figure 5.3. The output voltage no longer meets the steady-state operating requirements. The steady-state error is 8.4 percent. In addition, the peak response of the system is 2.282 V, which exceeds the 5 percent requirement.

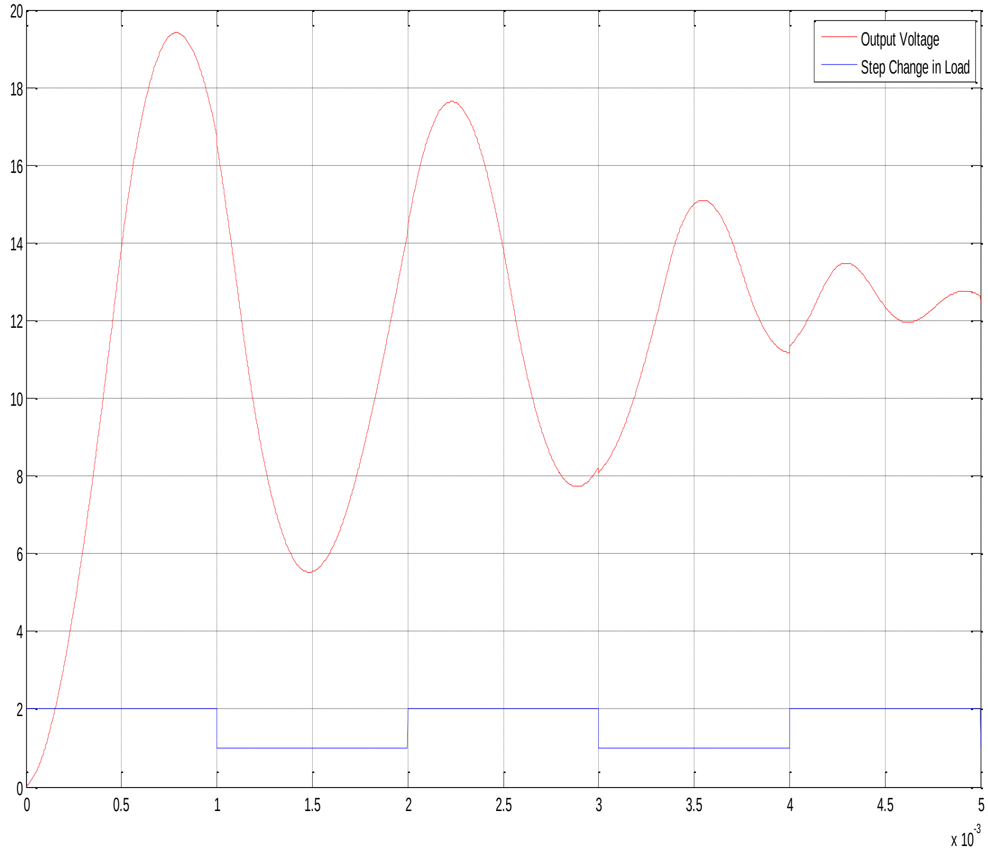

Figure 5.3 Response to 1A change in Load current with PI Compensated Buck Converter

5.2 Results With PID Compensated Buck Converter

5.2.1 Response to 20v DC Power Source with PIDCompensated Buck Converter

The response is shown below with settling time of about 2.3ms and rise time of about 1.33ms.

