

T.C.
BAHCESEHIR UNIVERSITY
GRADUATE SCHOOL OF EDUCATION
THE DEPARTMENT OF ELECTRICAL AND ELECTORNICS
ENGINEERING

Ö. BİRİNCİ

**24VDC 36W FLYBACK CONVERTER WITH EMC
IMPROVEMENT FOR RAILWAY APPLICATION**

MASTER'S THESIS
ÖZKAN BİRİNCİ

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ABSTRACT

24VDC 36W FLYBACK CONVERTER WITH EMC IMPROVEMENT FOR RAILWAY APPLICATION

Birinci, Özkan

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Nowadays, switching DC-DC power supplies are widely used in industry. The flyback converter is at the forefront of the railway industry due to its wide input voltage range at low powers, input and output isolation, reliability and simplicity of design. Due to the low efficiency of the flyback converter, soft switching and active clamping techniques are used to increase the efficiency, which results in high cost.

In this thesis, it is aimed to design a flyback converter with high efficiency according to literature, low cost according to industry, EMC filtered, DIMMABLE, compliant with EN 50155 and EN 55015 Railway standards to meet the need for led lighting in railway systems by using hard switching technique and passive RC snubber. The designed flyback converter complies with EN 50155 and TS EN 50121 Railway standards. Flyback converter with DIM feature has 77VDC-137.5VDC input voltage range, 24 VDC output voltage, 20VDC output voltage in DIM mode, 1.5A output current, 36W power and approximately 90% efficiency.

Key Words: DC-DC Converter, EMC Filter, Flyback Converter, Railway Application

ÖZ

24VDC 36W DEMİRYOLU UYGULAMASI İÇİN EMC İYİLEŞTİRMELİ FLYBACK DÖNÜŞTÜRÜCÜ

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Günümüzde anahtarlama DC-DC güç kaynakları endüstride yaygın bir şekilde kullanılmaktadır. Düşük güçlerde geniş giriş gerilimi aralığı, giriş ve çıkış izolasyonu, güvenilirlik ve tasarım sadeliği nedeniyle flyback dönüştürücü railway endüstrisinde ön planda yer almaktadır. Flyback dönüştürücü verimi düşük olması nedeniyle yumuşak anahtarlama ve aktif kenetleme teknikleriyle verimi yükseltilmektedir. Bu durum yüksek maliyete neden olmaktadır.

Bu tez çalışmasında sert anahtarlama tekniği ve pasif RC snubber kullanılarak literature göre yüksek verimli, endüstriye göre düşük maliyetli railway sistemlerinde led aydınlatma ihtiyacını karşılamak için EMC filtreli, DIMMABLE özellikli, EN 50155 ve EN 55015 Railway standartlarına uygun flyback dönüştürücü tasarımı amaçlanmaktadır. Tasarlanan flyback dönüştürücü DIM özellikli 77VDC-137.5VDC giriş gerilim aralığında, 24 VDC çıkış gerilimi, DIM modunda 20VDC çıkış gerilimi, 1.5A çıkış akımı, 36W gücü ve yaklaşık 90% verime sahiptir.

Anahtar Kelimeler: DC-DC Dönüştürücü, EMC Filtresi, Flyback Dönüştürücü, Demiryolu Uygulaması

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LIST OF SYMBOLS & ABBREVIATIONS

AC	Alternating Current
ACF	Active Clamp Flyback
AHB	Asymmetric Half Bridge
CCM	Continuous Conduction Mode
CMC	Current Mode Control
CMOS	Complementary Metal Oxide Semiconductor
COT	Constant Operating Time
DAB	Double Active Bridge
DC	Direct Current
DCM	Discontinuous Conduction Mode
EMC	Electromagnetic Compatibility
EMI	Electromagnetic Interference
EMC	Electromagnetic Compatibility
ESR	Equivalent Series Resistance
ESL	Equivalent Series Inductance
FFT	Fast Fourier Transform
FLC	Fuzzy Logic Control
GaN	Gallium Nitride
IBC	Interleaved Boost Converter
IGBT	Insulated Gate, Bipolar, Transistor
LED	Light Emitting Diode

MPC	Model Predictive Control
PBA	Printed Board Assembly
PCB	Printed Circuit Board
PFM	Pulse Frequency Mode
PI	Proportional Integral
PID	Proportional Integral Derivative
PV	Photovoltaic
PVE	Photovoltaic Emulator
PWM	Pulse Width Modulation
RCD	Resistor, Capacitor, Diode
SiC	Silicon Carbide
SMPS	Switch Mode Power Supply
THD	Total Harmonic Distortion
TVS	Threshold Voltage Start Up
ZCS	Zero Current Switching
ZVS	Zero Voltage Switching
A_e	Effective Area
A_L	Nominal Inductance
B	Magnetic Flux Density
C_{Lp}	Transformer Primary Leakage Inductance Capacitance
C_o	Output Filter Capacitor
C_{oss}	MOSFET Output Capacitance
C_s	Snubber Capacitor

C_{sw}	Snubber Test Capacitor
C_{CM}^{min}	EMC Common Mode Capacitance
D	Duty Cycle
f_s	Snubber Ringing Frequency
f_{sw}	Switching Frequency
f_{sw}^{max}	Maximum Switching Frequency
f_{co}^{LC}	LC Filter Cut-off Frequency
f_{co}^{EMC}	EMC Filter Cut-off Frequency
H	Magnetic Field Strength
I_{C_o}	Output Filter Capacitor Current
$I_{C_o}^{min}$	Output Filter Capacitor Minimum Current
$I_{C_o}^{max}$	Output Filter Capacitor Maximum Current
I_D	Output Diode Current
I_D^{min}	Output Diode Minimum Current
I_D^{max}	Output Diode Maximum Current
I_{DS}	MOSFET Drain-Source Current
I_{DS}^{max}	MOSFET Drain-Source Maximum Current
I_{in}	Flyback Converter Input Current
I_{out}	Flyback Converter Output Current
I_p	Transformer Primary Current
I_p^{max}	Transformer Maximum Primary Current
I_p^{min}	Transformer Minimum Primary Current

