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Analysis and Design of Pulse-Width Modulated Two-Switch Forward DC-DC Converter for Universal Laptop Adapter

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ANALYSIS AND DESIGN OF PULSE-WIDTH MODULATED TWO-SWITCH FORWARD DC-DC CONVERTER FOR UNIVERSAL LAPTOP ADAPTER

A thesis submitted in partial fulfillment of the requirements for the degree of Master of Science in Engineering

By

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B. Tech., Nalanda Institute of Engineering and Technology, India, 2008

2011
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I HEREBY RECOMMEND THAT THE THESIS PREPARED UNDER MY SUPERVISION BY Venkata Sai Aditya Kumar Choragudi ENTITLED Analysis and Design of a PWM Two-Switch Forward DC-DC Converter for Universal Laptop Adapter BE ACCEPTED IN PARTIAL FULFILLMENT OF THE REQUIREMENTS FOR THE DEGREE OF Master of Science in Engineering

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Abstract


The objective of this research is to analyze, design and simulate a two-switch forward pulse-width modulated (PWM) DC-DC converter. The Forward PWM DC-DC converter is originally derived from the buck converter by the addition of a transformer. The transformer is mainly used as a safety feature to electrically isolate input and output stages of the power converter and also to scale the voltage or current. The steady-state analysis is presented for the two-switch PWM DC-DC forward converter in continuous conduction mode (CCM). Based on the analysis, the design equations for two-switch PWM forward DC-DC converter are derived. A design example is given. Furthermore, a detailed procedure to design a high-frequency (HF) two-winding power transformer is presented for CCM. The 60 W/ 20 V two-switch forward converter is designed for universal laptop adapter. Simulation results are provided to validate the theoretical analysis.
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1 Background

1.1 Types of Power Supplies

Advancements in semiconductor fabrication and its proliferation in power semiconductor based power supplies is making the size of the energy conversion systems ever decreasing [1] - [5]. Majority of electronic equipments encompassing consumer, industrial, automotive applications is powered from a power source [4] - [6]. These power sources can either be an energy storage devices or power supplies which run off the utility lines. The DC supply voltage should be altered to appropriate level in order to match the load requirement. Power supply system using linear voltage regulator is called linear power supply and the power supply system using a switched-mode regulator is called switched-mode power supply system (SMPS) [1] - [7].

![Diagram of AC-DC power supply system. (a) With a linear voltage regulator. (b) With a switching regulator.](image)

Figure 1.1: AC-DC power supply system. (a) With a linear voltage regulator. (b) With a switching regulator.
1.1.1 Linear Power Supply

Figure 1.1(a) [1], [7] shows the block diagram of linear power supply system which outputs controlled DC using a linear voltage regulator. It is observed from the Figure 1.1(a) that, the AC is stepped down using a low-frequency (LF) line-transformer. This stepped down voltage is rectified and then filtered to get a DC voltage. At this stage, the resultant DC is uncontrollable. The nominal ac line voltage is 220 Vrms in India and Europe, and 110 Vrms in US [1]. The frequency of the line voltage is 50 Hz in India and 60 Hz in USA [1]. As the frequency of the ac line voltage is low, the transformer is heavier.

1.1.2 Switched-Mode Power Supply (SMPS)

Figure 1.1(b) [1] shows the power supply unit using a switched-mode regulator which rectifies the AC directly and then filters it to get a DC voltage. At this stage, the raw DC is unregulated and its level can fluctuate depending on the grid it is fed from [1]. This unregulated DC is fed to the switching regulator, which consists of fast semiconductor devices such as power MOSFETs and IGBTs, high-frequency (HF) transformer (in some cases), output rectifier and a filter. The MOSFETs are turned ON and OFF by a driver at a certain frequency typically ranging from 20 to 200 kHz and variable duty-ratio. Therefore, a pulsating voltage is appeared at the primary winding of the transformer with certain voltage and duty-ratio and as a consequence, to the secondary winding [3]. This voltage pulse train is rectified and then smoothened by an output filter [3]. Since, this power conversion scheme does not utilize an LF transformer, it is not relatively bulkier. Because, the switching frequency is much higher (by several order of magnitude) than the AC line frequency, the size and weight of the passive components are greatly reduced [1] - [7].
1.2 Advantages of SMPS over Linear Power Supplies

SMPS regulators are preferred to their linear counter parts for the following benefits:

- Since the semiconductor devices are either operated completely in saturation region or in the cutoff region, there will be less power dissipated compared to linear power supplies [19].

- SMPS uses semiconductor devices such as MOSFETs and IGBTs as switching devices. Linear power supplies uses BJTs and resistors to obtain controlled DC at the output. As a result, efficiency of a SMPS can exceed 90% where as it is 40-50 % for typical linear power supplies [3].

- Because the magnetic components like transformer, inductor and passive components like capacitor operate at very high frequencies ranging from 20 kHz to 1 MHz, the size and volume of these components are much smaller compared to the linear power supply that operate at 50 Hz. The power density of SMPS can exceed 20 times than the linear power supplies for same specifications [3].

With the focus on realizing lighter and smaller electronic equipment such as laptop computers, cellular phones and other portable devices, the importance of realizing higher energy density power converters has grown rapidly. The advancements in semiconductor based microcontrollers (DSP based or FPGAs) has enabled to implement necessary control mechanisms associated power supplies. Therefore, SMPS has become an integral part in present day power supplies.

1.3 DC-DC Converters in Power Electronics

The need for dc-dc converters for residential, industrial and commercial applications has grown extensively [19]. The underlying principle of switch-mode power conversion is the utilization of power semiconductors and magnetic components arranged in a
specific configuration and operating them by means of appropriate control strategies [19]. The four categories of converters in power electronics are 1) Rectifiers, which convert ac to dc, 2) Inverters, which convert dc to ac, 3) Choppers, which converter dc to dc, and 4) Cyclo-converters, which changes the frequency of the input signal to desired frequency. Some applications [22] of DC-DC converters are where, a 24 V battery in a truck should be stepped down to 12 V to charge a cellular phone. A 12 V battery in a car will be stepped down operate a CD/DVD player which run at 3 V. On a computer mother board, 6 V / 5 V voltage pin is stepped down to operate other chips that run at 2 V or so. A 110 V AC power line should be rectified and stepped down to several DC voltages to operate computer accessories which run at different voltage levels. A 6 V battery is stepped up to 100 V or more for testing the insulation voltage [22].

The goal in any converter is to modify voltage/current level associated with input power supply to match the load requirements.

1.4 Classification of DC-DC Converters

There are various types of dc-to-dc converters which are more suitable for specific applications than the other [22]. Some basic non-isolated converters are buck, boost, buck-boost converters [22]. Some isolated converters are derived from non-isolated converters. For example, forward converter which is an isolated converter is derived from buck converter by inserting a transformer. Similarly, flyback converter which is an isolated converter, is derived from buck-boost converter. Other classification, on the other hand, can be made from the step-up or step-down function, the converter does. Buck, flyback, forward converters are some of the step down converters. Boost converter is an example of a step-up converter [22].
1.5 Motivation

The PWM two-switch forward converter is proven to have some advantages than the conventional three winding forward converter [1], [8] - [13] mainly because of the reduced transistor voltage stress and the two diodes that effectively clamps the ringing, caused by the output capacitance of the transistors and the transformer leakage inductance to the maximum input voltage. A detailed analysis in steady-state and complete design methodology are not available in the literature. As the successful operation of transformer is crucial in two-switch forward converter, a thorough design of HF transformer is also necessary, which is not yet addressed in the literature. A detailed steady-state analysis is essential and can also aid the future researchers interested in this topology.

1.6 Objectives

1. To present a brief review on the analysis of a conventional forward PWM DC-DC converter and specifically highlight its drawbacks and some means to overcome few of them.

2. To present a detailed analysis of two-switch DC-DC forward converter in its steady state, operating in CCM.

3. To present a step-by-step procedure to design a HF two-winding forward transformer in CCM using the Area-product method.
2 Brief Review of Forward PWM DC-DC Converter

2.1 Background

A PWM forward converter is a widely used SMPS converter, which produces an isolated and controlled DC output voltage. When compared to flyback converter, forward converter is energy efficient and can be used for applications requiring output power between (100-500) W [1]. This chapter presents a brief overview on the analysis of forward converter in steady-state including the maximum voltage and current stresses the devices has to withstand [1] - [4], [9] - [13].

2.2 Circuit Description

The basic circuit configuration of forward converter without parasitic components is shown in the Figure 2.1. The switch $M$ is an n-channel power MOSFET switch. The magnetizing inductance $L_{mag}$ is referred to the primary side of the ideal transformer $T$. The two rectifier diodes connected to the transformer secondary winding are represented by $D_{r1}$ and $D_{r2}$ respectively. The filter inductor $L_f$ and the filter capacitor $C_f$ are connected after the rectification stage to suppress the higher order harmonics. There is a tertiary winding coupled to the primary winding of the transformer to

![Figure 2.1: A PWM DC-DC forward converter](image_url)
pump the magnetizing energy back to the source voltage $V_S$, through the tertiary diode $D_t$. $R_L$ is the load resistance. The duty-ratio $D_r$ is the fraction of on-time of the switch to the total switching period $T_s$.

### 2.2.1 Significance of Forward Converter

The significance of forward converter in modern world is due to its simple operation and robustness. Some of them are listed below.

1. The transformer used in the forward converters has the flux waveform symmetrical to the positive and negative halves despite of the abrupt load variations which makes the transformer to operate below saturation.

2. Forward converter provides galvanic isolation between the input and output since there is no energy transfer for a significant portion of switching cycle.

3. Unlike flyback converter which uses the transformer magnetizing inductance to drive the load, forward converter instead, uses a filter inductor which makes its transformer less heavier.

### 2.3 Analysis of Forward Converter in Steady State

We make some assumptions to make the analysis of forward converter simple [1]. The transformer leakage inductance, stray capacitances are neglected. The MOSFET switch $M$, the two rectifier diodes $D_{r1}$, $D_{r2}$, and the tertiary diode $D_t$ are assumed to be ideal. The passive components are assumed to be linear, and frequency independent [1]. The basic circuit operation of the ideal forward converter shown in the Figure 2.1 can be divided into three stages.

#### 2.3.1 Time Interval $0 < t \leq D_r T$

Figure 2.2(a) [1] shows the ideal equivalent circuit of the forward converter during the time interval $0 < t \leq D_r T$. The MOSFET switch $M$ and the rectifier diode $D_{r1}$,
in this time interval are turned ON. The rectifier diode $D_{r2}$ and tertiary diode $D_t$ are reverse biased. The primary winding of the ideal transformer and the magnetizing inductance $L_{mag}$ have the same voltage that can be expressed as [1] - [4], [9] - [13]

$$v_p = v_{Lmag} = V_S = L_{mag} \frac{di_{Lmag}}{dt}.$$  

(2.1)

From equation (2.1), the magnetizing inductance $L_{mag}$ has a current,

$$i_{Lmag} = \frac{1}{L_{mag}} \int_0^t v_{Lmag} \, dt = \frac{1}{L_{mag}} \int_0^t V_S \, dt = \frac{V_S}{L_{mag}} t,$$  

(2.2)

where its initial value is zero [1]. The voltage across the filter inductor $L_f$ is [1]

$$v_{Lf} = \frac{V_S}{n_p} - V_O = L_f \frac{di_{Lf}}{dt}. \quad (2.3)$$

From equation (2.3) the current through the filter inductor $L_f$ and the rectifier diode $D_{r1}$ is given by

$$i_{Dr1} = i_{Lf} = \frac{1}{L_f} \int_0^t v_{Lf} \, dt + i_L(0) = \left(\frac{\frac{V_S}{n_p} - V_O}{L_f}\right) t + i_{Lf}(0). \quad (2.4)$$

Therefore, the current through the transformer primary winding is given by [1]

$$i_p = i_s = \left(\frac{\frac{V_S}{n_p} - V_O}{n_p L_f}\right) t + \frac{i_{Lf}(0)}{n_p}, \quad (2.5)$$

and the switch current can be written as [1],

$$i_M = i_p + i_{Lmag} = \left(\frac{\frac{V_S}{n_p} - V_O}{n_p L_f}\right) t + \frac{i_{Lf}(0)}{n_p} + \frac{V_S}{L_{mag}} t. \quad (2.6)$$

The voltage across the rectifier diode $D_{r2}$ is [1]

$$v_{Dr2} = -v_{Lf} - V_O = -\frac{V_S}{n_p}. \quad (2.7)$$

From equation (2.7) the voltage across the tertiary winding is

$$v_t = -\frac{n_t}{n_p} V_S. \quad (2.8)$$
The voltage across the tertiary diode $D_t$ is [1]

$$v_{Dt} = - \left( \frac{n_t}{n_p} + 1 \right) V_S. \quad (2.9)$$

This interval is terminated as soon as the MOSFET $M$ gets turned OFF by a pulse driver.

Figure 2.2: Ideal equivalent circuits of forward converter operating in different time intervals. (a) $0 < t \leq D_r T$. (b) $D_r T < t \leq (D_r T + t_m)$. (c) $(D_r T + t_m) < t \leq T$. 
2.3.2 Time Interval \( D_r T < t \leq (D_r T + t_m) \)

The ideal equivalent circuit for the forward converter during this time interval is shown in the Figure 2.2(b). The MOSFET switch \( M \) and the rectifier diode \( D_{r1} \) are turned OFF, the diodes \( D_{r2} \) and \( D_t \) are turned ON. The filter inductor \( L_f \) has a voltage \[ v_{L_f} = -V_O = L_f \frac{d i_{L_f}}{dt}. \] (2.10)

From equation (2.1), the primary winding of the transformer and magnetizing inductance \( L_{mag} \) have the same voltage \[ v_p = v_{L_{mag}} = -\left( \frac{n_p}{n_t} \right) V_S = L_{mag} \frac{d i_{L_{mag}}}{dt}. \] (2.11)

From equation (2.11)
\[
i_{L_{mag}} = \frac{1}{L_{mag}} \int_{D_r T}^{t} v_{L_{mag}} dt + i_{L_{mag}}(D_r T) = -\frac{n_p V_S}{n_t L_{mag}} (t - D_r T) + \frac{D_r V_S}{f_s L_{mag}},
\] (2.12)
and
\[
i_{D_t} = \left( \frac{n_p}{n_s} \right) i_{L_{mag}} = -\frac{n_p^2 V_S}{n_t^2 L_{mag}} (t - D_r T) + \frac{n_p D_r V_S}{n_t f_s L_{mag}}.
\] (2.13)

The MOSFET switch voltage is \[ v_M = \left( \frac{n_p}{n_t} + 1 \right) V_S. \] (2.14)

The ideal waveforms of the forward converter is shown in Figure 2.3 [1]. Figure 2.4 and 2.5 shows the simulated waveforms of single-switch forward converter. The simulation data is provided.