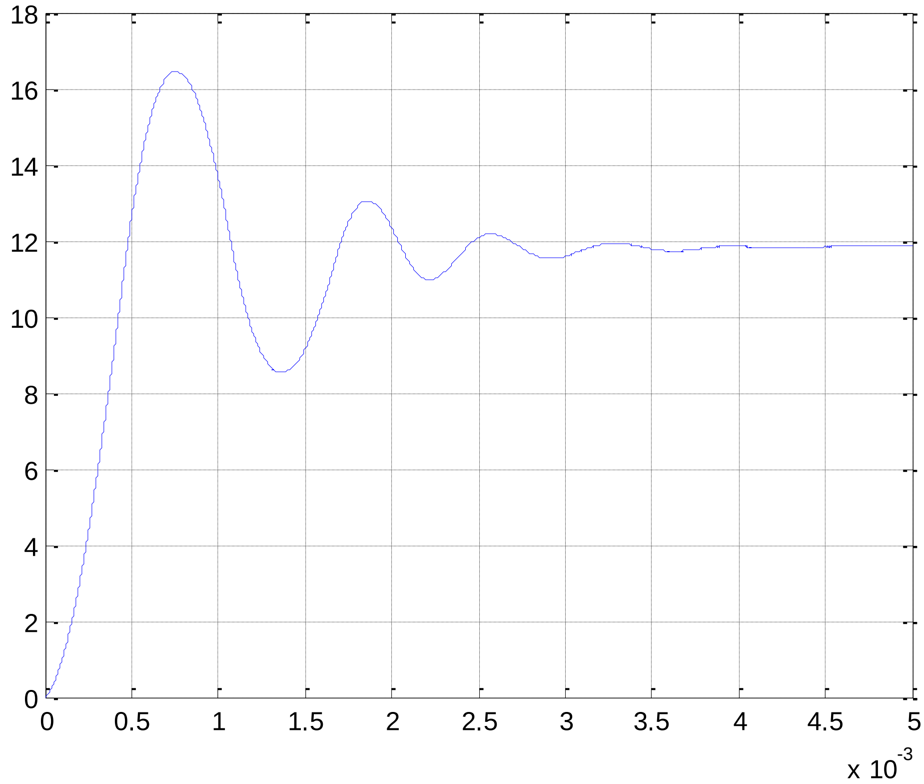

Figure 5.4 Response to 20v DC Power Source with PIDCompensated Buck Converter

5.2.2 Response to 1V change in Power source with PID Compensated Buck Converter

The response is shown below with settling time of about 2.8ms and rise time of about 5.88ms.

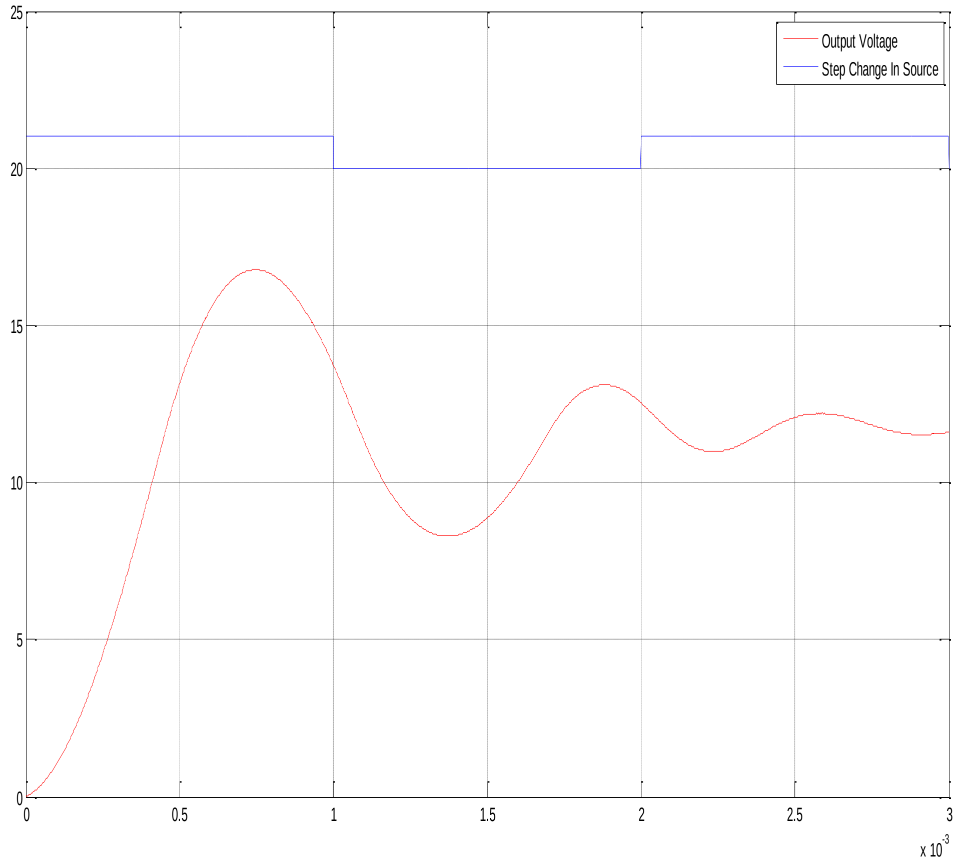

Figure 5.5 : Response to 1V change in power source with PID compensated Buck converter

5.2.3 Response to 1A change in Load current with PID Compensated Buck Converter

The response is shown below with settling time of about 3.3ms and rise time of about 3.94ms.