I_p^{ripple}	Transformer Primary Ripple Current
I_p^{mean}	Transformer Primary Average Current
I_p^{RMS}	Transformer Primary Effective Current
I_p^{peak}	Transformer Pimer Peak Current
I_s	Transformer Secondary Current
I_s^{max}	Transformer Maximum Secondary Current
I_s^{min}	Transformer Minimum Secondary Current
I_s^m	Transformer Secondary Decreasing Current
I_s^{peak}	Transformer Secondary Current Peak Value
I_s^{RMS}	Transformer Primary Effective Current
I_{sw}^{min}	Minimum Switching Current Limit
I_{sw}^{max}	Maximum Switching Current Limit
I_t^{OFF}	Switching Element Off State Current
L_{ag}	Core Air Range
L_{DM}^{max}	EMC Differential Mode Enductance
L_{LC}	LC Filter Inductance Value
L_p	Transformer Primary Magnetisation Inductance
L_p^{leak}	Transformer Primary Leakage Inductance
L_p^{max}	Transformer Maximum Primary Magnetisation Inductance
L_s	Transformer Secondary Magnetisation Inductance
L_t	Transformer Tertiary Winding Magnetisation Inductance
N_p	Transformer Primary Winding Number

N_{PS}	Transformer Primary Winding Coefficient
N_s	Transformer Secondary Winding Number
N_T	Transformer Tertiary Winding Number
N_{TS}	Transformer Tertiary Winding Coefficient
P_D^{loss}	Output Diode Power Loss
P_{in}	Flyback Converter Input Power
P_M^{loss}	MOSFET Total Power Loss
P_{MC}^{loss}	MOSFET Conduction State Power Loss
P_{MSW}^{loss}	MOSFET Switching Power Loss
P_{out}	Flyback Converter Output Power
P_s	Snubber Power Dissipation
R_C^{ESR}	Output Filter Capacitor Equivalent Resistance
R_{DS}^{on}	MOSFET Internal Resistance
R_L	Load Resistance
R_s	Snubber Resistance
R_{sns}	Current Limiting SENSE Resistor
T	Core Induction (Tesla)
t_{bt}	Backup Time
t_f	MOSFET Turn-Off Fall Time
t_{ON}	Switching Element Conduction State
t_{OFF}	Switching Element Cut-off State
t_{off}^{min}	MOSFET Off Time Value

t_r	MOSFET Turn-On Rise Time
T_{ring}^{sn}	Snubber Ringing Period
T_{ring}	Ringing Period Without Snubbbber
t_{th}	MOSFET Conduction Time Limit
V_{cs}	Clamp&Snubber Peak Voltage
V_D	Output Diode Voltage
V_D^{min}	Output Diode Minimum Voltage
V_D^{tOFF}	Output Diode Cut-Off Mode
V_{DS}	MOSFET Drain-Source Voltage
V_{DS}^{min}	MOSFET Drain-Source Minimum Voltage
V_f	Output Diode Forward Voltage
V_{in}	Flyback Converter Input Voltage
V_{in}^{max}	Maximum Input Voltage
V_{in}^{nom}	Nominal Input Voltage
V_{in}^{min}	Minimum Input Voltage
V_M^{ring}	MOSFET Ringing Voltage at Maximum Input Voltage
V_{out}	Flyback Converter Output Voltage
V_{out}^{ripple}	Output Voltage Ripple
V_p^{max}	Transformer Maximum Primary Voltage
V_p^{min}	Transformer Minimum Primary Voltage
V_p^{leak}	Transformer Primary Winding Leakage Inductance
V_{rf}	Reflection Voltage

V_s	Transformer Secondary Voltage
V_s^{max}	Transformer Maximum Secondary Voltage
V_s^{min}	Transformer Minimum Secondary Voltage
V_{sw}^{ripple}	Switching Element Maximum Voltage Stress
V_{sw}	Switching Element Drain-Source Voltage
V_s^{ripple}	Snubber Ringing Voltage
Z_p^{leak}	Transformer Primary Leakage Inductance Impedance
ΔI_{D-S}	Switching Element Drain-Source Current Variation
ΔV_{out}	Output Voltage Ripple Variation
$\Delta V_{C_o}^{ESR}$	Output Filter Capacitor Equivalent Resistance Variation
$\emptyset$	Magnetic Flux
η	Flyback Converter Efficiency
μ_0	Core Magnetic Permeability Coefficient

Chapter 1

Introduction

1.1 Overview

In electrical or electronic systems, the power supply is the most important component to provide power to the system. Different DC-DC converter topologies have been developed to meet the power requirements of electronic systems. Although different DC-DC converter topologies have been developed in recent years, flyback topology continues to be popular in the industry today due to its wide input voltage range, galvanic isolation, multiple output voltages, simplicity, ease of design, converter efficiency in the range of 70%-80% and relatively low cost requirements. The working principle of the flyback converter is to isolate the input and output line using a transformer while converting the input voltage value to a lower output voltage. The transformer is of vital importance on the converter. The transformer usually has two windings, while the primary winding is connected to the power supply and switching element. The secondary winding adjusts the output voltage according to the turns ratio and transformers with multiple secondary windings can give more than one output voltage.

With the development of lighting technology, LED luminaires are used today instead of fluorescent lamps for environmental lighting, office and home ceiling lighting, vehicle interior and exterior lighting due to low energy, high illumination area and low cost. In the industry, flyback converter topology is mostly used in the structure of LED drivers due to long life and cost. By adding DIM feature to the designed converter, the lighting level can be controlled. LED driver with DIM feature changes the visual field appearance, saves energy and increases the demand by users by providing application flexibility.

Cost, power losses and temperature performance play an important role in the design of the flyback converter. Considering the three factors, it is necessary to understand the flyback structure well, to estimate the power losses, temperature drops on the components and the cost. In addition, since the converter is high frequency due to its structure, the side components are small in size while high frequency causes EMI problems.

The switching element is switched on and off when current or voltage is applied to the switching element in the converter. Since the switching time on the switching element is not an instantaneous period, current and voltage overlap each other, losses occur on the switching element. Therefore, as the switching frequency increases, the loss on the switching element increases. This condition is called hard switching. (Rahman et al, 2021) Soft switching technique is used to reduce switching loss in high frequency converters. (Rahman et al, 2021) Thus, at high switching frequencies, converters can be used with small size and low switching losses.

Ringing occurs when the switching element is switched on and off. These resonances cause EMI problems. Snubber and clamp circuits are used to eliminate the ringing on the switching element. Snubber circuits are simple, useful, low cost and easily absorb energy due to the resistance used in its structure. In flyback converter, snubber and clamp circuits are also used on the output diode according to EMI, efficiency and thermal problems.

In the flyback transformer, there are losses in the core. Energy accumulates in the leakage inductance on the transformer and causes voltage fluctuation when the switching element switches to cut-off mode. Converter efficiency decreases due to voltage fluctuation. In addition, the leakage inductances in the transformer bring EMI problems and departure from the safe operating point due to some parasitic elements in the circuit. In order to reduce the losses on the transformer, it is necessary to estimate and calculate the parameters close to practice, while the winding of the transformer windings should be very well wound.