2.3.3 Time Interval \( (D_r T + t_m) < t \leq T \)

Figure 2.2(c) shows the forward converter ideal equivalent circuit during the time interval \( (D_r T + t_m) < t \leq T \). During this time interval, the MOSFET switch \( M \), and the diodes \( D_{r1} \), and \( D_t \) are turned OFF, the rectifier diode \( D_{r2} \) is turned ON. The two
The voltage across the MOSFET is [1]

\[ v_M = V_S, \]  
(2.15)

and the tertiary diode voltage is [1],

\[ v_{Dt} = -V_S. \]  
(2.16)

The voltage across the filter inductor \( L_f \) is given by [1]

\[ v_{L_f} = -V_O = L_f \frac{di_{L_f}}{dt}, \]  
(2.17)
2.4 Voltage and Current Stresses Across the Components

The voltage and current through the MOSFET have maximum values [1]

\[ V_{MM_{max}} = \left( \frac{n_p}{n_t} + 1 \right) V_{S_{max}} = \frac{V_{S_{max}}}{1 - D_{rMAX}} \] (2.18)

and

\[ I_{MM_{max}} = I_{Dr1MAX} + \Delta i_{LMAG(max)} = I_{O_{max}} + \frac{D_{r_{max}}}{n_p} + \Delta i_{LMag(max)}. \] (2.19)

The maximum peak voltage of the rectifier diode \( D_{r1} \) is [1]

\[ v_{Dr1M_{max}} = \frac{V_{S_{max}}}{n_t}. \] (2.20)

The peak voltage across the rectifier diode \( D_{r2} \) has a peak value from equation (2.7) is

\[ v_{Dr2M_{max}} = \frac{V_{S_{max}}}{n_p}. \] (2.21)

The two rectifier diodes \( D_{r1} \) and \( D_{r2} \) has same maximum value of peak currents given as [1]

\[ I_{Dr1M_{max}} = I_{Dr2M_{max}} = I_{O_{max}} + \frac{\Delta i_{LMag(max)}}{n_p}. \] (2.22)

The tertiary diode \( D_t \) has a maximum peak voltage from equation (2.9) is

\[ v_{DtM_{max}} = \left( \frac{n_t}{n_p} + 1 \right) V_{S_{max}} = \frac{V_{S_{max}}}{D_{rMAX}}, \] (2.23)

and its peak current can be expressed as [1]

\[ I_{DtM_{max}} = \frac{n_p}{n_t} \Delta i_{LMag(max)} = \left( \frac{n_p}{n_t} \right) \frac{D_{r_{min}} V_{S_{max}}}{f_s L_{mag}}. \] (2.24)

As we can see from equations (2.23) and (2.24), the peak voltage and the peak current in the tertiary diode \( D_t \) depends on the ratio \( \frac{n_t}{n_p} \).
2.4.1 Simulation Results

A single switch forward converter is simulated for the maximum output power $P_{O_{\text{max}}} = 60 \text{ W}$, output voltage $V_O = 20 \text{ V}$, input source voltage $V_S = 110 \text{ V}$, switching frequency $f_s = 100 \text{ kHz}$ and duty-ratio $D_{r_{\text{max}}} = 3.4 \mu\text{s}$ which has a time period of $T_s = 10 \mu\text{s}$.

The n-channel power MOSFET switch is chosen to be an international rectifier IRFPF30 which has maximum voltage $V_{DSS} = 900 \text{ V}$, maximum current $I_D = 3.6 \text{ A}$, and on-state resistance of $r_{DS} = 3.7 \Omega$ [27].

An ultra fast diode rectifier MUR890 is chosen as the tertiary diode $D_t$ which has a maximum DC blocking voltage of $V_R = 900 \text{ V}$, maximum forward current of $I_{F(AV)} = 8.0 \text{ A}$, forward voltage of $V_F = 1.8 \text{ V}$ [28]. The two rectifier diodes on the secondary side of the transformer is chosen to be MBR20200 high voltage schottky power diodes with $V_{DSS} = 200 \text{ V}$, $I_F = 20 \text{ A}$ and forward voltage $V_F = 0.95 \text{ V}$ [25]. The values of filter inductor and filter capacitor are $L_f = 300 \mu\text{H}$ and $C_f = 20 \mu\text{F}$. Load resistance $R_L = 6.67 \Omega$.

The transformer chosen is a three-winding linear transformer in saber sketch which has the same primary and tertiary turns $N_p = N_t = 22$. The secondary winding has $N_s = 11$ turns. The magnetizing and the primary winding inductance is $L_{mag} = L_p = 3 \text{ mH}$, and the secondary winding inductance $L_s = 1 \text{ mH}$.
Figure 2.4: Simulation results of a PWM single-switch forward DC-DC converter-1.
Figure 2.5: Simulation results of a PWM single-switch forward DC-DC converter-2.
3 Problem of Tertiary Winding in Forward PWM DC-DC Converter

3.1 Introduction

The conventional forward converter uses a tertiary winding and a diode to reset the transformer core after the MOSFET switch $M$ is turned off [7] - [12]. The tertiary winding is connected in opposite polarity to the magnetizing inductance $L_{mag}$ to give back the unwanted energy stocked in the core of the transformer to the input voltage source $V_S$. This chapter deals with the problems associated with the tertiary winding mechanism of core reset. Also, this chapter addresses some existing snubber circuit solutions (some of them are not fully explored) used to reset the core. Finally, this chapter makes a brief performance comparison of all the presented mechanisms.

3.2 Problems Associated with Tertiary Winding

1. As soon as the MOSFET switch $M$ is open [8], there exists a sudden change of flux in the windings of the transformer. According to Lenz law, the change of flux will generate theoretically infinite amount of negative voltages across the dot terminal of the transformer windings. Therefore, the tertiary diode $D_t$ should start conducting right after the turn off of the MOSFET $M$. But in reality, it acts like an anti-reset circuit due to its high impedance. Therefore sometimes a fastening capacitor $C$ can be used to speed-up the reset process [8].

2. The coupling between the primary and the tertiary winding must be very good because the tertiary winding has to give back the entire magnetization energy of the transformer to the source. For this, the primary and the tertiary windings are wound together known as bifilar wound. These bifilar windings should withstand high voltage stress and expensive compared to normal transformer wires.
3. The use of tertiary winding method of reset has a direct impact on duty-ratio. We know that the maximum duty-ratio [1]

\[ D_{r,\text{MAX}} = \frac{1}{n_t/n_p + 1}. \]  

(3.1)

As the ratio \( n_t/n_p \) increases from 0.25 to 4, \( D_{r,\text{MAX}} \) decreases from 0.8 to 0.2 [1]. For \( n_t = n_p \), \( D_{r,\text{MAX}} = 0.5 \) and the peak switch voltage [1]

\[ V_{MM} = \left( \frac{n_t}{n_p} + 1 \right) V_S = \frac{V_S}{1 - D_{r,\text{MAX}}}. \]  

(3.2)

As the ratio \( n_t/n_p \) increases from 0.25 to 4, \( V_{MM} \) increases from 1.25\( V_S \) to 5\( V_S \) [1]. Thus, with the decrease in the number of turns of the tertiary winding, there will be a simultaneous increase in the duty-ratio. At the same time, the voltage stress across the switch \( V_{MM} \) increases which demands to use the MOSFET of higher rating. High rating MOSFET have high \( r_{DS} \) increasing the conduction losses. The overall cost of the converter also increases. Therefore, we have to limit the converter duty-ratio to 50% for optimum performance [1].

4. The use of a magnetic component, i.e., winding as a core-reset, increases the risk of EMI problems. This effect is adverse at significantly high switching frequencies, usually above 100 kHz [8].

5. Magnetic effects are not easily predictable. For example, if the conduction of tertiary winding does not start because of its high impedance as the switch turns OFF, the negative voltage developed across the transformer dotted terminals may result in the failure of the whole converter [8].

6. The magnetic components like filter inductor, transformer and heat sinks will occupy more than 50 % of the total volume of the converter [2]. The tertiary winding still increases the size of the transformer [2].
The problems listed above arise in the conventional forward converter that uses tertiary winding and a diode as the core reset. Designing a reset circuit which overcomes most of the aforementioned problems is the main challenge for a power electronic designer at the industry level. There are several snubber circuits which can overcome some of the problems with tertiary winding. The existence of number of snubber circuits itself indicates that there is no particular circuit which can solve all the problems associated with the single-switch forward converter [8]. The following section presents various snubbers and their drawbacks.

### 3.3 Various Snubbers of Forward Converter

This section briefly presents the various types of forward converter snubbers used to transfer the energy present in the transformer core. As the operation of the conventional forward converter has already been discussed, the rest of the topologies are presented here. The main focus is to analyze the way the transformer core is being reset. However, the secondary side of the converter operation remains unaltered in every presented topology. To make the analysis less tedious, switch parasitic capacitances and transformer leakage inductances are neglected in some of the topologies.

#### 3.3.1 Using an \( RCD \) Snubber

![Figure 3.1: The Forward Converter employing \( RCD \) Snubber.](image-url)

Figure 3.1 shows the forward converter using an \( RCD \) snubber for core-reset [8].
When the MOSFET switch $M$ is open, the snubber diode $D_S$ will be forward biased and begins to charge the capacitor $C_S$ thereby de-magnetizing the core. The capacitor clamps the voltage across the primary winding to $D_r V_S$, where $D_r$ is the duty-ratio of the pulse. The voltage stress across the switch $v_M = 2V_{S\text{max}}$. The simplest way of resetting the transformer is accomplished by $RCD$ snubber. As the number of components required is less, the overall cost of the snubber is low. One disadvantage in this technique is the resistor $R_S$. Besides creating thermal problem, it dissipates power, reducing the efficiency. This effect is clearly observed if this scheme is used for applications requiring high input voltage [8].

### 3.3.2 Using an $LCDD$ Snubber

![Figure 3.2: The Forward Converter with $LCDD$ Snubber.](image)

Figure 3.2 shows the forward converter topology implementing $LCDD$ snubber. As soon as the MOSFET switch $M$ is open, the capacitor $C_r$ will be charged and clamped to $D_r V_S$ which is same as like in $RCD$ snubber. The way capacitor gets discharged makes the difference. Instead of using a resistor which dissipates the power, an inductor $L_r$ is used. [8] After the capacitor $C_r$ is fully charged, $LC$ resonance is formed between the inductor $L_r$ and diode $D_{S2}$ and a capacitor $C_r$. The maximum voltage stress across the main MOSFET $M$ is $2V_{S\text{max}}$ [8]. The disadvantages of this method are as follows [8]:

[8]
1. The modes of operation of this circuit are more because of the resonant behavior during the core reset. As a result, more number of iterations during the design are needed [8].

2. The conduction loss is added in the circuit because of the high-frequency resonance between $L_r$ and $C_r$. This loss can easily diminish the saving of the power loss caused by $R_S$ from $RCDD$ snubber. Therefore, improvement in efficiency is not substantial [8].

3. If the converter is operated at higher input voltages, the size of the inductor $L$ needed could be large [8].

### 3.3.3 Using a Resonant Reset

![Figure 3.3: The Forward Converter with Resonant reset.](image)

Figure 3.3 shows the forward converter topology implementing resonant reset [8].

[8] This topology makes use of the parasitics to reset the core. As soon as the main MOSFET switch $M$ is open, the magnetizing inductance of the transformer $L_m$ and the switch output capacitance $C_{p1}$ resonates and automatically resets the transformer. The voltage stress across the MOSFET is $v_M = 2V_S$. Additional capacitor is needed to obtain a well-defined resonant frequency is one major disadvantage in this approach [8].
3.3.4 Using a Switching Snubber

Figure 3.4: The Forward Converter with Switching Snubber.

Figure 3.4 shows the forward converter implementing switching snubber [8]. This topology makes use of another MOSFET and a capacitor $C_s$ to reset the transformer core [8]. As soon as the main MOSFET $M$ is turned OFF by a driver, the capacitor $C_s$ will be charged. To discharge the capacitor, the switch $S_S$ is turned on. The main switch $M$ has a voltage stress of $1.3V_S$ across it. While turn off, the capacitor $C_s$ clamps the voltage across the switch. Making use of negative switch current is the primary idea of this soft-switching scheme. The switch $M$ has a body diode which is made to conduct using the negative current and the switch in effect, can be turned ON with nearly zero voltage. The switch $M_S$ is turned on at 0.7 V and turned off at optimum conditions. Efficiency improvement is substantial [8]. The disadvantages of this topology are listed below:

1. This switching snubber requires an additional MOSFET, needless to say, its gate drive as well [8].

2. The magnetizing inductance used in this topology accounts for 30-50% increase in the overall conduction losses [8].

3. These kind of converters undergo sub-harmonic oscillations which prevents its use in industrial applications [8].
3.3.5 The Two-Switch Forward Converter: not fully explored

![Diagram of the Two-Switch Forward Converter]

Figure 3.5: The Two-Switch Forward Converter.