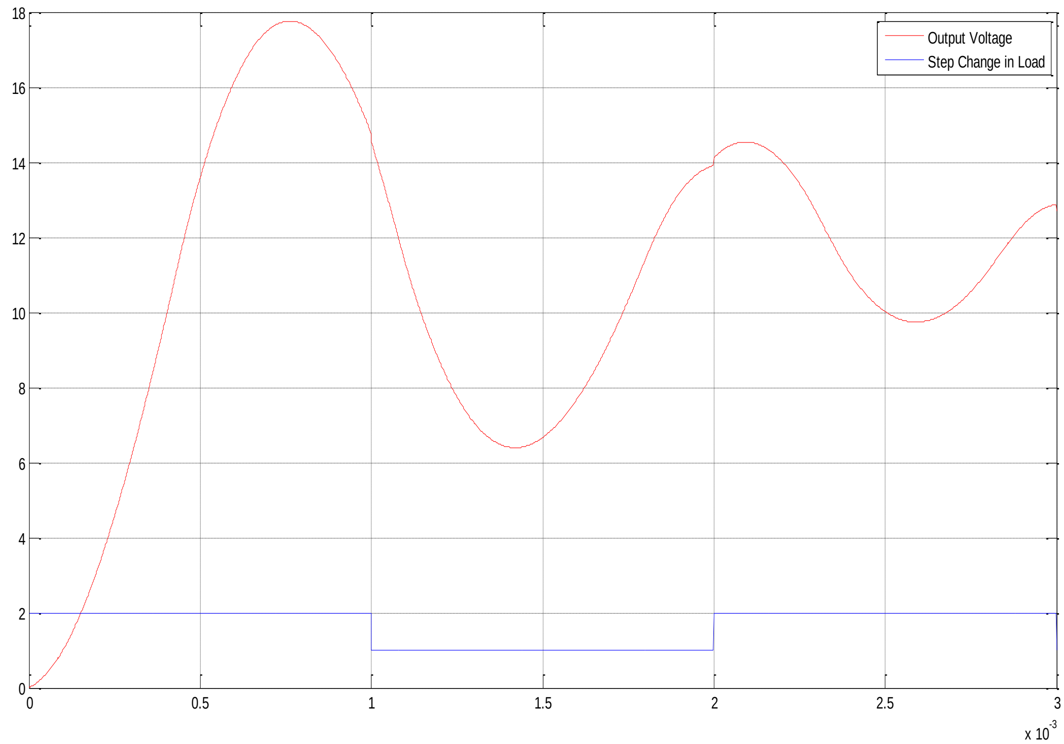

Figure 5.6 Response to 1A change in Load current with PID Compensated Buck Converter

5.3 Experimental Results for Fuzzy Controllers

5.3.1 Response to 20 V DC Power Source

The response of the open-loop system and the system compensated by a fuzzy logic PID controller for a 20 V DC power source can be seen in Figure 5.7. Both responses have zero steady-state error since the initial condition of the duty-cycle, D is 0.6, is chosen so that is met. The open-loop response has a maximum overshoot of 70 percent while the closed-loop response has a maximum overshoot of 6.5 percent. In addition, the settling time has been reduced from 2.75 msec to 1.10 msec. However, the rise time has been increased from 0.2 msec to 0.6 msec.