1.2 Problem Statement

Nowadays, flyback converters continue to be popular, while the cost increases to achieve high efficiency. The converter efficiency decreases when hard switching is performed on the switching element of the flyback converter. In order to increase the converter efficiency, techniques such as soft switching technique, increasing the printed circuit board (PCB) layers are applied. Soft switching technique increases the efficiency by reducing switching losses while active clamp circuits are applied both on the switching element and on the output diode, which increases the number of components and brings the cost problem. When a hard switching technique is used in the converter, the capacitive effects of the switching element, the leakage inductances of the transformer and other parasitic side components cause a high EMI problem and the output voltage fluctuation is very high. This damages the load connected to the converter and makes the converter useless. Generally, DIM converters are not in high demand in the industry due to cost. There is a high cost to make the converter suitable for the industry by eliminating the mentioned problems. The high cost reduces the demand for the converter.

1.3 Literature Review

Switched DC-DC converters are widely used in industry and are used in a variety of applications. DC-DC converters obtain a constant and regulated DC voltage by minimizing power loss from a variable DC voltage. DC-DC converters are classified as insulated and non-isolated. In converters, isolation is created by connecting the input ground of the converter to the output ground. The converters that are widely used today and known as conventional converters in the literature are buck, boost, buck-boost and flyback topologies. In this section, studies on traditional DC-DC converter topologies are reviewed.

The buck converter is a non-isolated converter. It produces a regulated output voltage lower than the input voltage. Buck converter has semiconductor switching element, inductance, diode and filter capacitor. It works with the principle of discharging the inductance by the load while the source voltage charges the inductance of the converter. The buck converter has two modes of operation continuous conduction mode (CCM), discontinuous conduction mode (DCM), and these modes are determined by the current in the inductance of the converter. In the buck converter topology, the output voltage level is obtained by controlling the duty ratio of the PWM (Pulse Width Modulation) signal of the buck converter switch. Buck converter LED drivers are used in wide application areas such as DC motor speed control, regulated power supplies, television and telephone. (Sevim & Bozkurt, 2021)

The boost converter provides a regulated output voltage higher than the input voltage. In other words, because the rectified input voltage level is lower than the output voltage, it works inversely with respect to the buck converter. The boost converter works with the principle of increasing and releasing the magnetic energy accumulated in the inductor it has. The boost converter has two operating modes, continuous conduction mode and discontinuous conduction mode. The CCM mode of the step-up converter is commonly used because of its high efficiency, but the DCM mode may also be preferred due to the use of small inductors. The boost converter is commonly used in milliwatt-to-megawatt power conversion applications. (Forouzesh & Siwakoti, 2017)

The flyback converter is the isolated version of the buck-boost converter. Therefore, the input ground is isolated from the output ground by a transformer. It is generally used in applications below 100W and at high frequencies. (Gerstner et al, 2021) Flyback converter has semiconductor switching element, diode, capacitor and the most important element is transformer. The flyback converter has two operating modes, continuous conduction mode and discontinuous conduction mode, and these modes are determined by the magnetizing current of the transformer. It is widely used in daily used electronic devices such as batteries of electric vehicles, chargers, computer power supplies. (Bektaş et al, 2018)

The performance of the flyback converter is investigated in conduction mode and noise sources have been identified. The supply connected to the flyback converter in this study has a maximum of 22VDC, 17mA current and 22mW power. In steady state analysis it is operated in discontinuous conduction mode and the reasons are explained (higher efficiency, faster transient response, smaller transformer). Within the scope of the study, an EMI filter and snubber circuit were designed on the flyback converter, and the converter was isolated and designed in accordance with certain safety rules. Noise sources that cause performance degradation of this converter have been identified and their effects have been reduced. (Mohammed & Nafie, 2015)

Soft start isolated bidirectional full bridge flyback converter for high power applications is investigated. Within the scope of the study, three different prototypes with RCD snubber, active compression and flyback snubber circuits proposed in working condition have been designed, and the designed bidirectional flyback converter has 48VDC low edge voltage, 360VDC high edge voltage, 1.5kW power and 25 kHz switching frequency. It has been observed that the efficiency of the designed insulated bidirectional full bridge flyback converter is between 90-92%. As a result of the study, it has been reported that the voltage changes caused by the current difference between the flyback snubber circuit, the current fed inductor and the transformer can be alleviated. (Wu et al, 2010)

A closed loop flyback converter with 9VDC-24VDC input voltage, 3.3V output voltage and 1A output current with 100kHz operating frequency has been applied. Required parameters for flyback converter design have been calculated and simulated considering that it is recommended for computer power supply units and will be useful for portable devices that need low voltage requirements. PI (Proportional-Integral) controller was used in the simulation design. The simulation results were found to be consistent. (Martin, 2020)

The flyback transducer was analyzed and mathematically modeled, the transducer was designed according to the parameters suitable for the modeling and tested with the simulation of the designed transducer. The converter designed has an input voltage of 50-70VDC, an output voltage of 6VDC, an output current of 0.2A and a switching frequency of 65kHz. It has been reported that the converter designed by

analyzing within the scope of the study can be used for high frequency converters used in low power applications. (Coruh et al, 2010)

An asymmetric half-bridge (AHB) flyback converter is proposed where the asymmetric half-bridge flyback topology provides a high margin for the transformer turn ratio. Active clamp rotation and symmetrical half-bridge rotation steady state and operating principles analyzes were examined, and the AHB flyback converter has lower voltage stress than ACF. Experimental studies have been carried out on the prototype with 19.3VDC output voltage in the 1MHz, 65W, 100-250VDC input range and it has been observed that the peak efficiency is 96.5% and the system efficiency is over 93% at 127VDC input voltage, resulting in high efficiency in all input voltage ranges and high frequency. (Ouyang & Andersen, 2020)

Flyback design is made with M57962L IGBT driver and its performance is explained. The flyback converter designed using CCM, which is one of the steady-state states in the study has +15VDC input voltage, -8VDC output voltage, 150 mA output current, 75kHz switching frequency. The output voltage is analyzed using the Zener diode and the TL431 regulator separately to the supply circuit of the converter. It has been reported that the output voltage of the TL431 regulator is 5% better than the zener diode, and the performance of the flyback converter is sufficient to drive the IGBT driver. (Kumar & Pachori, 2020)