Figure 4.1 shows the forward converter with two MOSFET switches $M_1$ and $M_2$. This method uses two demagnetization diodes to reset the core of the transformer. The switches are simultaneously turned off, with the help of a gate driver. Then, the two demagnetization diodes $D_{c1}$ and $D_{c2}$ gets forward biased and give the magnetizing energy in the transformer back to the input voltage source $V_S$. The clamping diodes connects the drain terminals of the switches $M_1$ and $M_2$ to the source $V_S$. Therefore, even a slight increase in the drain voltage of the switches above $V_S$ makes the diodes to conduct clamping them exactly to $V_S$ [1].

3.4 A Brief Performance Comparison

With respect to various factors, the following section presents a brief comparision between the different core reset circuit topologies discussed above.

3.4.1 Component Stresses and Efficiency

As the two switches are clamped to the maximum input voltage, the two-switch topology is the best and efficient way of core reset when the voltage stress across the switches are concerned [8].
3.4.2 Small-Signal Dynamics

All of the above discussed topologies have small-signal dynamics similar to the conventional forward DC-DC converter which is the derivation from the buck converter [8].

3.4.3 Noise/EMI and Parts Count

Noise levels in two-switch topology and resonant reset topology is medium and the rest of the topologies have high noise/EMI. The forward topology with resonant reset has very less number of parts needed. Only one capacitor is needed and the whole circuit works by its parasitic capacitances and inductances [8].

The two-switch topology despite of its duty-ratio limitation at 50%, it has less voltage stress, low EMI and relatively high efficiency. A detailed analysis in steady-state, design equations, design procedure are not available in the literature. Therefore, a detailed analysis is made on the two-switch forward converter in its steady-state operating in CCM.
4 PWM Two-Switch Forward DC-DC Converter

4.1 Introduction

A PWM two-switch DC-DC forward converter operating in CCM is presented in this chapter [1]. Current and voltage equations including the stresses of all the components are presented [9] - [13]. The DC voltage transfer function for both loss less and lossy converter is derived [1]. The power losses across all the components including the gate drive of transistors are estimated [1]. A design example with universal input laptop adapter specifications is illustrated.

4.2 Circuit Description

A two-switch forward converter is depicted in Figure 4.1 [1]. The two MOSFET transistors $M_1$ and $M_2$ connected across the source voltage are n-channel power MOSFETs. The two demagnetization diodes $D_{c1}$ and $D_{c2}$ are connected across input voltage source and transformer primary winding. The two diodes on the transformer secondary side are the rectifier diodes denoted by $D_{r1}$ and $D_{r2}$ respectively. $V_S$ represents the source voltage and $R_L$ represents the load resistance. The switching period is represented by $T_s$, which the the reciprocal of the switching frequency $f_s$. The duty-ratio $D_r$ is defined as the fraction of the ON time of the switch $t_{ON}$ to the total time period $T_s$ [13]. The magnetizing inductance $L_{mag}$ is referred to the transformer primary side [1], [9] - [13].

4.3 Two-Switch PWM DC-DC Forward Converter : Steady-State Analysis

The operation of an ideal two-switch forward dc-dc converter neglecting the transformer and switch parasitic components can be divided into three stages. The two MOSFETs $M_1$ and $M_2$ using a pulse driver are turned ON simultaneously and the current through the magnetizing inductance $i_{L(mag)}$ begins to increase with a slope of
Figure 4.1: The Two-Switch PWM Forward DC-DC Converter.

$V_S/L_{mag}$ [1]. The rectifier diode $D_{r1}$ drives the load in this stage. When the switches are turned OFF at $t = D_rT$, the clamping diodes $D_{c1}$ and $D_{c2}$ gets forward biased and clamps the maximum voltage across the transistors to the DC input voltage $V_S$. This is the fine characteristic that makes the two-switch forward converter so popular. The magnetizing current $i_{L(mag)}$ decreases with a slope of $-\frac{V_S}{L_{mag}}$ [1]. On the secondary side of the transformer, the filter inductor current $i_{L_f}$ is diverted to the free-wheeling diode $D_{r2}$, which drives the load. When the magnetizing current falls to zero, the two demagnetizing diodes on the transformer primary side are turned OFF. The inductor current $i_{L_f}$ continues to drive the load through $D_{r2}$ till next switching cycle begins. Since the pulses driving the transistors are unipolar, the transformer core operates in the first quadrant of the hysteresis curve [21]. The two-switch forward converter is suitable to applications where the power level is ranging between 30-500 W [1], [9] - [13]. The following assumptions are made in analyzing the two-switch forward converter [1].

1. The parasitic capacitance and lead inductance of the diodes are assumed to be zero [1].

2. The leakage inductance and stray capacitance of the transformer are neglected [1].
3. The input voltage source $V_S$ is ideal with zero output impedance [1].

4.3.1 Time Interval $0 < t \leq D_r T$

During the time interval $0 < t \leq D_r T$, the two MOSFET switches are simultaneously turned ON by a pulse driver. Figure 4.2(a) shows the ideal equivalent circuit of two-switch forward converter during the time interval $0 < t \leq D_r T$. The relationship among the transformer voltages to primary and secondary turns ratio is [1]

$$v_p : v_s = N_p : N_s,$$  \hfill (4.1)

where $N_p$ and $N_s$ are the number of turns of the primary and secondary, respectively. When the two MOSFET switches $M_1$ and $M_2$ are ON, the magnetizing inductance $L_{mag}$ of the transformer and the primary winding have the same voltage given as [1]

$$v_p = v_{Lmag} = V_S = L_{mag} \frac{d i_{L(mag)}}{dt}.$$  \hfill (4.2)

The initial condition of the magnetizing inductance is $i_{L(mag)}$ is zero [1]. Therefore, from equation (4.2), its current can be written as [1, 9, 13],

$$i_{L(mag)}(t) = \frac{1}{L_{mag}} \int_0^t v_{L(mag)} \, dt = \frac{1}{L_{mag}} \int_0^t V_S \, dt = \frac{V_S}{L_{mag}} t.$$  \hfill (4.3)

The magnetizing current has a peak value at $t = D_r T$ and is given by [1]

$$\Delta i_{L(mag)} = i_{L(mag)}(D_r T) = \frac{V_S D_r T}{L_{mag}} = \frac{D_r V_S}{f_s L_{mag}},$$  \hfill (4.4)

from which

$$\Delta i_{L(m)(max)} = \frac{D_r V_S \Delta i_{L(m)(max)}}{f_s L_{mag}(min)}.$$  \hfill (4.5)

From equation (4.5) the minimum magnetizing inductance is

$$L_{mag}(min) = \frac{D_r V_S \Delta i_{L(m)(max)}}{f_s \Delta i_{L(m)(max)}},$$  \hfill (4.6)
where $\Delta i_{Lm(max)}$ is 10% of the maximum peak current of the ideal transformer primary $I_{pmax}$. From equation (4.1), the transformer secondary winding voltage can be expressed as

$$v_s = \frac{v_p}{n} = \frac{V_S}{n}.$$  \hspace{1cm} (4.7)

The filter inductor $L_f$ has a voltage of [1]

$$v_{L_f} = \frac{V_S}{n} - V_O = L_f \frac{di_{L_f}}{dt}. \hspace{1cm} (4.8)$$

So, from equation (4.8) the current through the rectifier diode $D_{r1}$ and the filter inductor $L_f$ is,

$$i_{Dr1} = i_{L_f} = \frac{1}{L_f} \int_0^t v_{L_f} \, dt + i_{L_f}(0) = \frac{(\frac{V_S}{n} - V_O)}{L_f} \, t + i_{L_f}(0). \hspace{1cm} (4.9)$$

Hence, the transformer primary winding current can be expressed as [1]

$$i_p = \frac{i_s}{n} = \frac{(\frac{V_S}{n} - V_O)}{nL_f} \, t + \frac{i_{L_f}(0)}{n}, \hspace{1cm} (4.10)$$

and the two MOSFET currents are given as [1], [9],

$$i_{M1} = i_{M2} = i_p + i_{L(mag)} = \frac{(V_S - V_O)}{nL_f} + \frac{i_{L_f}(0)}{n} + \frac{V_S}{L_{mag}} \, t. \hspace{1cm} (4.11)$$

The rectifier diode $D_{r2}$ has a voltage of [1]

$$v_{Dr2} = -v_{L_f} - V_O = -\frac{V_S}{n}. \hspace{1cm} (4.12)$$

The voltage across the two de-magnetization diodes $D_{c1}$ and $D_{c2}$ is [1]

$$v_{Dc1} = v_{Dc2} = -V_S. \hspace{1cm} (4.13)$$

At time $t = D_r T$, the switches are turned OFF terminating this time interval.
Figure 4.2: Equivalent circuits for different time intervals for PWM two-switch DC-DC forward converter. (a) $0 < t \leq D_r T$. (b) $D_r T < t \leq (D_r T + t_m)$. (c) $(D_r T + t_m) < t \leq T$. 
4.3.2 Time Interval $D_r T < t \leq (D_r T + t_m)$

The ideal equivalent circuit during $D_r T < t \leq (D_r T + t_m)$ is shown in Figure 4.2(b). On the primary side of the transformer, the two switches $M_1$ and $M_2$ are OFF, the clamping diodes $D_{c1}$ and $D_{c2}$ clamps the voltage across the two transistors to maximum input voltage. On the secondary side, the diode $D_{r1}$ is OFF and $D_{r2}$ is ON. The voltage across filter inductor $L_f$ is [1]

$$v_{L_f} = -V_O = L_f \frac{di_{L_f}}{dt}. \quad (4.14)$$

From equation (4.14) the current through the inductor $L_f$ and the diode $D_{r2}$ is
Figure 4.4: Simulated waveforms of the PWM two-switch DC-DC forward converter-1
Figure 4.5: Simulated waveforms of the PWM two-switch forward converter-2
found to be
\[ i_{L_f} = i_{D_{r2}} = \frac{1}{L_f} \int_{D_rT}^{t} v_{L_f} dt + i_{L_f}(D_rT) = -\frac{V_O}{L_f} \int_{D_rT}^{t} dt + i_{L_f}(D_rT) = -\frac{V_O}{L_f} (t-D_rT) + i_{L_f}(D_rT). \] (4.15)

The filter inductor peak-to-peak value of the ripple current is [1]
\[ \Delta i_{L_f} = i_{L_f}(D_rT) - i_{L_f} = \frac{V_O T(1-D_r)}{L_f} = \frac{V_O (1-D_r)}{f_s L_f}. \] (4.16)

The voltage across the transformer primary and the magnetizing inductance \(L_{mag}\) is [1]
\[ v_p = v_{L(mag)} = -V_S = L_{mag} \frac{d i_{L(mag)}}{dt}. \] (4.17)

From equation (4.17), the magnetizing current and current through the clamping diodes \(D_{c1}\) and \(D_{c2}\) can be expressed as,
\[ i_{D_{c1}} = i_{D_{c2}} = i_{L(mag)} = \frac{1}{L_{mag}} \int_{D_rT}^{t} v_{L(mag)} dt + i_{L(mag)}(D_rT) = -\frac{V_S}{L_{mag}} (t-D_rT) + \frac{D_r V_S}{f_s L_{mag}}. \] (4.18)

The maximum current of the two clamping diodes \(D_{c1}\) and \(D_{c2}\) occurs at \(t = D_r T\) which is given by [1]
\[ i_{D_{c1M}} = i_{D_{c2M}} = I_{D_{c1}}(D_rT) = \frac{D_r V_S}{f_s L_{mag}}. \] (4.19)

From equation (4.17), the voltages across the transformer secondary and the diode \(D_{r1}\) are
\[ v_s = v_{D_{r1}} = \frac{V_S}{n}, \] (4.20)

and the voltage seen at the two MOSFETs \(M_1\) and \(M_2\) is [1]
\[ v_{M1} = v_{M2} = V_S. \] (4.21)

The magnetizing current reaches zero at time \(t = D_r T + t_m\) terminating this interval.
4.3.3 Time Interval \((D_rT + t_m) < t \leq T\)

Figure 4.3(c) shows the ideal equivalent circuit for the two-switch forward converter for the time interval \((D_rT + t_m) < t \leq T\). The two MOSFET switches \(M_1\) and \(M_2\), diodes \(D_{c1}, D_{c2}, D_{r1}\) during this time interval are turned OFF and the free-wheeling diode \(D_{r2}\) is ON. The voltage across the transformer winding, \(L_{mag}\) and the rectifier diode \(D_{r2}\) are \(v_1 = v_2 = v_{L(mag)} = v_{Dr2} = 0\) [1]. The voltage across the two switches and the two clamping diodes \(D_{c1}\) and \(D_{c2}\) are [1]

\[
v_{M1} = v_{M2} = \frac{V_S}{2},
\]

and

\[
v_{Dc1} = v_{Dc2} = -\frac{V_S}{2},
\]

respectively. From equations (4.14) and (4.15), the voltage across the filter inductor \(L_f\) and the current through the rectifier diode \(D_{r2}\), inductor \(L_f\) are given by

\[
v_{Lf} = -V_O = L_f \frac{di_{Lf}}{dt}
\]

and

\[
i_{Dr2} = i_{Lf} = \frac{-V_O}{L_f} (t - (D_rT + t_m)) + i_{Lf}(D_rT + t_m).
\]

4.3.4 Simulation Data

Figure 4.4 and 4.5 shows the simulation results of the PWM two-switch forward converter using two IRF740 power MOSFETs as the two switches \(M_1\) and \(M_2\), two MR826 power diodes as the two clamping diodes \(D_{c1}, D_{c2}\) and two MBR2540 Schottky diodes as rectifier diodes \(D_{r1}\) and \(D_{r2}\) respectively. The input voltage of the converter is \(V_S = 110\) V, and outputs 20 V at the load, operating at maximum duty-ratio of \(D_{rmax} = 0.44\) \(\mu\)s. The filter inductor and filter capacitor values are tweaked for the simulation to \(L_f = 340\) \(\mu\)H and \(C_f = 20\) \(\mu\)F. The HF transformer used for the simulation has the following dimensions: Cross-sectional area \(A_c = 0.864\) \(\text{cm}^2\),
length of the core $l_c = 33.3$ cm and the volume of the core $V_c = 2.88$ cm$^3$ [21]. The magnetizing inductance of the transformer $L_{mag}$ is 4 mH.