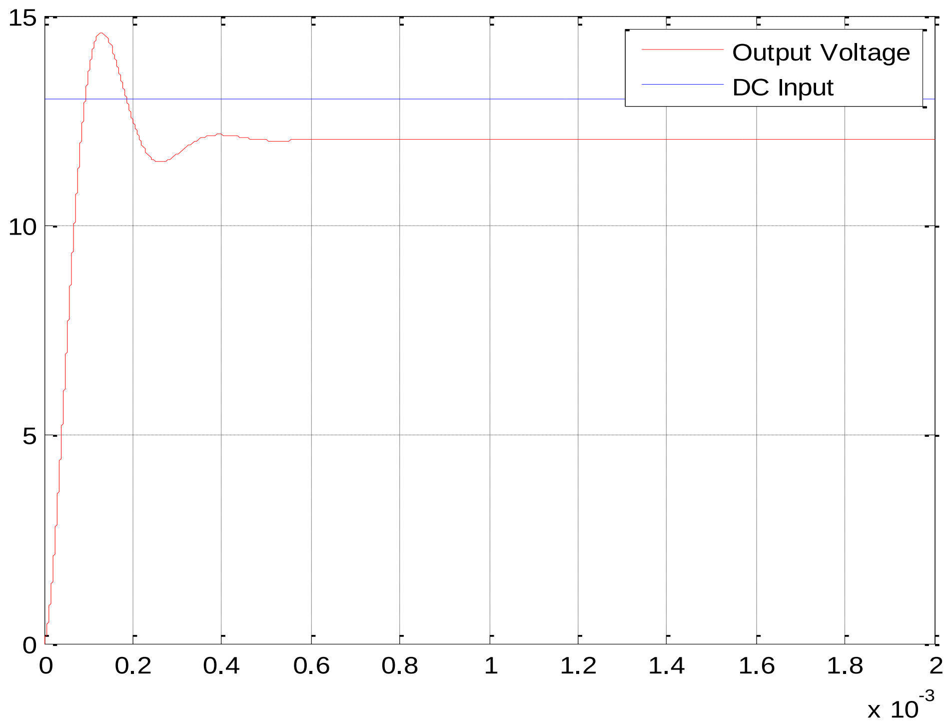

Figure 5.7 Response to 20 V DC Power Source

5.3.2 Response to 1 V Step Change in Power Source

The time response of the closed-loop system compensated by a fuzzy logic PID controller for a

1 V step change in the source can be seen in Figure 5.8. The figure shows that the steady-state error improves from 8 percent to less than 1 percent for the simulation. The steady-state error of the compensated system eventually decreases to zero because of the integral action. In addition, the maximum overshoot improves from 6 percent to less than 1 percent.

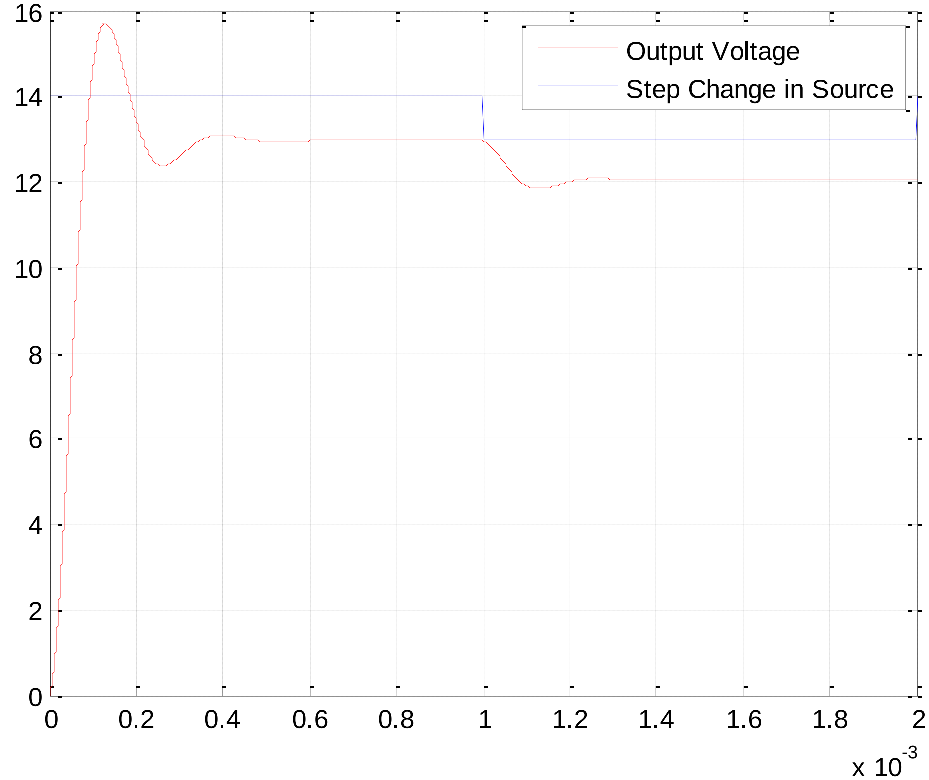

Figure 5.8 Response to 1 V Step Change in Power Source

5.3.3 Response to 1 A Step Change in Load Current

The time response of the closed-loop system compensated by a fuzzy logic PID controller to 1 A step change in the load can be seen in Fig 5.9. It can be seen from this figure that the steady-state error decreases from 1 percent to zero. In addition, the maximum undershoot improves from 12 percent to 3 percent and the maximum overshoot improves from 7 percent to 2 percent.