A study was conducted on the use of SiC MOSFET in high input voltage controlled power supply design using flyback topology, simulation techniques for SiC modeling were discussed and it was concluded that the difficulties in design caused cost increase. An RCD snubber circuit has been used in the flyback converter design, and the converter has a switching frequency of 80kHz, an input voltage range of 320-1000VDC, an output voltage of 24VDC, 15VDC and two 5VDC, a total of 4 output voltages. According to the analysis, SiC MOSFET is recommended for control power supplies with high input voltage in renewable energy applications. (Jagannath et al, 2020)

The analysis and design of the flyback converter, which gives multiple output voltages using the flyback topology, is examined. The designed flyback converter

provides 220VAC input voltage, 180VDC and 5VDC multiple output voltages. The flyback converter is proposed to power the Nixie Tube and 5VDC microcontroller, while the full load analysis is done with calculations for continuous conduction mode (CCM) and discontinuous conduction mode (DCM) analysis. Closed-loop control of the converter is performed using a PI control block with 220VAC rectified input voltage and 5VDC reference voltage, while for 5VDC output voltage, closed-loop using a PID (Proportional-Integral-Derivative) Controller with $2.5\ \Omega$ resistive load the control shows the desired output voltage. (Gupta et al, 2020)

It is explained that SiC MOSFETs are more efficient than Silicon (Si) MOSFETs due to low transmission and switching losses, by using multi-output flyback converter designed and using SiC MOSFET due to low power loss. RCD clamp circuit is designed using steady-state discontinuous conduction mode. The multi-output flyback converter, designed and simulated in practice, has 250-850VDC input voltage, two 24VDC output voltages, two 1A output currents, 20kHz switching frequency and 24W output power. The designed converter was simulated and implemented as hardware. Designed in the 250VDC-850VDC input voltage range, the converter provided 85.14% efficiency at low input voltage and 76.02% at high input voltage. The ripple voltage of the designed multi-output flyback converter is less than 1% and the output voltage regulation is reported to be good, while the converter is efficient using SiC MOSFET and the designed converter is provided for general power units, industrial fields and aerospace technology. (Nagesha et al, 2015)

A study has been made about the asymmetric half-bridge flyback converter and it has been emphasized that the control method is clear, simple and reliable by using hybrid switching technique for the converter. It has been stated that the asymmetric half-bridge flyback converter has advantages such as fewer components, less cost, higher power and efficiency, and in the simulation created in this study to evaluate the theoretical information, the converter has an input voltage range of 90VDC-270VDC, 19VDC output voltage, 0.8566A-3, output current range of 42A, a switching frequency of 100kHz, and an output power of 65W. The simulation results showed that the hybrid switching technique proposed in the study can be applied on the flyback converter. (Chen et al, 2016)

A miniature converter with a frequency of 400kHz and multiple outputs of 5VDC, 15VDC, 5VDC is designed with current mode controlled flyback topology. It has 22VDC-36VDC input voltage range, -5VDC/0.2A, 15VDC/0.7A, 5VDC/4A output voltage and current. While the designed flyback converter is experimentally verified with 70% efficiency, a soft start circuit is applied to reduce the voltage on the switching element MOSFET. (Chandra et al, 2015)

Owing to the voltage equalization need of battery series, the flyback converter was designed as a voltage equalizer methodology. Designed for voltage equalization of four 12V battery series, the flyback converter has 48VDC input voltage, 12VDC output voltage, 50kHz switching frequency and 100W power. A prototype was built with two outputs for the converter theoretical analysis and four outputs to validate the theoretical analysis. The activity stages, waveforms, design parameters and experimental results of the converter during the voltage stabilization of the battery series are given. It has been reported from the experimental results that the voltage stabilization of the battery series is provided by its own voltage stabilization feature in the converter DCM operating state, while the output voltages of the Flyback converter in the DCM operating state are balanced. (Costa et al, 2018)

Design information about the flyback power converter, which is designed according to the high voltage impulse load requirements, is presented, and it is also used to charge the energy storage capacitor that meets the flyback topology, charging time and voltage level requirements. Peak current mode and boundary conduction mode methods are used to charge the energy storage capacitor. In this study, a 5W high-voltage flyback converter prototype and simulation was made to generate pulses with 500Hz and 1kHz pulse repetition and 1kV voltage. The simulation results confirming the designed flyback converter are presented while it is explained that the efficiency of the flyback converter decreases due to the leakage inductance in the converter, the winding capacitance delays the charging time, and the leakage inductance increases the energy level lost when the power level increases. (Ravi et al, 2020)

In this article, it is studied about lossless snubber and flyback converter with base level transformer. An experimental prototype was built using a bipolar transistor with

a voltage of 6V and a current of 900mA. In the converter, the RCD snubber circuit is analyzed lossless, while at full load, the DCM operates in steady state mode. In the worst case, it has been observed that the average efficiency is about 69%, the lossless efficiency is in the range of 63%-75%, and the efficiency measured by RCD is in the range of 58%-62%. In the tests carried out on the experimental prototype, the theoretical analysis was confirmed, it was reported that the cost of the base level transformer was reduced by about 10% of the design cost, and the design is provided for low-power devices, mobile phone chargers and energy-saving applications. (Zhao et al, 2012)

An analysis is made by constructing a state space model using the transfer function while the flyback converter is operating in continuous conduction mode. Steady-state continuous conduction mode was used for design analysis, while the mean was modeled using converter, small AC signal modeling, power circuit and control parameters. While the designed flyback converter has 110-250VDC input voltage range, 48VDC output voltage, 50kHz switching frequency and 250W output power, it has been reported that the converter is unstable in continuous conduction mode in the simulation results. (T. Halder, 2020)

Simulations were made on the boost converter topology and transferred to the PCB (Printed Circuit Board) in practice. High efficiency is achieved by applying a high frequency (up to 500 kHz) GaN and SiC based DC-DC boost converter, ZVS (Zero Voltage Switched) two-phase boost converter topology. By investigating the feasibility of interspersed ZVS boost converters providing high frequency operation with a power density exceeding 10kW/dm³, it was found that careful design, especially thermal considerations, is required for the converter to achieve high power density. Despite simulations suggesting that it is possible to achieve high power density by experimentally designed, simulated, and validated the converter, problems have been identified that limit the output power of the converter. (Zimoch et al, 2019)

A soft-switching boost converter with a new simple passive lossless suppressor is proposed, offering the advantages of a minimum number of components, low cost, and low voltage stress on the switch. It is seen that the proposed converter has better efficiency and EMI values than other boost converters while experiments have been

made with the 220W prototype and it has been reported that the efficiency of the converter is 3% better than the efficiency of conventional boost converters. (Shamsi et al, 2012)

The boost converter, which converts 3.3VDC input voltage to 5VDC output voltage, is provided for use with mobile devices. The converter with Soft-Start feature has 100mA output current, 1MHz switching frequency, and the switching frequency is quite high. The control of the proposed step-up converter consists of a block that controls the on and off state of the switch. (Mo et al, 2017)