### 4.4 Maximum Limit on Duty-ratio

Same like in conventional forward converter, the transformer core reset is vital for safe operation of two-switch forward converter. If the core is not completely reset, more and more energy will be accumulated in the transformer core in the consequent switching cycles and results in failure of the converter [1]. Therefore, there is a maximum permissible value of duty-ratio $D_{rMAX}$ one should avoid to ensure the safe operation of the converter. One can express the condition for transformer core reset as [1]

$$ (D_rT + t_m) \leq T \quad (4.26) $$

At $D_{rMAX}$, equation (4.26) will be

$$ D_{rMAX}T + t_m = T, \quad (4.27) $$

from which

$$ t_m = (1 - D_{rMAX})T. \quad (4.28) $$

From equations (4.8), (4.14), and (4.24), volt-second balance can be established as [1],

$$ V_S D_{rMAX}T = V_S(1 - D_{rMAX})T. \quad (4.29) $$

Re-arranging this equation yields, $D_{rMAX} = 0.5$. Therefore, one should not operate the two-switch forward converter with duty-ratio greater than 0.5 [1]

### 4.5 Voltage and Current Stresses of the Components

The peak voltage across the two MOSFETs $M_1$ and $M_2$ [1]

$$ V_{M1max} = V_{M2max} = V_{Smax}, \quad (4.30) $$
from which, the peak current through the two switches can be expressed as,

\[ I_{M1}^{\text{MAX}} = I_{M2}^{\text{MAX}} = \frac{I_{Dr1}^{\text{MAX}}}{n} + \Delta i_{L\text{mag}(\text{max})}. \quad (4.31) \]

The maximum voltage of the two rectifier diodes is given by [1]

\[ v_{Dr1}^{\text{MAX}} = v_{Dr2}^{\text{MAX}} = \frac{V_{S\text{max}}}{n}, \quad (4.32) \]

The peak currents through the two rectifier diodes on the secondary side of the transformer are [1]

\[ I_{Dr1}^{\text{MAX}} = I_{Dr2}^{\text{MAX}} = I_{O\text{max}} + \frac{\Delta i_{L\text{mag}(\text{max})}}{n}. \quad (4.33) \]

The maximum voltage at the two de-magnetizing diodes \( D_{c1} \) and \( D_{c2} \) is [1]

\[ v_{Dc1}^{\text{MAX}} = v_{Dc2}^{\text{MAX}} = -V_{S\text{MAX}}. \quad (4.34) \]

The peak current through the clamping diodes \( D_{c1} \) and \( D_{c2} \) is [1]

\[ I_{Dc1}^{\text{MAX}} = I_{Dc2}^{\text{MAX}} = \Delta i_{L\text{mag}(\text{max})} = \frac{D_{r\text{min}} V_{S\text{max}}}{f_s L_{\text{mag}}} = \frac{D_{r\text{min}} V_{S\text{max}}}{f_s L_{\text{mag}}}. \quad (4.35) \]

## 4.6 Determination of the DC Voltage Transfer Function \( M(D) \)

From Figures 4.2(b), 4.2(c) and from equations (4.8), (4.14), the voltage across the filter inductor \( L_f \) is \( v_{Lf} = \left( \frac{V_s}{n} - V_O \right) \) for \( 0 < t \leq D_r T \) and \( v_{Lf} = -V_O \) for \( D_r T < t \leq T \). Establishing volt-second balance, we get [1],

\[ \left( \frac{V_s}{n} - V_O \right) D_r T = V_O (1 - D_r) T. \quad (4.36) \]

Re-arranging equation (4.36) to get the dc voltage transfer function of loss less converter as

\[ M(D) \equiv \frac{V_O}{V_S} = \frac{I_S}{I_O} = \frac{D_r}{n}, \quad \text{for} \quad D_r \leq D_{r\text{MAX}}. \quad (4.37) \]

From equation (4.37) one can obtain the dc current transfer function as

\[ I(D) \equiv \frac{I_O}{I_S} = \frac{n}{D_r}, \quad \text{for} \quad D_r \leq D_{r\text{MAX}}. \quad (4.38) \]
4.7 Minimum Inductance at CCM/DCM Boundary

The inductor current waveform $i_{L_f}$ at CCM/DCM boundary is shown in the Figure 4.6 [1]. From this waveform, we can express the current through the filter inductor as [1]

$$i_{L_f} = -\frac{V_O}{L_f}(t - D_rT) + i_{L_f}(D_rT), \quad \text{for} \quad D_r T < t \leq T,$$

resulting in

$$i_{L_f}(T) = -\frac{V_OT(1 - D_r)}{L_f} + i_{L_f}(D_rT) = 0,$$

from which

$$\Delta i_{L_{f_{max}}} = \frac{V_O T(1 - D_{r_{min}})}{L_{f_{min}}} = \frac{V_O (1 - D_{r_{min}})}{f_s L_{f_{min}}}.$$  \hspace{1cm} (4.41)

At CCM/DCM boundary, the output current $I_O$ and output resistance $R_L$ can be expressed as [1]

$$I_O(B) = \frac{\Delta i_{L_{f_{max}}}}{2} = \frac{V_O (1 - D_{r_{min}}))}{2f_s L_{f_{min}}} = \frac{V_O}{R_{L_{max}}}$$ \hspace{1cm} (4.42)

and

$$R_L(B) = \frac{V_O}{I_O(B)} = \frac{2f_s L_f}{1 - D_r}, \quad \text{for} \quad D_r \leq D_{r_{max}}.$$ \hspace{1cm} (4.43)
Therefore, the minimum value of the inductance \(L_f\) can be found as [1]

\[
L_f(\text{min}) = \frac{V_O(1 - D_{r\text{min}})}{2f_s I_O(B)},
\]  

(4.44)

4.8 Theoretical Calculation of Power Losses and Estimation of Efficiency in a Two-switch PWM DC-DC Forward Converter Operating in CCM

The equivalent circuit of two-switch forward converter with parasitic resistances is shown in the Figure 4.7 [1]. [1] \(r_{DS}\) is the switch drain-to-source resistance in each MOSFET when it is ON. Since, the circuit has two similar fast recovery diodes in the primary side of the transformer, let \(R_{Fc}, V_{Fc}\) be the forward resistance and forward voltage of the two clamping diodes \(D_{c1}, D_{c2}\) respectively. Let \(R_{Fr}, V_{Fr}\) be the forward resistance and its forward voltage of the two rectifier diodes \(D_{r1}, D_{r2}\) respectively. The equivalent series resistance (ESR) of the filter capacitor and filter inductor are symbolized by \(r_{Cf}\) and \(r_{Lf}\) respectively. Let \(r_{w1}\) and \(r_{w2}\) be the winding resistances of the primary and secondary windings of the transformer respectively [1].

![Figure 4.7: Equivalent circuit to determine the losses of each component in two-switch forward converter.](image)

The current through the two MOSFET switches \(M_1\) and \(M_2\) can be approximated
by [1]
\[ i_{M1} = i_{M2} = \begin{cases} \frac{I_O}{n} & \text{for } 0 < t \leq D_r T; \\ 0 & \text{for } D_r T < t \leq T. \end{cases} \] (4.45)

From equation (4.45), the RMS current through the two MOSFETs are given as,
\[ I_{M1\text{rms}} = I_{M2\text{rms}} = \sqrt{\frac{1}{T} \int_0^T i_{M1}^2 \, dt} = \sqrt{\frac{1}{T} \int_0^{D_r T} \left( \frac{I_O}{n} \right)^2 \, dt} = \frac{I_O \sqrt{D_r}}{n}, \] (4.46)

and the conduction losses in each of the MOSFETs,
\[ P_{rDS1} = r_{DS} I^2_{M1\text{rms}} = \frac{r_{DS} D_r I^2_O}{n^2} = \frac{D_r r_{DS} I_O}{n^2 R_L} P_O. \] (4.47)

The switching loss in each MOSFET is given as [1]
\[ P_{\text{sw1}} = f_s C_O V_S^2 = \frac{f_s C_O V^2_O}{M(D)^2} = \frac{f_s C_O R_L}{M(D)^2} = \frac{n^2 f_s C_O R_L}{P_O}. \] (4.48)

From equations (4.47) and (4.48) the total power dissipation in each of the MOSFETs is given as [1],
\[ P_{FET} = P_{rDS} + \frac{P_{\text{sw}}}{2} = \frac{D_r r_{DS} I^2_O}{n^2} + \frac{1}{2} \frac{f_s C_O V^2_S}{M(D)^2} = \left( \frac{D_r r_{DS}}{n^2 R_L} + \frac{f_s C_O R_L}{2M(D)^2} \right) P_O. \] (4.49)

The transformer primary winding conduction loss is given by [1]
\[ P_{rw1} = r_{w1} I^2_{M1\text{rms}} = \frac{r_{w1} D_r I^2_O}{n^2} = \frac{D_r r_{w1} I_O}{n^2 R_L} P_O. \] (4.50)

The two diodes located on the transformer secondary, \( D_{r1} \) and \( D_{r2} \) are identical. Therefore, the rectifier diode \( D_{r1} \) has current that can be approximated as [1]
\[ i_{Dr1} = \begin{cases} I_O & \text{for } 0 < t \leq D_r T; \\ 0 & \text{for } D_r T < t \leq T. \end{cases} \] (4.51)

resulting in its rms value
\[ I_{Dr1\text{rms}} = \sqrt{\frac{1}{T} \int_0^T i_{Dr1}^2 \, dt} = \sqrt{\frac{1}{T} \int_0^{D_r T} I^2_O \, dt} = I_O \sqrt{\frac{1}{T}(D_r T)} = I_O \sqrt{D_r}. \] (4.52)

The power loss due to \( R_{Fr} \) [1]
\[ P_{RFr} = R_{Fr} I^2_{Dr1\text{rms}} = D_r R_{Fr} I^2_O = \frac{D_r R_{Fr}}{R_L} P_O. \] (4.53)
The average current through the rectifier diode $D_{r1}$ is [1]

$$I_{Dr1} = \frac{1}{T} \int_0^T i_{Dr1} \, dt = \frac{1}{T} \int_0^{D_rT} I_O \, dt = D_r I_O, \quad (4.54)$$

which results in the power loss due to $V_{Fr}$ of the diode $D_{r1}$ [1],

$$P_{VFr1} = V_{Fr} I_{Dr1} = V_{Fr} I_O D_r = \frac{D_r V_{Fr}}{V_O} P_O. \quad (4.55)$$

From equations (4.53) and (4.55), the total conduction loss in the rectifier diode $D_{r1}$ is given as

$$P_{Dr1} = P_{VFr} + P_{RFr} = V_{Fr} I_O D_r + R_{Fr} I_O^2 D_r = P_O D_r \left( \frac{V_{Fr}}{V_O} + \frac{R_{Fr}}{R_L} \right). \quad (4.56)$$

The power loss in transformer winding $r_{w2}$ is [1]

$$P_{rw2} = r_{w2} I_{Dr1 rms}^2 = D_r r_{w2} I_O^2 = \frac{D_r r_{w2}}{R_L} P_O. \quad (4.57)$$

From equations (4.50) and (4.57), the power loss in both the primary and secondary windings is

$$P_{rw} = P_{rw1} + P_{rw2} = \left( \frac{r_{w1}}{n^2} + r_{w2} \right) \frac{D_r}{R_L} P_O. \quad (4.58)$$

The rectifier diode current $i_{Dr2}$ can be approximated by [1]

$$i_{Dr2} = \begin{cases} 0, & \text{for } 0 < t \leq D_r T, \\ I_O, & \text{for } D_r T < t \leq T. \end{cases} \quad (4.59)$$

from which we obtain its rms value as,

$$I_{Dr2 rms} = \sqrt{\frac{1}{T} \int_0^T i_{Dr2}^2 \, dt} = \sqrt{\frac{1}{T} \int_{D_rT}^T I_O^2 \, dt} = I_O \sqrt{1 - D_r}. \quad (4.60)$$

and the power losses in $R_{Fr}$ [1]

$$P_{RFr} = R_{Fr} I_{Dr2 rms}^2 = (1 - D_r) R_{Fr} I_O^2 = \frac{(1 - D_r) R_{Fr}}{R_L} P_O. \quad (4.61)$$

From equation (4.59), the average current through $D_{r2}$ can be found as [1]

$$I_{Dr2} = \frac{1}{T} \int_0^T i_{Dr2} \, dt = \frac{1}{T} \int_{D_rT}^T I_O \, dt = (1 - D_r) I_O, \quad (4.62)$$
which results in power loss due to $V_{Fr}$ of the diode $D_{r2}$ is obtained as \[ P_{V_{Fr}} = V_{Fr}I_{D_{r2}} = V_{Fr}I_O(1 - D_r) = \frac{(1 - D_r)V_{Fr}}{V_O}P_O. \] (4.63)

From equations (4.61) and (4.63), the overall conduction loss in the diode $D_{r2}$ is given by

\[
P_{D_{r2}} = P_{V_{Fr}} + P_{RFr} = V_{Fr}I_O(1 - D_r) + R_{Fr}I_O^2(1 - D_r) = P_O(1 - D_r) \left( \frac{V_{Fr}}{V_O} + \frac{R_{Fr}}{R_L} \right). \]

The currents through the clamping diodes $D_{c1}$ and $D_{c2}$, which is usually $10\%$ of the maximum peak current of the ideal transformer primary can be approximated as \[ i_{Dc1} = i_{Dc2} = \begin{cases} 
0, & \text{for } 0 < t \leq D_r T, \\
0.1\frac{I_O}{n}, & \text{for } D_r T < t \leq (D_r T + t_m), \\
0, & \text{for } (D_r T + t_m) < t \leq T.
\end{cases} \] (4.65)