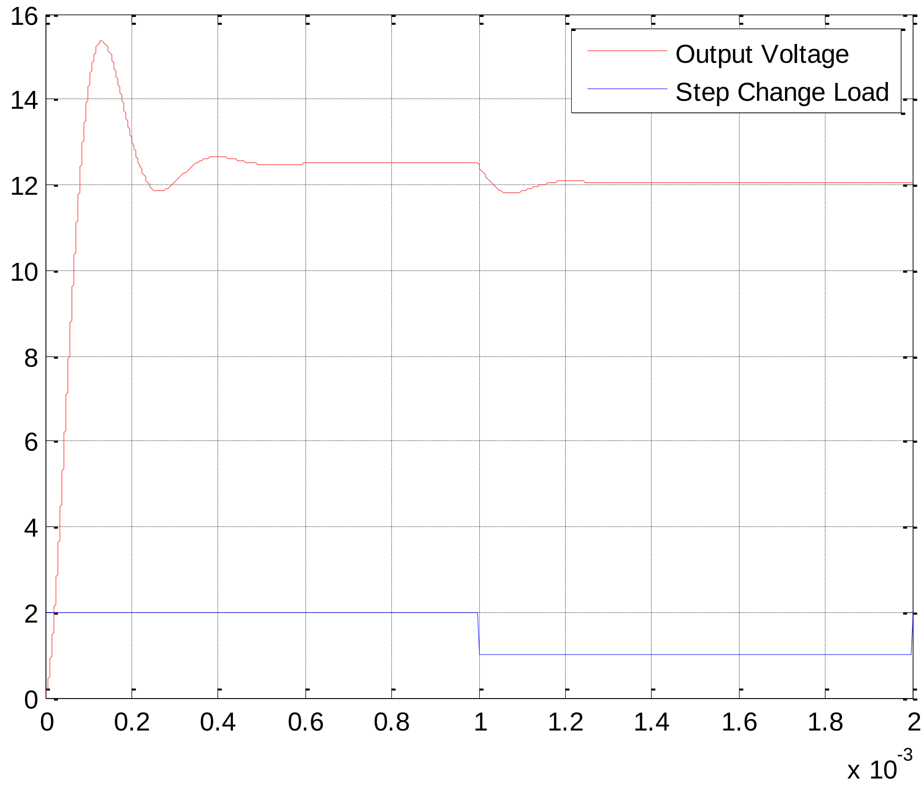

Figure 5.9 Response to 1 A Step Change in Load Current

6. Conclusions

Current trends in electronics require operation at low voltages with higher currents. In order to supply power to these electronics, PWM power converters are required which are able to deliver these voltages and currents efficiently. One such converter capable of delivering these requirements is the buck converter. The circuit is capable of generating lower voltages with a more median duty cycle.

In addition, digital control schemes are replacing the use of analog control schemes when controlling PWM power converters. Digital schemes, implemented through the use of DSPs, offer immunity to component variations, digital system compatibility, and the ability to incorporate advanced control schemes which is not available in analog counterparts.

One advanced control scheme which can be implemented with DSP is fuzzy logic control. Fuzzy logic control is a nonlinear control which control the duty cycle. Control of the duty cycle, in turn, controls the output voltage of the system. the time-domain response of the closed-loop system is improved with respect to the open-loop system. The overall speed of the system is also increased, as seen by the decrease of the settling time when the converter is connected to the power source. The system is also capable of fully rejecting disturbances by reducing the steady-state error to zero for a connection to a 20 V DC power source, a 1 A step change in load current, and 1 V step change in the power source voltage. In addition, the stability of the system is improved by reducing the overshoot/undershoot in all simulations. In all, the overall performance of the system is improved compared with the PI and PID controller.

Further investigation of the buck converter requires modeling and running a co-simulation between MATLAB/Simulink and circuit modeling software; either Synopsys Saber or Orcad PSpice. In addition, further investigation of the control of the buck converter required investigation into current-mode control and using fuzzy logic PID to regulate it. Finally, to test the control and circuit designs, the circuit should be constructed from the corresponding components, a programmed DSP provide the fuzzy logic controller, and the entire system tested to ensure functionality.

6.1. Future Work

In future work, the system will focus on how to optimize the digital controller, which includes: reducing the number of the instructions to increase the achievable sampling frequency, implementing the synchronization of the sampling and switching cycle, tuning up the coefficients of the compensator and increasing the bandwidth of the low-pass filter to improve the transient response of the system.

A more complicated and nonlinear algorithm will be attempted in DSP to improve system performance, which is generally difficult to implement in analog circuits. This algorithm may include adaptive and predictive control or a control based upon efficiency peaking, thermal management and other factors.

The DSP can also play an important role in managing, monitoring and testing of the power system. DSP can be used to execute data acquisition, signal conditioning, filtering, spectral analysis and transient capture, and it has the capability to monitor and control systems concurrently. This is particularly useful in adaptive control because the controller can dynamically change the structure or parameters in real time in response to variations in system behavior. Testing capabilities built into power electronics systems are used to identify system parameters for automatic tuning of controller gains and to locate faults in the event of a failure. Such systems facilitate testing by performing a selected acquisition of data and recording the stimulated system responses.

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