A new design approach is presented that provides very high conversion efficiency in low voltage, high power isolated boost DC-DC converters. Detailed test results of the 1.5-kW full-bridge reinforced dc-dc converter confirm the theoretical analysis and show very high conversion efficiency while worst case efficiency at maximum load and minimum input voltage is 96.8% while the maximum efficiency of the proposed converter is 98%. It is recommended for use in fuel cell applications. (Nyman & Andersen, 2010)

A study has been made about the Interleaved Boost Converter (IBC) topology and it is recommended for renewable energy applications. Interleaved Boost Converter has advantages over traditional buck converter, such as high efficiency, low input current ripple, electromagnetic emission, high voltage boost, low output voltage ripple and low switching loss. Within the scope of the study, three interspersed amplifier transducers, uncoupled, coupled and reverse coupled, were designed, analyzed and simulated. The simulation results of all three converters were compared and it was observed that the Coupled converter reduced the current fluctuations compared to the non-coupled inductors, while the observations made that the IBC gave low input current fluctuation for fuel cell applications. (Rahavi et al, 2012)

Boost converter design, simulation and experimental prototype were made for use in PV module applications, and two separate controllers were used to control the duty cycle of the designed converter. The output voltage of the converter was controlled using Proportional-Integral and Fuzzy Logic Control (FLC) in the converter. The input voltage of the designed boost converter is 21VDC, the output

voltage calculated according to the equations is 24VDC, the simulation result is 24.7VDC, and the experimental prototype is 28VDC. It has been observed that there is little difference between the experimental prototype results and the simulation results, but it has been reported that the FLC controlled transducer has better performance and can respond with high speed and accuracy compared to the PI controlled transducer. (Mohamed et al, 2016)

The controller was designed and simulated by modeling the boost converter for use in renewable energy applications. A Proportional-Integral controller is designed so that the converter, which is modeled by using the DC-DC boost converter, small signal analysis and transfer function, can accept the load change and give a constant output voltage. The PI controller root positioning method was used, and it was presented that the closed loop performance of the PI controller boost converter designed with the root position method by performing controller simulation in the MATLAB/SIMULINK environment is better than the closed loop performance of the uncontrolled boost converter. It has been reported that the PI controller boost converter designed with the Root Position Method improves the system response and the designed controller can be used in different applications. (Shadoul et al, 2017)

DC-DC step-up converter is designed and it is proposed to connect the designed converter to the PV system for maximum power point monitoring. Soft switching is used for continuous conduction mode and discontinuous conduction mode while steady-state situations are analyzed and the designed converter load is simulated by changing it to 12kW, 200kW and 500kW. In the simulation results, it has been observed that the converter works well and it is reported that the output voltage changes of the converter can be solved with the inverter control strategy. (Hasaneen & Mohammed, 2008)

A new Booster Interleaved Boost converter with high efficiency and low voltage stress has been proposed by conducting a study on DC-DC boost converter for renewable energy system fuel cell and photovoltaic system applications. Interspersed converter structure is used to reduce output voltage fluctuations and input current fluctuations. The proposed converter combines a flyback converter with switched capacitors, the flyback converter is used to avoid high duty ratio and expand the

amplifier gain of the boost converter. Continuous conduction mode is used by performing steady-state analysis and it is reported that continuous conduction mode is more suitable than discontinuous conduction mode. An experimental prototype has been made and analyzed while the designed Boost converter has 24VDC input voltage, 200VDC output voltage, 50kHz switching frequency and 200W power. Experimental results are reported to verify the analysis and design of the designed transducer, observing that the duty cycle of the transducer is low and the travel time is short in the prototype results. (Ling et al, 2015)

A 175W multi-output SMPS is modeled and simulated by examining the power quality of the multi-output forward boost switched-mode power supply and calculating the comprehensive design parameters of the advanced boost converter. In the study, the input voltage changes of the modeled converter, the effect of the load line change on the converter and its performance were evaluated. Output voltage is controlled using Proportional-Integral (PI) by making closed loop control. The investigated forward boost converter is 220Vrms 50Hz input voltage, 5VDC, 12VDC, -5VDC, -12VDC output voltages, 18A, 6A, 0.3A, 0.8A output currents, respectively. It has 175W output power and 50kHz switching frequency. In the presented advanced boost converter, it is stated that the total harmonic distortion (THD) input current drops to 10% at 50% load and the supply voltages are in the range of 170V- under all operating conditions at full load, while the output voltage fluctuations are in the order of 270V, and output voltage fluctuations are in the order of 2%. (Singh et al, 2011)

A boost converter that can start at the mean threshold voltage has been designed. The threshold voltage initiation (TVS) for the designed transducer is examined, the transducer is fabricated, tested, and the results are presented. This converter is designed with CMOS (Complementary metal oxide semiconductor) process, also the converter has a fluctuating input voltage range of 0.15VDC – 0.9VDC, output voltage of 1.2VDC and load current range of 0.1mA – 1mA. The peak current control of the converter works in discontinuous conduction mode, reducing the switching loss at low loads, and it is stated that the switching frequency works depending on the load in PFM (pulsed frequency mode) control, and the switching frequency increases as the load value increases. As a result of the study, the designed converter results are

reported as 700kHz and 15kHz, respectively, with the highest and lowest switching frequency, 83% efficiency and 2% output ripple voltage. (Sze et al, 2008)

Model predictive control (MPC) strategy for the boost converter used in various applications such as energy storage systems, solar energy systems has been examined. The goal of the MPC strategy is to formulate a one-step prediction of behavior problems to manage the converter. With this strategy method, the inductor current is designed and referenced. The MPC created in the study has a 10VDC input voltage and a switching frequency of 20kHz. It has been reported that the converter plays an active role in maintaining the steady state operating point when the system parameters and resistive load are changed between 50% and 200% by using the observer to change the system dynamics within the scope of the study. (Cheng et al, 2016)

The controller is designed for boost converter, resistor load and unknown external input voltage and observer-based boost converter is designed. The designed boost converter has 6VDC input voltage, 200kHz switching frequency and 4W output power. It is stated that the controller of the designed boost converter is asymptotically stable in closed loop, it is experimentally modeled and tested against input voltage and load uncertainties. It has been reported that the controller and boost converter designed according to the test results are resistant to variable input voltages and load changes while the controller design presented in this study is recommended for buck and buck-boost topologies. (Oucheriah & Guo, 2013)