Since $10\%$ of peak of the transformer primary is equal $10\%$ of $I_O$, the rms current through the two clamping diodes can be expressed as \[ I_{Dc1rms} = I_{Dc2rms} = \left( \frac{0.1I_O}{n} \right) \sqrt{D_r}. \] (4.66)

The power loss in $R_{Fc}$ is

\[
P_{RFc} = R_{Fc}I_{Dc1rms}^2 = 10 \times 10^{-3} \times R_{Fc}D_r \left( \frac{I_O}{n} \right)^2 = 10 \times 10^{-3}P_O \times D_r \left( \frac{R_{Fc}}{n^2R_L} \right). \]

The average value of the current through the two clamping diodes $D_{c1}$ and $D_{c2}$ is found to be

\[
I_{Dc1} = I_{Dc2} = \frac{1}{T} \int_0^T i_{D1} dt = \frac{1}{T} \int_{D_r T}^{D_r T + t_m} \left( \frac{0.1 I_O}{n} \right) \, dt = \frac{0.1I_O D_r}{n}.
\] (4.68)

from which the power loss associated with the voltage $V_{F1}$ of the diodes $D_{c1}$ and $D_{c2}$ is obtained as \[ P_{V_{Fc}} = V_{Fc}I_{Dc1} = V_{Fc}I_{Dc2} = 0.1 \times V_{Fc} \left( \frac{I_O D_r}{n} \right) = 0.1 \times D_r V_{Fc} \left( \frac{P_O}{nV_O} \right). \] (4.69)
Thus, the overall conduction loss of the clamping diodes $D_{c1}$ and $D_{c2}$ will be

$$P_{Dc1} = P_{Dc2} = P_{VFc} + P_{RFc} = \left[ 0.1 \left( \frac{P_O}{nV_O} \right) V_{Fc}D_r \right] + \left[ 0.01 \times P_O D_r \left( \frac{R_{Fc}}{n^2 R_L} \right) \right]$$

$$P_{Dc1} = P_{Dc2} = \frac{P_O D_r}{n} \left[ 0.1 \left( \frac{V_{Fc}}{V_O} \right) + \left( \frac{0.01}{n} \right) \left( \frac{R_{Fc}}{R_L} \right) \right].$$

The current through the inductor $L_f$ can be approximated as [1]

$$i_{Lf} \approx I_O, \quad (4.70)$$

which gives its rms value,

$$I_{Lf rms} = I_O. \quad (4.71)$$

The conduction loss due to $r_{Lf}$ in the filter inductor $L_f$ is given by [1]

$$P_{rLf} = r_{Lf} I_{Lf rms}^2 = r_{Lf} I_O^2 = r_{Lf} \left( \frac{P_O}{R_L} \right). \quad (4.72)$$

The filter capacitor current $i_{Cf}$ can be written as [1]

$$i_{Cf} = \left\{ \begin{array}{ll}
\frac{\Delta i_{Lf} t}{D_r T} - \frac{\Delta i_{Lf}}{2}, & \text{for } 0 < t \leq D_r T; \\
-\frac{\Delta i_{Lf} (t-D_r T)}{(1-D_r)T} + \frac{\Delta i_{Lf}}{2}, & \text{for } D_r T < t \leq T.
\end{array} \right. \quad (4.73)$$

The filter capacitor RMS current can be found as [1]

$$I_{Cf rms} = \sqrt{\frac{1}{T} \int_0^T i_{Cf}^2 dt} = \frac{\Delta i_{Lf}}{\sqrt{12}} = \frac{V_O (1 - D_r)}{\sqrt{12 f_s L_f}}, \quad (4.74)$$

and the power loss due to $r_{Cf}$ of the filter capacitor [1]

$$P_{rcf} = r_{Cf} I_{Cf rms}^2 = \frac{r_{Cf} (\Delta i_{Lf})^2}{12} = \frac{r_{Cf} R_L (1 - D_r)^2}{12 f_s^2 L_f^2} P_O. \quad (4.75)$$

The overall power loss is given by [1]

$$P_{LS} = 2(P_{rDS1} + P_{sw1}) + (P_{Dc1} + P_{Dc2}) + (P_{rw1} + P_{rw2}) + (P_{Dr1} + P_{Dr2}) + (P_{rLf} + P_{rcf}). \quad (4.76)$$

$$P_{LS} = 2 \left( \frac{D_r r_{DS1}}{n^2} I_O^2 + \frac{1}{2} f_s C_O V_S^2 \right) + 2 \times \left( 0.01 \times \frac{R_{Fc} I_O^2 D_r}{n^2} + 0.1 D_r \frac{V_{Fc} I_O}{n} \right) \quad (4.77)$$
\[
\begin{aligned}
&+ (V_F R_I O + R_F r_I O^2) D_r + \left( D_r r_w I_O^2 + \frac{r_w D_r I_O^2}{n^2} \right) + r_{LI} I_O^2 + \frac{r_{CF}(\Delta i_{LI}^2)}{12}.
\end{aligned}
\]

\[
P_{LS} = \left[ \left( \frac{2D_r r_{DS}}{n^2 R_L} + \frac{f_s C_O R_L}{M(D)^2} \right) + D_r \left( \frac{0.1V_F}{n V_O} + \frac{0.01 R_F c}{n^2 R_L} \right) + \left( \frac{V_F}{V_O} + \frac{R_F r}{R_L} \right) \right.
\]

\[
\left( \frac{D_r r_w 1}{n^2 R_L} + \frac{D_r r_w 2}{R_L} \right) + \left( \frac{r_{LI} R_L}{R_L} + \frac{r_{CF} R_L (1 - D_r)^2}{12 f_s^2 L_f^2} \right) \right] P_O.
\]

(4.78)

Thus, the overall efficiency of the converter is given by [1]

\[
\eta_c = \frac{P_O}{P_O + P_{LS}} = \frac{1}{1 + \frac{P_{LS}}{P_O}}
\]

\[
= \frac{1}{1 + \frac{V_F}{V_O} + \frac{(2r_{DS} + r_w 1 + 0.01 R_F c) D_r}{n^2 R_L} + \frac{R_F r + D_r r_w 2}{R_L} + \frac{0.1 D_r V_F}{n V_O} + \frac{f_s C_O R_L}{M(D)^2} + \frac{r_{CF} R_L (1 - D_r)^2}{12 f_s^2 L_f^2}}.
\]

(4.79)

### 4.9 DC Voltage Transfer Function for a Lossy Converter

The average value of the source current of the converter is [1]

\[
I_S = \frac{1}{T} \int_0^T i_{M1} \, dt = \frac{1}{T} \int_0^T \frac{I_O}{n} \, dt = \frac{D_r I_O}{n},
\]

(4.80)

from which, one can get the dc current transfer function as,

\[
I(D) \equiv \frac{I_O}{I_t} = \frac{n}{D_r}.
\]

(4.81)

The converter efficiency [1]

\[
\eta_c = M(D) I(D) = \frac{n M(D)}{D_r}.
\]

(4.82)

From equation (4.82), the dc voltage transfer function of the lossy two-switch forward converter can be found as

\[
M(D) = \frac{\eta_c}{I(D)} = \frac{\eta_c D_r}{n}
\]

(4.83)

which gives
\[
D_r = n \left[ 1 + \frac{V_{Fr}}{V_O} + \frac{(2r_{DS}+r_{w1}+0.01R_{Fe})D_r}{n^2R_L} + \frac{R_{Fr}+r_{Lf}+D_rr_{w2}}{R_L} + \frac{0.1D_r V_{Fe}}{nV_O} + \frac{f_s C_O R_L}{M(D)^2} + \frac{r_{Cf} R_L (1-D_r)^2}{12 f_s^2 L_f^2} \right].
\]

From equation (4.83), the duty-ratio is
\[
D_r = \frac{nM(D)}{\eta_c}. \quad (4.84)
\]

Substitute equation (4.84) into (4.79), the efficiency of the two-switch forward converter can be obtained as,
\[
\eta_c = \frac{N_{\eta_c}}{D_{\eta}} \quad (4.85)
\]

where
\[
N_{\eta_c} = 1 - \left( \frac{2r_{DS}+r_{w1}+0.01R_{Fe}}{n^2R_L} + \frac{r_{w2}}{R_L} + \frac{0.1V_{Fe}}{nV_O} - \frac{n r_{Cf} R_L}{6n f_s^2 L_f^2} \right) nM(D)
\]
\[
+ \left\{ \left( \frac{2r_{DS}+r_{w1}+0.01R_{Fe}}{n^2R_L} + \frac{r_{w2}}{R_L} + \frac{0.1V_{Fe}}{nV_O} - \frac{n r_{Cf} R_L}{6n f_s^2 L_f^2} \right) nM(D) - 1 \right\}^2
\]
\[
- \left[ 1 + \frac{V_{Fr}}{V_O} + \frac{R_{Fr}+r_{Lf}}{R_L} + \frac{f_s C_O R_L}{M(D)^2} + \frac{r_{Cf} R_L}{12 f_s^2 L_f^2} \right] \left[ \frac{n^2 r_{Cf} R_L M(D)^2}{3 f_s^2 L_f^2} \right]^{1/2} \quad (4.86)
\]

and
\[
D_{\eta} = 2 \left[ 1 + \frac{V_{Fr}}{V_O} + \frac{R_{Fr}+r_{Lf}}{R_L} + \frac{f_s C_O R_L}{M(D)^2} + \frac{r_{Cf} R_L}{12 f_s^2 L_f^2} \right]. \quad (4.87)
\]
Design of a PWM Two-Switch DC-DC Forward Converter in Continuous Conduction Mode

A PWM two-switch forward converter operating in CCM is designed to meet the following specifications: Output voltage $V_O = 20$ V, maximum output power $P_{O_{\text{max}}}$ = 60 W, ripple factor $\frac{V}{V_O} \leq 1\%$, switching frequency $f_s = 100$ kHz. The input voltage is universal ac utility line rectified voltage.

The maximum and minimum values of the input source voltage $V_S$ are [1]

$$V_{S\text{max}} = 375 \text{ V},$$

and

$$V_{S\text{min}} = 110 \text{ V},$$

The dc voltage transfer function can have the maximum and minimum values given as [1]

$$[M(D)]_{\text{max}} = \frac{V_O}{V_{S\text{min}}} = \frac{20}{110} = 0.1818,$$

and

$$[M(D)]_{\text{min}} = \frac{V_O}{V_{S\text{max}}} = \frac{20}{375} = 0.0533.$$ (5.4)

The output current can have the maximum and minimum values given as [1]

$$I_{O\text{max}} = \frac{P_{O\text{max}}}{V_O} = \frac{60}{20} = 3.0 \text{ A},$$

and

$$I_{O\text{min}} = \frac{P_{O\text{min}}}{V_O} = \frac{6}{20} = 0.3 \text{ A}.$$ (5.6)

From equations (5.5) and (5.6), the load resistance can have maximum and minimum values

$$R_{L\text{max}} = \frac{V_O}{I_{O\text{min}}} = \frac{20}{0.3} = 66.67 \Omega,$$

and

$$R_{L\text{min}} = \frac{V_O}{I_{O\text{max}}} = \frac{20}{3} = 6.67 \Omega.$$ (5.8)
Assuming the converter efficiency $\eta_c = 85\%$, and maximum duty-ratio $D_{rMAX} = 0.47$, one can calculate the turns ratio of the transformer as [1]

$$n = \frac{\eta_c D_{rmax}}{[M(D)]_{max}} = \frac{0.85 \times 0.47}{0.1818} = 2.19.$$  \hspace{1cm} (5.9)

Pick $n=2.18$. From equation (5.9) the duty-ratio $D_r$ can have the maximum and minimum values given by [1]

$$D_{rmax} = \frac{n[M(D)]_{max}}{\eta_c} = \frac{2.18 \times 0.1818}{0.85} = 0.466,$$  \hspace{1cm} (5.10)

and

$$D_{rmin} = \frac{n[M(D)]_{min}}{\eta_c} = \frac{2.18 \times 0.0533}{0.85} = 0.136.$$  \hspace{1cm} (5.11)

The calculated value of $D_{rmax}$ is less than the assumed $D_{rMAX}$. For the converter to be in CCM, the minimum value of filter inductance $L_{f(min)}$ required for the converter is [1]

$$L_{f(min)} = \frac{R_{Lmax}(1 - D_{rmin})}{2f_s} = \frac{66.67 \times (1 - 0.136)}{2 \times 100 \times 10^3} = 288 \mu H.$$  \hspace{1cm} (5.12)

Pick $L_f = 345 \mu H$. In the filter inductor $L_f$, the peak-to-peak current ripple is calculated as [1]

$$\Delta i_{L_f(max)} = \frac{V_O(1 - D_{rmin})}{f_s L_f} = \frac{20 \times (1 - 0.136)}{100 \times 10^3 \times 345 \times 10^{-6}} = 0.5005.$$  \hspace{1cm} (5.13)

The ripple voltage $V_r$, which is 1% of the output voltage (from the specification) [1],

$$V_r = 0.01V_O = 0.01 \times 20 = 0.2 \text{ V}.$$  \hspace{1cm} (5.14)

From equations (5.13) and (5.14), the filter capacitor ESR can be calculated as [1]

$$r_{Cf(max)} = \frac{V_r}{\Delta i_{L_f(max)}} = \frac{0.2}{0.5008} = 0.4 \Omega.$$  \hspace{1cm} (5.15)

Choose $r_{Cf} = 0.35 \Omega$.