In this article, buck converter topology is designed, modeled and simulated for LED driver design. The buck converter designed in the study has 15VDC input voltage, 10.2VDC output voltage, 25kHz switching frequency and 700mA average output current. In the circuit presented for the designed and modeled buck converter topology, MOSFET application was made instead of diode. System use of MOSFET, it aims to increase the system efficiency to 5% and above. The converter has been simulated closed loop and open loop. It has been reported that the closed loop gives more accurate results when the closed loop and open loop data are compared in the simulation results. (Deekshitha & Shenoy, 2017)

The lower limits of the power conversion that occurs when the Buck converter is used as a voltage regulator and the feasibility of the lower power voltage regulator source are examined, and the feasibility of the Buck converter is examined by using the capacitor, inductor criteria and CMOS parameters. In the study, comparison and benchmarking were made using 0.50 μ and 0.18 μ CMOS Standard technology specifications, and the buck converter design was made using 0.50 μ m standard CMOS. The buck converter has 3.3VDC input voltage, 1.65VDC output voltage, 250mA output current, 1.023W input power, 0.417 transmit power, 400MHz operating frequency and 41.33% efficiency. It has been reported that the buck converter designed within the scope of the study cannot be used in low power voltage regulation applications when the parameters of large inductance values. (Zamora-Mejía et al, 2016)

In order to increase the heavy load efficiency, buck converter with heavy load mode and baby-buck mode buck converter is recommended in order to increase the light load efficiency. In order to reduce the switching loss, the proposed converter light load discontinuous conduction mode and constant operating time (COT) control are combined. As a result of the experiments, the designed converter has 5VDC input voltage, 1.2VDC output voltage, 10-700mA output current range, 14.6-250kHz operating frequency range and 70-80.6% efficiency; converter works fine in both modes. (Lin et al, 2020)

A controller was designed for linear and non-linear buck converters, and the controller designed for buck converter was designed using PID control technique in order to regulate the output voltage by simulating. The designed converter has 48VDC input voltage, 10VDC output voltage and 53kHz operating frequency while the buck converter whose controller is designed operates in continuous conduction mode. The PID controller design and linear and non-linear models of the closed-loop buck converter system were simulated and tested in three different ways using the simulation input voltage variation, reference voltage variation and output load variation parameters. It is proposed to follow with zero steady-state error and be implemented for a buck converter. (Kumar et al, 2021)

Digital voltage mode PI controller design for buck converter in digital control systems is examined, converter and controller are modeled, designed, simulated and experimentally analyzed on prototype. The converter designed within the scope of the study has 24VDC input voltage, 12VDC output voltage, 3A output current, 36W output power and 200kHz switching frequency. The converter designed within the scope of the study has 24VDC input voltage, 12VDC output voltage, 3A output current, 36W output power and 200kHz switching frequency while It has been reported that the designed converter operates in continuous conduction mode with higher linearity than the boost converter. The converter designed in this study is designed as an asynchronous buck converter due to the temperature factor while the designed control system has a converter, a current and a voltage sensor. The transfer function of the designed transducer was evaluated using AC-sweep analysis while within the scope of the study, it has been reported that mathematical analysis is important for the design of the system, compensator design and mathematical calculations for correct system indentation, and aggressive controllers have a faster conversion response in the change of load parameters over time. (Suskis & Nikonorov, 2018)

The internal current loop of the buck converter is investigated with the digital average current mode control (CMC) technique, and a discrete time frame is proposed for the analysis of current loop instability and current loop controller setting in digital environment. The buck converter prototype was made within the scope of the study and designed with a CMC controller using a TMS320F28335 microcontroller from Texas Instruments, the buck converter has an input voltage range of 3-6VDC, an output voltage of 1.5V and an operating frequency of 200kHz. From the experimental analysis results, it works in accordance with the proposed analysis, the FFT of the inductor current contains harmonics and is unstable in the waveform, the current loop is limited by the applied input voltage, and the CMC can be used for steady-state analysis for DC-DC converters. (Rao et al, 2020)

It is proposed to calculate and design passive components of buck converter, capacitor, output resistor and inductor for use in Photovoltaic Emulator (PVE) applications. The buck converter has a 20kHz switching frequency and a 1% output ripple factor, which is designed to ensure that the proposed converter operates

smoothly in continuous conduction mode and has a lower output voltage ripple factor than the system specifications. It has been observed that the buck converter designed with the passive components calculated within the scope of the study is positioned above zero at the minimum inductor current level and the inductor current is above zero and works smoothly in the continuous conduction mode, and it has been reported that the designed buck converter can keep the output voltage ripple factor below 1% and the system is consistent. (Ayop et al, 2020)

A buck converter design and simulation has been made using PI with variable input voltage and output load, low-order output voltage fluctuating peak. The buck converter designed in the study has 12VDC input voltage, 5VDC output voltage and 100kHz switching frequency. The buck converter designed in the study has 12VDC input voltage, 5VDC output voltage and 100kHz switching frequency while open and closed loop continuous conduction mode interrupted conduction mode analyzes were performed. It has been observed that the buck converter, designed and simulated within the scope of the study, works in accordance with the changes in the duty cycle parameters, input voltage and load resistance when the capacitance and inductance values are kept constant. (Shinde et al, 2018)

A zero current switching (ZCS) semi-resonant buck converter is proposed for use in high power applications and the parameters of the buck converter are calculated mathematically and simulated in a computer environment for testing. The control of the proposed converter is semi-resonance and pulse frequency modulation control instead of the pulse width modulation control of the traditional buck converter, and the buck converter recommended within the scope of the study has 120VDC input voltage, 80VDC output voltage, 5A output current, 400W power and 40kHz operating frequency. The efficiency of the designed buck converter is 96.3% in the simulation results and it is reported that it is 10-15% higher than conventional buck converters, while it is also said that the proposed converter can be used in airplanes, electric vehicles and scooters. (Farrakhov et al, 2020)

The output parameters of the buck converter operating in discontinuous conduction mode with pulse frequency modulation (PFM) control are examined and formulas are proposed for the estimation of the output parameters of the buck

converter. The current source approach and voltage source approach have been used to estimate the output parameters of a PFM controlled buck converter operating in DCM mode, and it has been reported that it is an accurate way to measure the spectral form of the output voltage fluctuation without requiring technical and detailed knowledge of the converter's design. Within the scope of the study, formulas designed and found about current and voltage source approaches to estimate output parameters in DCM mode were analyzed and a comparison was made with the output spectrum of a standard 2.25MHz, 600mA buck converter from Texas Instruments. The designed buck converter has 4.2VDC input voltage, 1.8VDC output voltage and 2.25MHz operating frequency. As a result of the analysis of the formulas designed and found in the study, it has been reported that low-power systems can help to predict the power supply noise parameters. (Tao & Fayed, 2012)

Comprehensive analysis and design of the buck converter for photovoltaic applications is proposed, and the simulation and experimental prototype of the buck converter, which has been analyzed and designed, is made. Buck converter fixed input voltage 80VDC and 20kHz switching frequency with a maximum output power of 164W and 60% converter efficiency has been found, and it has been reported that the losses on the MOSFET used as the switching element in the converter design and the diode used in the converter design reduce the system efficiency. (Masri et al, 2012)

In this article, a constant current buck converter is designed for LED driving application. The designed converter was analyzed and simulated in the specified topology, and an experimental prototype with dimming was made. The converter designed in the study has 16VDC input voltage, 300mA output current and 500kHz switching frequency. In the experimental results, it has been reported that the simulation and prototype of the converter are compatible, the dimming level shows the correct result by adjusting the duty cycle when the pulse width modulation (PWM) signal is changed, and the converter provides a constant output current of 300mA when the output voltage is changed. (Mrabet et al, 2020)

This article proposes a new bidirectional DC-DC converter application. The converter proposed in the study has more boost-down voltage gains than the bidirectional buck-boost converter used in the literature. In addition, it has been

suggested that the average value of the switch current is lower than that of the bidirectional buck-boost converter under the same electrical conditions as the bidirectional buck-boost converter. Experimental waveform analysis and steady-state analysis were performed and it was observed to be compatible. A prototype circuit with 14VDC input voltage, 42VDC output voltage, 50kHz switching frequency and 4.7A output current was used to determine the performance accuracy of the dual battery system used on the automobile. At full load, the boost mode efficiency is 92.7% and the down mode efficiency is 93.7%. The measured efficiency range is 92.7%–96.2% in boost mode, 93.7%–96.7% in downgrade mode. It is more efficient than the traditional bidirectional Buck/Boost converter. (Yang & Liang, 2012)

Baars et al. reported the design of three-phase double active bridge (DAB) topology and buck-boost topology for use in railway applications. There is also discussion about auxiliary inductors and hard pull mode to expand the soft switching region using soft switching. In order to reduce the losses in the system, the parameters are calculated and tested in the parameters determined while a prototype is made according to these parameters, resulting in an output power of 80 kW and an efficiency of 95.6%. (Baars et al, 2014)

A four-switch buck-boost converter has been applied for a high-current power supply, and the designed four-switch buck-boost converter has an input voltage range of 6.5-36VDC, an output voltage of 12VDC, a switching frequency of 300kHz and maximum output power of 60W. It has been reported that the switching frequency is proportional to the power losses in the designed circuit. The converter is designed for printed circuit board (PCB) applications, and a four-layer printed circuit board is designed to achieve high output current and an experimental prototype is made and the maximum efficiency of the converter in the measurement results is 99% at 1.1A output current and 97.2% at 4.5A output current. (Brusey, 2021)

A buck-boost converter has been proposed, which produces a negative output voltage with a low duty cycle and outputs a lower output voltage than the input voltage. The proposed converter has 1200VDC input voltage, -940VDC output voltage and 200kHz switching frequency by using elements with buck converter topology, applied to buck-boost topology and formed by switching element MOSFET, diode, inductor

and output capacitor. The proposed buck-boost converter was compared with the buck converter analysis and simulation. Within the scope of the study, analysis and simulation results reported that when the switching frequency of the traditional buck converter and the proposed buck-boost converter are changed, the traditional buck converter shows better efficiency and performance, and is also suitable for use in the proposed buck-boost converter line drivers and receiver applications. (Soheli et al, 2018)

Sunkara and Kondrath, presented the design procedure of DC-DC Buck-Boost converter with PWM control, bidirectional and reverse output operating in continuous conduction mode and verified by simulation. The buck-boost converter designed in the study operates at $24\text{VDC} \pm 4\text{VDC}$ input voltage, -15VDC output voltage, 1-3A output current and 100kHz switching frequency in buck mode, while it operates at -15VDC input voltage, 24VDC output voltage and 1A output current in boost mode. The total power of the designed buck-boost converter is 45W. Within the scope of the study, it has been reported that the designed converter has an efficiency between 89-96.5% and the efficiency of the converter can be reduced by using anti-parallel power diodes in the study. (Sunkara & Kondrath, 2013)

A simple buck-boost LED driver is presented for battery-powered led array applications, and in order to increase the efficiency of the buck-boost converter, the converter can be operated in two different modes, buck and boost, by soft switching on the switching element. The offered converter has 12VDC input voltage, 32VDC output voltage, 100kHz switching frequency and 26W output power. An experimental prototype of the proposed converter has been made and tested, with the converter having an efficiency of 90% at rated load and with the input diode, while its efficiency is 93.1% at full load and without the input diode. In the study, it was understood in the experimental results that the converter is suitable for practical use and can be used in applications where high efficiency and various colored LEDs are selected. (Pouladi et al, 2019)

A Zero Voltage Source (ZVS) single-switch buck-boost converter designed for photovoltaic (PV) applications is presented, and the proposed converter is used to regulate the voltage from the PV module and convert it to output voltage. The

converter simulation proposed in the study and the prototype made converter after the simulation have 36VDC input voltage, 200VDC output voltage, 240W output power and 50kHz switching frequency. It has been reported that the converter designed with ZVS soft switching without using auxiliary switching elements in the converter design and designed with the converter amplified can be used in PV applications by finding that the desired and regulated output voltage values are correct. (Kiran & Ezhilarasi 2015)

This article presents the mathematical modeling and design of bidirectional DC/DC buck-boost converter for use in electric vehicle applications. In the study, the converter designed by simulating the proposed converter using a proportional-integral (PI) controller has 200VDC input voltage, 350VDC output voltage, 40kHz switching frequency and 1.5kW output power. It was reported that the converter, which was designed and simulated in the parameters determined within the scope of the study, is suitable for electric and hybrid vehicle applications. (Vacheva et al, 2019)

Si and GaN structured switching elements of the integrated double buck-boost converter designed for use in LED applications have been investigated. The designed integrated double buck-boost converter has 220Vrms input voltage and 58.4VDC output voltage while in the study, power MOSFET and GaN transistor, which are two different switching elements in Si structure, were compared on the converter and three different switching elements added to the converter were tested at switching frequencies ranging from 50-150kHz. As a result of the switching elements tested and compared on the converter within the scope of the study, it has been reported that GaN-based switching elements operate at lower operating temperatures and are 5% more efficient than Si-based switching elements. (Duarte et al, 2017)

A transformerless single-switch buck-boost converter is designed and experimentally simulated while the input voltage of the designed converter is 11VDC, the output voltage is 26VDC in the step-up condition and the output voltage is 10VDC in the step-down condition. The converter designed by examining steady-state continuous conduction mode and discontinuous conduction mode operates in continuous conduction mode. In the simulation results of the converter designed within the scope of the study, it has been reported that the voltage gain is higher than other

conventional converters (eg conventional buck-boost converter, ZETA, CUK and SEPIC) and is suitable for applications such as LED drivers and electric vehicles. (Magar et al, 2018)

The above studies have been reviewed and in this thesis, a high efficiency Flyback converter design in accordance with EN 50155, TS EN 50121 and IP5K1 standards will be presented.