Therefore, the filter capacitance $C_f$ can be chosen as [1]

$$C_{f(min)} = \max \left\{ \frac{D_{rmax}}{2f_s r_{Cf}}, \frac{1 - D_{rmin}}{2f_s r_{Cf}} \right\} = \frac{1 - D_{rmin}}{2f_s r_{Cf}} = \frac{1 - 0.136}{2 \times 10^5 \times 0.4} = 12.3 \mu F.$$  \hspace{1cm} (5.16)
Choose filter capacitor $C_f = 20 \, \mu F/0.35 \, \Omega$. The two rectifier diodes has current and voltage stresses given as [1]

$$V_{Dr1M(max)} = V_{Dr2M(max)} = \frac{V_{S(max)}}{n} = \frac{375}{2.18} = 172.02 \, \text{V}, \quad (5.17)$$

and

$$I_{Dr1M(max)} = I_{Dr2M(max)} = I_{O(max)} + \frac{\Delta i_{Lf(max)}}{2} = 3 + \frac{0.5008}{2} = 3.2504 \, \text{A}. \quad (5.18)$$

respectively.

The peak current through the transformer primary winding [1]

$$I_{p(max)} = \frac{I_{Dr1M(max)}}{n} = \frac{3.25}{2.18} = 1.49 \, \text{A}. \quad (5.19)$$

The peak current through the magnetizing inductance is assumed to be 10 % of the primary winding peak current [1]

$$\Delta i_{Lmag(max)} = 0.1 \times I_{p(max)} = 0.1 \times 1.49 = 0.149 \, \text{A}, \quad (5.20)$$

thereby, calculating the minimum value of the magnetizing inductance [1]

$$L_{mag(min)} = \frac{D_{rmin} V_{S(max)}}{f_s \Delta i_{Lmag(max)}} = \frac{0.136 \times 375}{100 \times 10^3 \times 0.149} = 3.4 \, \text{mH}. \quad (5.21)$$

Pick $L_{mag} = 4 \, \text{mH}$. The two demagnetization diodes $D_{c1}$ and $D_{c2}$ has maximum current and voltage stresses given as [1]

$$V_{Dc1M(max)} = V_{Dc2M(max)} = V_{S(max)} = 375 \, \text{V}, \quad (5.22)$$

$$I_{Dc1M(max)} = I_{Dc2M(max)} = \Delta i_{Lmag(max)} = 0.149 \, \text{A}. \quad (5.23)$$

The two MOSFET switches $M_1$ and $M_2$ has a maximum voltage stress of [1]

$$V_{M1M(max)} = V_{M2M(max)} = V_{S(max)} = 375 \, \text{V}, \quad (5.24)$$

and maximum current stress of [1]

$$I_{M1M(max)} = I_{M2M(max)} = \frac{I_{O(max)}}{n} + \Delta i_{Lmag(max)} = \frac{3}{2.18} + 0.149 = 1.5258 \, \text{A}. \quad (5.25)$$
The two MOSFET switches $M_1$ and $M_2$ are chosen to be two n-channel IRF740 power MOSFETs which has $V_{DSS} = 400$ V, $r_{DS(on)} = 0.55$ Ω, $I_D = 10$ A, output capacitance $C_O = 100$ pF and the gate charge $Q_{g(max)} = 63$ nC [1], [24]. For the two clamping diodes, $D_{c1}$ and $D_{c2}$, two fast recovery silicon MR826 diode rectifiers are chosen which has a reverse voltage of $V_R = 600$ V, $I_F = 5$ A, forward voltage drop of $V_{F1} = 1.0$ V and forward resistance of $R_{F1} = 0.2$ Ω [1], [26]. Two MBR20200 high voltage power schottky barrier diodes are chosen for the two rectifier diodes $D_{r1}$ and $D_{r2}$, which has a maximum dc blocking voltage of $V_R = 200$ V and maximum current of $I_F = 20$ A with a forward voltage drop of $V_{F2} = 0.95$ V and forward resistance of $R_{F2} = 95$ mΩ at a junction temperature $T_J = 20^\circ$C [25]. The power losses across the components are calculated and efficiency is estimated for $V_{Smin} = 110$ V, maximum output power $P_{O(max)} = 60$ W and with minimum load resistance $R_{L(min)} = 6.67$ Ω. The ohmic loss when each MOSFETs is ON, is given by [1]

$$P_{rDS} = \frac{D_{rmax} r_{DS} I_{Omax}^2}{n^2} = \frac{0.466 \times 0.55 \times 3^2}{2.18^2} = 0.4856 \text{ W, \hspace{1cm} (5.26)}$$

and the switching loss is across the MOSFET with $C_O$ as its parasitic output capacitance [1],

$$P_{sw} = f_s C_O V_{Smin}^2 = 10^5 \times 100 \times 10^{-12} \times 110^2 = 0.1211 \text{ W.} \hspace{1cm} (5.27)$$

Assume the winding resistance of the transformer primary, $r_{w1} = 75$ mΩ, the power loss is [1],

$$P_{rw1} = \frac{r_{w1} D_{rmax} I_{Omax}^2}{n^2} = \frac{0.075 \times 0.466 \times 3^2}{2.18^2} = 0.06615 \text{ W. \hspace{1cm} (5.28)}$$

The power loss in the rectifier diode $D_{r1}$ due to $R_{F2}$ [1]

$$P_{RFr1} = D_{rmax} R_{F2} I_{Omax}^2 = 0.466 \times 0.095 \times 3^2 = 0.3984 \text{ W, \hspace{1cm} (5.29)}$$

and due to $V_{F2}$, the power loss in the diode $D_{r1}$ is given as [1]

$$P_{VFr1} = V_{F2} I_{Omax} D_{max} = 0.95 \times 3 \times 0.466 = 1.3281 \text{ W.} \hspace{1cm} (5.30)$$
From equations (5.29) and (5.30), the total conduction losses in diode $D_{r1}$ is

$$P_{Dr1} = P_{RFr1} + P_{VFr1} = 0.3984 + 1.3281 = 1.7265 \text{ W.}$$  \hspace{1cm} (5.31)

Let us assume that the winding resistance of the transformer secondary $r_{w2} = 20 \text{ m}\Omega$, the power loss in $r_{w2}$ is \[1\]

$$P_{rw2} = D_{max} r_{w2} I_{Omax}^2 = 0.466 \times 0.02 \times 3^2 = 0.084 \text{ W.}$$ \hspace{1cm} (5.32)

Hence, from the equations (5.28) and (5.32), power loss in both primary and secondary windings is

$$P_{rw} = P_{rw1} + P_{rw2} = 0.06615 + 0.084 = 0.1501 \text{ W.}$$ \hspace{1cm} (5.33)

The power loss in the rectifier diode $D_{r2}$ due to $R_{F2}$ is \[1\]

$$P_{RFr2} = (1 - D_{rmax}) R_{F2} I_{Omax}^2 = (1 - 0.466) \times 0.095 \times 3^2 = 0.456 \text{ W,}$$ \hspace{1cm} (5.34)

and due to $V_{F2}$, the conduction loss in the diode $D_{r2}$ is \[1\]

$$P_{VFr2} = (1 - D_{rmax}) V_{F2} I_{Omax} = (1 - 0.466) \times 0.95 \times 3 = 1.52 \text{ W.}$$ \hspace{1cm} (5.35)

Hence, from equations (5.34) and (5.35), the overall power loss in the rectifier diode $D_{r2}$ is

$$P_{Dr2} = P_{RFr2} + P_{VFr2} = 0.456 + 1.52 = 1.978 \text{ W.}$$ \hspace{1cm} (5.36)

In the two clamping diodes $D_{c1}$ and $D_{c2}$, the forward resistance $R_{F1}$ causes the power loss given as \[1\]

$$P_{RFc1} = P_{RFc2} = 10 \times 10^{-3} \times R_{F1} \times \left( \frac{I_O}{n} \right)^2 = 0.01 \times 0.2 \times \left( \frac{3}{2.18} \right)^2 = 3.79 \text{ mW.}$$ \hspace{1cm} (5.37)

The loss in the clamping diodes $D_{c1}$ and $D_{c2}$ due to the forward voltage is

$$P_{VFc1} = P_{VFc2} = 0.1 \times \frac{I_O}{n} \times V_{F1} D_{rmax} = 0.1 \times 1.376 \times 1.0 \times 0.466 = 0.064 \text{ W.}$$ \hspace{1cm} (5.38)
From equations (5.37) and (5.38), the total power loss in the diodes \(D_{c1}\) and \(D_{c2}\) is

\[
P_{Dc1} = P_{Dc2} = P_{RFC1} + P_{VFC1} = 3.79 \times 10^{-3} + 64 \times 10^{-3} = 67.8 \text{ mW. (5.39)}
\]

Depending on the value of the filter inductor, the inductor ESR is assumed as \(r_{Lf} = 100 \text{ m}\Omega\), the power loss is calculated as [1]

\[
P_{rLf} = r_{Lf} I_{Omax}^2 = 0.1 \times 3^2 = 0.9 \text{ W. (5.40)}
\]

The power loss due to \(r_{Cf}\) of the filter capacitor \(C_f\) is [1]

\[
P_{rCf} = \frac{r_{Cf} (\Delta i_{Lf(max)})^2}{12} = 0.005 \times 0.5004^2 = 0.1 \text{ mW. (5.41)}
\]

The overall power loss in all the components is given as [1]

\[
P_{Losses} = 2 \times (P_{rDS} + P_{su}) + (P_{w1} + P_{w2}) + (P_{Dc1} + P_{Dc2}) + (P_{Dr1} + P_{Dr2}) + (P_{rLf} + P_{rCf})
\]

\[
= 2 \times (0.4856 + 0.1211) + (0.06615 + 0.084) + (2 \times 0.0678) + (1.7265 + 1.978)
\]

\[
+ (0.9 + 0.1 \times 10^{-3}) = 6.6987 \text{ W, (5.42)}
\]

from which, one can calculate theoretical efficiency [1]

\[
\eta_t = \frac{P_{Omax}}{P_{Omax} + P_{Losses}} = \frac{60}{60 + 6.698} = 89.95 \approx 90\%.
\]

The voltage pulse applied at the gates of the two switches is \(V_{GSm} = 15 \text{ V}\), therefore power loss in gate drive is estimated as [1]

\[
P_{GD} = f_s Q_{gmax} V_{GSm} = 10^5 \times (63 \times 10^{-9}) \times 10 = 0.063.
\]

From equations (4.83) - (4.87), the converter efficiency and duty-ratio can be calculated. Figures (5.1) - (5.6) show the variation of duty-ratio, efficiency vs load resistance and output current. From the plots, it is observed that the efficiency at the minimum input voltage is higher, than at the maximum input voltage. Also, it can be observed that, the duty-ratio \(D_r\) is dependant more on the input voltage source \(V_s\) than on the load resistance \(R_L\). If the load resistance changes from minimum to its maximum value, the duty-ratio has almost a constant value.
Figure 5.1: Plot of efficiency $\eta_c$ vs source voltage $V_S$ for various value of load resistance $R_L$.

Figure 5.2: Plot of duty-ratio $D_r$ vs input voltage $V_S$ for various values of load resistance $R_L$. 
Figure 5.3: Plot of efficiency $\eta_c$ vs output current $I_O$ for various values of input voltage $V_S$.

Figure 5.4: Plot of duty-ratio $D_r$ vs output current $I_O$ for various values of input voltage $V_S$. 
Figure 5.5: Plot of efficiency $\eta_c$ vs load resistance $R_L$ for various values of input voltage $V_S$.

Figure 5.6: Plot of duty-ratio $D_r$ vs load resistance $R_L$ for various values of input voltage $V_S$. 
6 Design of a HF Transformer for Two-Switch DC-DC Forward Converter Operating in CCM

6.1 Introduction

The transformer is an important magnetic component in any isolated SMPS DC-DC converter [17]. Most importantly, the successful operation of transformer in two-switch forward converter is crucial because, the unwanted energy stored in the transformer magnetizing inductance, non-magnetic regions, should be discharged completely before the commencement of the next switching cycle [1], [23]. Therefore, proper design of transformer is necessary in a two-switch forward converter. This chapter presents the important factors to be considered while designing the transformer in a DC-DC converter [17], [23]. A step-by-step design procedure of a two-winding HF transformer using Area-Product method in CCM mode of operation is presented. A design example is illustrated [16] - [23]. The three important factors to be considered while designing a transformer [17], [23] are 1) Core Material, which conducts the magnetic flux, 2) Core, which accommodates windings and provides a path for the establishment of flux and 3) Conductors, which carry the current in the transformer.

6.2 Core Material

Magnetic core material is one of the important factors to be considered because, it is responsible for conducting the magnetic flux $\phi$ in the transformer core. There are two types of magnetic materials [17]. Soft and Hard magnetic materials. It is not easy to magnetize and demagnetize a hard magnetic material. Soft magnetic materials, on the other hand are easy to magnetize and demagnetize [17]. Therefore they are used as core materials in the design of transformers, inductors which operate at HF. The important parameters to be considered while choosing a material are [17], [18]:

- [17], [18]:
1) relative permeability $\mu_{rc}$ 2) resistivity $\rho_c$ and, 3) saturation flux density $B_s$.

### 6.3 Core

The main purpose of a core in any magnetic component is to reduce the reluctance and to provide a well-defined path for magnetic flux [17]. Cores are made in wide variety of shapes and sizes such as Pot cores, ETD cores, PQ cores, RM cores etc. Cores made with ferro-magnetic material are chosen because of their wide frequency range, low cost and low eddy current losses [17], [18], [23].

### 6.4 Conductor

Conductors come in various shapes. Some of them are square wire, round wire, rectangular wire and litz wire [17], [23]. The conductor is chosen based upon the skin depth of the material [17], [20].

### 6.5 Area-Product Method

Area-product method gives a fair estimate of selecting a core based on the amount of energy it has to handle [17]. The peak value of the magnetic flux linkage from basic magnetic theory can be written as [7], [17], [18]

$$\lambda_{\text{peak}} = N_p A_c B_{\text{peak}} = L_p I_{\text{pmax}}.$$  \hspace{1cm} (6.1)

Where $A_c$, $B_{\text{peak}}$ and $L_p$ represents the cross-sectional area of the core, peak magnetic flux density and primary winding inductance of the transformer respectively. The maximum value of current density through a winding is given as [17], [18]

$$J_{\text{max}} = \frac{I_{\text{pmax}}}{A_{\text{wp}}},$$  \hspace{1cm} (6.2)

where $A_{\text{wp}}$ is the cross-sectional area of the primary winding wire. The window area of the transformer which accommodates the primary and the secondary windings is
given by [17]

\[ W_a = \frac{N_p A_{wp} + N_s A_{ws}}{k_u}. \]  

Assume that the two windings occupy the same area such that \( N_s A_{wp} = N_s A_{ws} \) and substituting equation (6.2) in equation (6.3), the window area

\[ W_a = \frac{2N_p A_{wp}}{k_u} = \frac{2N_p I_{p_{max}}}{k_u J_{max}}. \]  

(6.4)

The primary winding is taken into consideration here. The core-area product [17]

\[ A_p = W_a A_c = \left( \frac{2 \times N_p I_{p_{max}}}{k_u J_{max}} \right) \left( \frac{L_p I_{p_{max}}}{N_p B_{peak}} \right). \]  

(6.5)

Substituting equation (6.1) in equation (6.5), one can obtain the core-area product as

\[ A_p = \frac{2L_p I_{p_{max}}^2}{k_u J_{max} B_{peak}} = \frac{4W_{max}}{k_u J_{max} B_{peak}} \]  

(6.6)

where \( W_{max} \) is the maximum energy the core has to handle.

### 6.6 Design of a Two-Switch Forward Transformer for CCM

The specifications to design a transformer for the two-switch forward converter are:

- The input source voltage is ac universal rectified line voltage \( 110 < V_s \leq 375 \text{ V} \),
- output voltage is \( V_O = 20 \text{ V} \), output current ranges from \( 0.3 < I_O \leq 3 \text{ A} \) operating at a switching frequency of \( f_s = 100 \text{ kHz} \).

The steps to design a two-winding forward transformer for CCM operation are as follows. The output power can have the maximum and minimum values given by [17]

\[ P_{O_{max}} = V_O I_{O_{max}} = 20 \times 3 = 60 \text{ W} \]  

(6.7)

and

\[ P_{O_{min}} = V_O I_{O_{min}} = 20 \times 0.3 = 6 \text{ W}. \]  

(6.8)

The minimum load resistance is [17]

\[ R_{L_{min}} = \frac{V_O}{I_{O_{max}}} = \frac{20}{3} = 6.66 \text{ \Omega}. \]  

(6.9)
The maximum load resistance is [17]

\[ R_{Lmax} = \frac{V_O}{I_{Omin}} = \frac{20}{0.3} = 66.66 \, \Omega. \] (6.10)

The minimum dc voltage transfer function is given by [17]

\[ [M(D)]_{min} = \frac{V_O}{V_{Smax}} = \frac{20}{375} = 0.0533. \] (6.11)

The maximum dc voltage transfer function is given by [17]

\[ [M(D)]_{max} = \frac{V_O}{V_{Smin}} = \frac{20}{110} = 0.1818. \] (6.12)

Assuming efficiency of the converter \( \eta_c = 85\% \) and the maximum duty-ratio \( D_{rmax} = 0.47 \), from equation (6.12), one can calculate the turns ratio of the transformer given as

\[ n = \frac{\eta_c D_{rmax}}{[M(D)]_{max}} = \frac{0.85 \times 0.47}{0.1818} = 2.197. \] (6.13)

Pick transformer turns ratio \( n = 2.18 \). The minimum and maximum values of the duty-ratio is [17]

\[ D_{rmin} = \frac{n[M(D)]_{min}}{\eta_c} = \frac{2.18 \times 0.0533}{0.85} = 0.1368, \] (6.14)

\[ D_{rmax} = \frac{n[M(D)]_{max}}{\eta_c} = \frac{2.18 \times 0.1818}{0.85} = 0.4663. \] (6.15)

The minimum filter inductance of the converter to operate in CCM is determined by [17]

\[ L_{fmin} = \frac{R_{Lmax}(1 - D_{rmin})}{2f_s} = \frac{66.67 \times (1 - 0.136)}{2 \times 100 \times 10^3} = 289.137 \, \mu H. \] (6.16)

Pick \( L_f = 345 \, \mu H \). The peak-to-peak filter inductor current ripple is [1], [17]

\[ \Delta i_{Lfmax} = \frac{V_O(1 - D_{rmin})}{f_s L_f} = \frac{20 \times (1 - 0.136)}{100 \times 10^3 \times 345 \times 10^{-6}} = 0.5005A. \] (6.17)

The minimum peak-to-peak filter inductor current ripple is given as [1]

\[ \Delta i_{Lfmin} = \frac{V_O(1 - D_{rmax})}{f_s L_f} = \frac{20 \times (1 - 0.4633)}{100 \times 10^3 \times 345 \times 10^{-6}} = 0.3094A. \] (6.18)
The two rectifier diodes have current stresses given by [1]

\[ \Delta I_{Dr1\text{max}} = \Delta I_{Dr2\text{max}} = I_{O\text{max}} + \frac{\Delta i_{Lf\text{max}}}{2} = 3 + \frac{0.5783}{2} = 3.2505 \text{ A.} \]  

(6.19)

The ideal transformer maximum current through the primary winding is given as [1]

\[ I_{p\text{max}} = \frac{I_{Dr1\text{max}}}{n} = \frac{3.2502}{2.18} = 1.4909 \text{ A.} \]  

(6.20)

The magnetizing current is assumed as 10 % of the peak current through the transformer primary winding [1]

\[ \Delta i_{L\text{mag}(\text{max})} = 0.1 \times I_{p\text{max}} = 0.1 \times 1.4909 = 0.149 \text{ A.} \]  

(6.21)

The minimum value of the magnetizing inductance can be calculated as [1]

\[ L_{\text{mag}(\text{min})} = \frac{D_{r\text{min}} V_{S\text{max}}}{f_s \Delta i_{L\text{mag}(\text{max})}} = \frac{0.1368 \times 375}{100 \times 10^3 \times 0.149} = 3.44 \text{ mH.} \]  

(6.22)

Choose \( L_{\text{mag}} = 4 \text{ mH.} \) From basic magnetic theory, one can calculate the energy stored in the transformer magnetic field which is given by [17]

\[ W_{\text{max}} = \frac{1}{2} L_{\text{mag}} \Delta i_{L\text{mag}(\text{max})}^2 = \frac{1}{2} \times 4 \times 10^{-3} \times 0.149^2 = 44.4 \mu\text{J.} \]  

(6.23)

### 6.7 Selecting a Core

For selecting a core, we assume some factors [18]. Core window utilization factor \( k_u = 0.3 \), maximum current density \( J_{\text{max}} = 5 \text{ A/mm}^2 \) and \( B_{\text{peak}} = 0.2 \text{ T.} \) From equation (6.6), core area product can be calculated as

\[ A_p = \frac{4W_{\text{max}}}{k_u J_{\text{max}} B_{\text{peak}}} = \frac{4 \times 44.4 \times 10^{-6}}{0.3 \times 5 \times 10^6 \times 0.2} = 0.06 \text{ cm}^4. \]  

(6.24)

The core selected based on the value of \( A_p \) is PQ-42614 core and type of material is P-type [17]. The core parameters are \( A_p = 0.1720 \text{ cm}^4, A_c = 0.864 \text{ cm}^2, l_c = 33.3 \text{ mm,} \)

\( V_c = 2.88 \text{ cm}^3 \) and \( \mu_{rc} = 2500 [21]. \) The core mechanical dimensions are: \( A = 2.72 \text{ cm,} \)

\( B = 0.594 \text{ cm, C = 1.9 cm, 2D = 6.7 mm, E = 2.205 cm, F = 1.22 cm} [21]. \)

From equation (6.5), core window area can be calculated as

\[ W_a = \frac{A_p}{A_c} = \frac{0.1720}{0.864} = 1.99 \text{ cm}^2. \]  

(6.25)
6.8 Selection of Transformer Primary Winding Wire

From equation (6.2) the primary winding cross-sectional area is given by

\[ A_{wp} = \frac{I_{pmax}}{J_{max}} = \frac{1.49}{5} = 0.298 \text{ mm}^2. \] (6.26)

Depending on the value of wire cross-sectional area of the primary winding, the wire selected is AWG20. The wire parameters are, \( d_{ip} = 0.812 \text{ mm}, \) \( d_{op} = 0.879 \text{ mm}, \) \( A_{wsp} = 0.5188 \text{ mm}^2 \) and \( \frac{R_{wldc}}{l_w} = 0.03323 \text{ Ω/m} \) [7]. The air gap length in the transformer core is preset to

\[ l_g = 0.1 \times 10^{-3}. \] (6.27)

The transformer primary turns can be calculated as [17]

\[ N_p = \sqrt{\left( \frac{l_g}{\mu_o A_c} \right)} = \sqrt{\left( \frac{0.1 \times 10^{-3} \times 0.95 \times 10^{-3}}{4\pi \times 10^{-7} \times 1.25 \times 10^{-4}} \right)} = \sqrt{604.7} = 24.5. \] (6.28)

Pick \( N_p = 24. \) We know that \( n = \frac{N_p}{N_s} \)

\[ N_s = \frac{N_p}{n} = \frac{24}{2.18} = 11.3 \] (6.29)

Pick \( N_s = 11. \)

6.9 Calculation of Number of Layers in Primary Winding

From magnetics data sheet, the core height is given by [20], [21]

\[ H_c = 2 \times D = 6.7 \text{ mm} \] (6.30)

For each layer, the number of turns can be a maximum of [20]

\[ N_{L1} = \frac{H_c}{d_{o1}} = \frac{6.7}{0.879} = 7.62 \] (6.31)

Pick \( N_{L1} = 7. \) For the total number of turns, the layers required would be [20]

\[ N_{Lp} = \frac{N_p}{N_{L1}} = \frac{24}{7} = 3.4. \] (6.32)
Pick \( N_{lp} = 4 \). Therefore, per each layer, the number of turns can be calculated as \[ N_{L1} = \frac{N_p}{N_{lp}} = \frac{24}{4} = 6. \] (6.33)

The magnetic flux density has a peak value given by \[ B_{\text{peak}} = \frac{\mu_0 N_p I_{1 \text{max}}}{l_g + \frac{l_c}{\mu_{rc}}} = \frac{4\pi \times 10^{-7} \times 24 \times 1.4909}{0.1 \times 10^{-3} + \left(\frac{33.3 \times 10^{-3}}{2500}\right)} = 0.3968 \, \text{T}. \] (6.34)

The ac component of the magnetic flux density has a maximum value given by \[ B_{m(\text{max})} = \frac{\mu_0 N_p \left(\frac{\Delta i_{L \text{mag}(\text{max})}}{2}\right)}{l_g + \frac{l_c}{\mu_{rc}}} = \frac{4\pi \times 10^{-7} \times 24}{0.1 \times 10^{-3} + \left(\frac{33.3 \times 10^{-3}}{2500}\right)} \times \left(\frac{0.149}{2}\right) = 0.0198 \, \text{T}. \] (6.35)

The material chosen is \( P \)-type with parameters \( a = 1.63 \), \( k = 0.0434 \) and \( b = 2.62 \) at a frequency of 100 kHz and are obtained from manufacturers data sheet \([21]\). The power loss density in the core is given by \([17]\), \([21]\)

\[ P_v = k(f_s \, \text{in kHz})^a \left(10B_{m(\text{max})} \, \text{in T}\right)^b \] (6.36)

\[ P_v = 0.0434 f_s^{1.63} \left(10B_{m(\text{max})}\right)^{2.62} = 0.0434 \times 100^{1.63} \times (10 \times 0.0198)^{2.62} = 1.14 \, \text{mW/cm}^3. \]

The core loss is \([17]\)

\[ P_c = V_c P_v = 2.88 \times 1.14 \times 10^{-3} = 3.3 \, \text{mW}. \] (6.37)

The primary winding wire length can be calculated as \([17]\)

\[ l_{wp} = N_p l_T = 24 \times 5.38 = 129.12 \, \text{cm} \] (6.38)

where \( l_T \) is the mean turn length \([17]\)

\[ l_T = \pi \left(\frac{F + E}{2}\right) = \pi \left(\frac{1.22 + 2.205}{2}\right) = 5.38 \, \text{cm}. \] (6.39)
Pick \( l_{wp} = 135 \text{ cm} \)

For the transformer primary winding, the dc and low frequency resistance and the dc and low frequency resistance power loss are given as [17]

\[
R_{wpdc} = l_{wp} \left( \frac{R_{wdc1}}{l_w} \right) = 135 \times 10^{-2} \times 0.03323 = 0.045 \Omega, \quad (6.40)
\]

and

\[
P_{wpdc} = R_{wpdc} I_{Imax}^2 = 0.045 \times 1.4909^2 = 0.0997 \text{ W} \quad (6.41)
\]

Assuming \( F_{Rph}=4 \). The total power loss in the transformer primary winding is obtained as [17]

\[
P_{wp} = F_{Rph} P_{wpdc} = 4 \times 0.0997 = 0.3988 \text{ W}. \quad (6.42)
\]

### 6.10 Selection of Transformer Secondary Winding Wire

The transformer secondary winding have the maximum current given as [17]

\[
I_{smax} = I_{Omax} + \frac{\Delta i f_{max}}{2} = 3 + \frac{0.5005}{2} = 3.2502 \text{ A} \quad (6.43)
\]

From (6.2) secondary winding wire cross-sectional area can be calculated as

\[
A_{ws} = \frac{I_{smax}}{J_{max}} = \frac{3.2502}{5} = 0.65 \text{ mm}^2. \quad (6.44)
\]

By comparing the value wire cross-sectional area of the secondary winding, the wire selected is AWG19. The parameters are \( d_{is} = 0.912 \text{ mm}, d_{os} = 0.98 \text{ mm}, A_{ws} = 0.6531 \text{ mm}^2 \) and \( \frac{R_{wdc2}}{l_w} = 0.02639 \Omega/\text{m} \) [7]. The secondary winding wire length can be calculated as [17]

\[
l_{ws} = N_s l_T = 11 \times 5.38 = 59.18 \text{ cm}. \quad (6.45)
\]

Pick \( l_{ws} = 65 \text{ cm} \).
### 6.11 Calculation of Number of Layers in Secondary Winding

For each layer, the number of turns can be a maximum of [20]

\[
N_{L2} = \frac{H_c}{d_{os}} = \frac{6.7}{0.98} = 6.86. \tag{6.46}
\]

For the whole number of turns, the layers required would be [20]

\[
N_{Ls} = \frac{N_s}{N_{L2}} = \frac{11}{6.86} = 1.603. \tag{6.47}
\]

Pick \(N_{Ls} = 2\). Each layer can have the number of turns given as [20]

\[
N_{L2} = \frac{N_s}{N_{Ls}} = \frac{11}{2} = 5.5. \tag{6.48}
\]

Pick \(N_{L2} = 6\). The transformer secondary winding DC and low frequency resistance is given by [17]

\[
R_{wsdc} = \left( \frac{R_{wdc2}}{l_w} \right) l_{ws} = 0.02639 \times 65 \times 10^{-2} = 17 \text{ mΩ}, \tag{6.49}
\]

and the dc and low frequency power loss is given by [17]

\[
P_{wsdc} = R_{wsdc}(D_{rmax}I_{Omax})^2 = 17 \times 10^{-3} \times (0.48 \times 3)^2 = 35.25 \text{ mW}. \tag{6.50}
\]

Assume \(F_{Rsh} = 2\). The power loss in the transformer secondary winding is given by [17]

\[
P_{ws} = F_{rsh}P_{wsdc} = 2 \times 35.25 \times 10^{-3} = 70 \text{ mW}. \tag{6.51}
\]

From equations (6.41) and (6.50), the dc and low frequency power loss for both the windings of the transformer can be calculated as

\[
P_{wdc} = P_{wpdc} + P_{wsdc} = 99.7 \times 10^{-3} + 35.25 \times 10^{-3} = 0.135 \text{ W}. \tag{6.52}
\]

The power loss in the two windings of the transformer [17]

\[
P_w = P_{wp} + P_{ws} = 0.3988 + 0.07 = 0.4688 \text{ W}. \tag{6.53}
\]
From equations (6.37) and (6.53) the total power loss in the transformer is given by the sum of core loss and the winding loss

\[
P_{\text{total}} = P_c + P_{wt} = 3.3 \times 10^{-3} + 0.4688 = 0.4721 \text{ W.} \tag{6.54}
\]

At maximum output power, the efficiency of the transformer is calculated as [17]

\[
\eta = \frac{P_{\text{Omax}}}{P_{\text{Omax}} + P_{\text{total}}} = \frac{60}{60 + 0.4721} = 99.22\%. \tag{6.55}
\]

The surface area of the PQ core can be calculated from [17]

\[
A_t = \pi A \left( \frac{A}{2} + 2B \right) = \pi \times 2.72 \left( \frac{2.72}{2} + 2 \times 0.594 \right) = 21.3 \text{ cm}^2. \tag{6.56}
\]

The surface power loss density is [17]

\[
\psi = \frac{P_{\text{total}}}{A_t} = \frac{0.466}{21.31} = 0.021 \text{ W/cm}^2. \tag{6.57}
\]

The rise in the temperature [17]

\[
\Delta T = 450\psi^{0.826} = 450 \times 0.021^{0.826} = 19.13\degree \text{C.} \tag{6.58}
\]

The window utilization factor is [17]

\[
k_u = \frac{N_pA_{wp} + N_sA_{ws}}{W_a} = \frac{24 \times 0.518 \times 10^{-6} + (11 \times 0.65 \times 10^{-6})}{1.99 \times 10^{-4}} = 0.1. \tag{6.59}
\]
7 Comparision of a Single-Switch and Two-Switch Forward Converters

A two-switch topology is one of the better solutions, in terms of safe transformer core reset and component stresses [7] - [16]. This chapter presents a comparision between a single-switch forward converter and two-switch forward converter component stresses with simulation results. Efficiency curves are plotted for both the converters for same specifications $V_S = 110$ V, $V_O = 20$ V, $f_s = 100$ kHz, $P_{Omax} = 60$ W. The saber sketch schematics of the two converters are included.

7.1 Transistor Voltage Stress

Figure 7.1 show the saber schematic of two-switch forward converter. The two transistors in the two-switch forward converter at the given specifications has a voltage stress exactly equal to the source voltage $V_S$ as shown in Figure 7.2. On the other hand, the single switch forward converter, whose saber schematic is shown in the Figure 7.3, has a voltage stress of $(1 + \frac{n_p}{n_t}) V_S$, which is some times more than double the input voltage as shown in Figure 7.4, depending on the $\frac{n_p}{n_t}$ ratio [1]. This is because, the tertiary winding, not always has same number of turns like the primary winding.

If there is any leakage inductance in the transformer of the single-switch forward converter, i.e, for $k < 1$, the resonance formed between the leakage inductance of the transformer and the output capacitance of the switch creates ringing as shown in the Figure 7.5, which is difficult to control in single-switch forward converter. This is a magnetic effect in the transformer which makes the voltage stress across the MOSFET unpredictable. However, in two-switch forward converter, the clamping diodes $D_{c1}$ and $D_{c2}$ perfectly clamps the ringing to the source voltage $V_S$ as shown in the Figure 7.6 and the energy is given back to the source.
Figure 7.1: Saber sketch schematic of a two-switch forward converter.
7.2 Conduction Losses Across the Switch

A two-switch forward converter has two power MOSFET switches whose drain-to-source resistance $r_{DS}$ appears to be more than $r_{DS}$ of a forward converter which uses a single switch. But, as the voltage stress on the switch in a conventional forward converter is more, compared to a two-switch one, a transistor with higher voltage blocking capability should be used. A transistor with high voltage blocking capability has more drain-to-source resistance $r_{DS}$ than total $r_{DS}$ of two lower rating transistors. For example: An international rectifier IRFPF30 n-channel power MOSFET is used as a switch in forward converter whose $r_{DS} = 3.7 \, \Omega$ [27]. In a two switch forward converter, two IRF740 power MOSFETs are used each of which has a drain-to-source resistance of $r_{DS} = 0.55 \, \Omega$ [24]. Therefore, the conduction losses of the single-switch forward converter is more than the conduction losses in a two-switch forward converter using two switches.
Figure 7.3: Saber sketch schematic of a PWM DC-DC forward converter.
Figure 7.4: Switch voltage in single-switch forward converter with no leakage inductance.

Figure 7.5: Switch voltage in single-switch forward converter with leakage inductance.
7.3 Effect of Leakage Inductance

Figures 7.4 and 7.5 shows the voltage across the MOSFET switch with and without leakage inductance. It is observed from the Figure 7.5 that switch has to withstand high-voltage stress if the transformer has any leakage inductance. Also, it is not easy to determine or estimate the ringing associated with the leakage inductance of the transformer and the output capacitance of the MOSFET. If the design of transformer has any unwanted leakage because of imperfect coupling, the parasitic resonance can burn the MOSFET switch. Therefore, while choosing a MOSFET in single-switch forward converter, a switch with high-voltage blocking is selected to ensure that it can handle the resonant phenomena. On the other hand, from the Figures 7.2 and 7.6, the two-switch forward converter has absolutely no ringing. When the two switches $M_1$ and $M_2$ are turned off, the leakage inductance present in the transformer will resonate with the output capacitances of the two MOSFETs. But, if the drain terminal of any
MOSFET tries to exceed the input voltage $V_S$, the corresponding demagnetization diode will conduct and clamps the maximum switch voltage to the input voltage $V_S$. As a result, the high-voltage ringing is completely avoided in a two-switch forward converter.

<table>
<thead>
<tr>
<th>Type</th>
<th>MOSFET Voltage Stress</th>
<th>Demagnetization diode</th>
<th>Ringing</th>
<th>Total $r_{DS}$</th>
</tr>
</thead>
<tbody>
<tr>
<td>Single-switch</td>
<td>$(1 + \frac{n_p}{n_t}) V_{S_{max}}$</td>
<td>$- (1 + \frac{n_t}{n_p}) V_{S_{max}}$</td>
<td>Un predictable</td>
<td>High</td>
</tr>
<tr>
<td>Two-switch</td>
<td>$V_{S_{max}}$</td>
<td>$-V_{S_{max}}$</td>
<td>No ringing</td>
<td>Lower</td>
</tr>
</tbody>
</table>

Table 1: Comparison of the Component Stresses.

Table 1 compares the transistor voltage stress, demagnetization diode(s) voltage stress in single-switch and two-switch forward converters. It is clearly observed that the resonant phenomenon which causes ringing is absolutely avoided in two-switch forward converter and as a result, the equivalent drain-to-source resistance is much smaller in two-switch forward converter.

7.4 Efficiency Curves

Figure 7.7 shows the efficiency $\eta_c$ vs output power $P_O$ for same specifications $V_S = 110$ V, $V_O = 20$ V, and $f_s = 100$ kHz. It is clearly seen that the two-switch forward converter has got good efficiency and at maximum output power, approximately 92%, close to the theoretically calculated efficiency which is approximately 90%. Single-switch forward converter, has got 86% efficiency at same specifications.

7.5 Advantages of Two-Switch PWM DC-DC Forward Converter Based on the Analysis

Addition of a high-side MOSFET and a clamping diode to the primary side of the transformer in forward converter brings many advantages specifically in systems requiring high voltages at the input. The advantages are listed below [9]-[13].

1. The maximum voltage stress of the two MOSFETs is clamped to the DC
Because of some abrupt load variations, if the drain voltage of the MOSFET tries to exceed the source voltage, the two demagnetization diodes $D_{c1}$ and $D_{c2}$ get forward biased, thereby clamping the voltage across the switches to $V_S$.

2. The two clamping diodes, in addition to clamping the voltage across the magnetizing inductance, they clamps the voltage spikes reflected due to the the leakage inductance too [9] - [16].

3. Neither the design nor the main transistors have to dispose the magnetizing energy. The clamping diodes arrangement automatically pumps the energy back to the input [10].

4. There is absolutely no ringing caused by the output capacitance of the two switches and the leakage inductance of the transformer because of the clamping diodes. As a result, the system noise and hence the losses will reduce improving
the efficiency [10].

5. The two-switch topology connected across the source voltage and primary winding of the transformer accomplishes the job of transformer core-reset more effectively and easily than any of the snubber circuits discussed [8], [10].

6. In single-switch forward converter, the maximum peak voltage stress across the transistor is dependant of the leakage inductance, physical arrangement of the circuit, type of switch and diode used etc. In two-switch forward converter, the maximum transistor voltage stress is $V_S$ and there is no problem of the voltage spikes due to high-frequency resonance caused due to leakage inductance [10].

7. The power loss caused by the ESR of the high side MOSFET in two-switch topology appears to cause extra dissipation of power. But, careful examination of the data sheets, the ESR of two lower rating MOSFETs is less than $r_{DS}$ of one high rated MOSFET [10].

8. As we exactly know the peak voltage stress across each MOSFETs, one can safely use the one which has lower $r_{DS}$ thereby reducing the conduction losses [10].

9. Even though, the there is a gate drive power loss for two transistors, in two-switch topology, the elemenation of leakage inductance effects, and usage of relatively low $r_{DS}$ transistors will have a good impact on the efficiency [10].

10. Other than two clamping diodes, $D_{c1}$, $D_{c2}$ and an extra high side MOSFET, there is no need for any other passive snubber components either to dispose the magnetizing energy or to control the effects of leakage inductance [10].

11. The benefit of two-switch topology is more clear in applications with high voltages at the input. This is because, high input voltage needs a winding which
should have more number of turns which increases the leakage inductance and thereby power losses [10].

12. The two-switch topology is an extended version of single switch with the addition of a MOSFET and a diode. Inspite of these many with two-switch topology, these benefits are at modest cost [10].

7.6 Drawbacks of Two-switch Forward Converter

Even though the two-switch topology offers many benefits, there are some drawbacks and limitations too. They are as follows:

1. The two-switch forward converter has two switches. High-side and a low-side switch. The low-side switch is driven with a gate pulse referenced to ground. But, the high-side switch should be driven with respect to its source. This source voltage is unpredictable and changes with load.

2. The pulse at the gate of high-side transistor must be higher at all times than its source voltage. Moreover, the driving circuit should be simple such that it should not effect the converters overall efficiency. Therefore, the two-switch topology poses real challenge to the designer to build a robust driver.

3. We cannot turn the switch ON, for more than some fixed fraction of the switching cycle because the duty cycle is limited to 50%. If duty cycle exceeds this, then the energy that was present in the transformer magnetizing inductance in the previous cycle gets added with the energy in the consequent cycles and eventually collapse.

4. Because the duty cycle is limited, we cannot operate the converter to get wide range of output voltages.
5. The two switches in two-switch forward converter experience from hard switching during turn-on and turn-off. Obviously, hard switching of the converter has comparatively low efficiency than the soft switching one [13].

7.7 Future Work

A thorough analysis presented in this thesis, is on two-switch forward converter operating in CCM. This work can be extended to a detailed analysis on the converter operating in DCM. In addition to that, this converter can be operated for wide range of input voltages automatically if a voltage mode control loop is designed. This loop has a comparator which sense the output voltage, compares the change in the output voltage with a sawtooth waveform to change the duty-ratio and thereby correcting the output voltage. Also, the losses in the transformer windings due to the effect of eddy currents can be calculated using Fourier series.
8 Bibliography

References