1.4 Goal of Thesis

In this thesis the scope of this thesis, a DC-DC flyback converter with 77VDC-132VDC input voltage range, 24VDC output voltage, 1.5A output current, 36W power, EMC filter, DIMMABLE feature, EN 50155 and EN 50121 railway standards has been designed to meet the lighting needs of railway systems. EN 50155 standard is required for railway applications-electronic devices used in rail vehicles. EN 50121-3-2 standard is required for electromagnetic compatibility for electronic devices used in railway applications.

According to 50155 standard requirements, vehicles in railway applications use batteries up to 110VDC to save weight and volume. However, many systems inside the railway vehicle require 12VDC and 24VDC input voltage. The main task of the DC-DC converter in the systems in the vehicles in railway applications is to convert 110VDC to 12VDC or 24VDC, while increasing 12VDC and 24VDC to 110VDC. The nominal input voltage values specified in the standard are 24VDC, 28VDC, 36VDC, 48VDC, 72VDC, 96VDC, 110VDC, but the recommended values in the standard are 24VDC, 72VDC and 110VDC. The EN50155 standard specifies $\pm 30\%$ input voltage variation including voltage ripple. The converter should have a minimum input voltage of 77VDC and a maximum input voltage of 137.5VDC at 110VDC nominal input voltage. In this case, the converter with a nominal input voltage of 110VDC must be designed in such a way that no damage or malfunction occurs in the voltage range of 77VDC and 137.5VDC.

According to the 50155 standard requirements, the operating temperature classes vary. Temperature classes OT1 -25°C to $+55^{\circ}\text{C}$ and OT2 -40°C to $+55^{\circ}\text{C}$ apply to equipment in passenger compartments and driver's cab with a reference permanent temperature of 25°C . Temperature classes OT3 -25°C to $+70^{\circ}\text{C}$ and OT4 -40°C to $+70^{\circ}\text{C}$ apply to equipment in cabins/cabins with a reference permanent temperature of 45°C . Classes OT5 and OT6 are applicable for PBA (Printed Board Assembly) in a sub-rack or in an enclosure for semiconductor drive units or combustion engine control units when required due to the internal enclosure temperature. The interior lighting device of the railway vehicle is in the interior passenger compartment of the vehicle, i.e. OT1 and OT2 temperature class, but the DC-DC flyback converter will be installed in the panel section of the cabins in the passenger compartment. For this reason, the flyback converter designed within the scope of this thesis is designed for OT4 temperature class.

According to EN 50121-3-2 standard requirements, the electromagnetic effects of the electronic device shall be arranged in such a way as to minimize the electromagnetic effect of the electronic equipment of the rolling stock. In EN 50121-3-2 standard, there should be no noise at frequencies between 150kHz-500kHz at db higher than 99 dB μ V peak and no noise at frequencies between 500kHz-30MHz at db higher than 93 dB μ V peak.

In the design of the flyback converter, it is designed to achieve higher efficiency than the literature by using a simple snubber circuit and without using an active clamp circuit, minimising switching losses, using a double-layer PCB board, reducing the output voltage fluctuation while performing hard switching. It is aimed to increase user demand compared to other railway power supplies due to its low cost, DIMMABLE feature, IP5K1 protection (protection against dust and dripping water) and high efficiency. In addition, it is aimed to present a design approach by simulating the selection of critical components in the design of flyback converters, estimating the power and heat losses that may occur, and providing low-cost, reliable and long-lasting products suitable for the industry.

1.5 Thesis Layout

General information about the flyback converter is given in Chapter 1 in the thesis while designing a low-cost flyback converter within the scope of the thesis. Afterwards, a problem statement was made and literature research and literature follow-up were carried out. During the literature research, the usage areas and applications of the flyback converter are stated and the purpose of the thesis is mentioned.

In Chapter 2, the working principle, operating modes and theoretical component formulae of the flyback converter are given. Suppression cell, transformer design, switching element and diode characteristics are mentioned.

In Section 3, the details of the proposed design are given. Converter parameters, transformer design, switching element and diode selection, snubber design, EMI filter design, information about the integrated used, LTspice circuit simulation and circuit diagram are given. Power loss estimation is made for the elements used in the converter design.

In Section 4, converter prototype test details and test results are given.

In Section 5, the test results of the prototype flyback converter are presented and recommendations are given.

Chapter 2

Theoretical Background

2.1 Flyback Converter

Among the switched DC-DC converter topologies, the Flyback converter occupies a large place in the industry. Flyback converter has become a popular and distinguished topology today due to its simple structure, low cost, small component components, galvanic isolation, versatile applications, wide input range value and wide usage area. The flyback converter is typically used for applications in the 20W-200W power rating range, while it is used in industry for applications below 100W-150W. The use of an output inductor in the flyback converter structure has a lower cost and the losses in the converter are at lower levels compared to other topologies. In this case, the changes in the load section are more efficient than the forward converters. Another important feature of the flyback converter is that it can provide multiple output voltages. (Kamil, 2007) The flyback converter is an isolated version of the typical buck-boost converter topology. The input and output sections of the converter are isolated while the transformer is energized when the switching element of the converter is in the conduction state. The flyback converter has many advantages as well as disadvantages. The transformer that the converter has has leakage effects when stray effects are neglected, the voltage stress and resonance of the switching element are high. Due to the high voltage stress, it is necessary to use a switching element that can operate at high voltage levels for the design. In this case, the converter design results in high cost. In addition, high voltage stress and ringing cause the transmission loss problem and reduce the converter efficiency. (Pesonen, 2014) The feature that distinguishes the flyback converter from the buck and boost converter is that it has a transformer that provides isolation between the input and output voltage and is used for energy storage. The flyback converter typically consists of a switching element (MOSFET), transformer, diode and filter capacitor. Fig. 1 shows the

conventional circuit diagram of the Flyback converter topology. The elements in the circuit diagram are considered ideal.

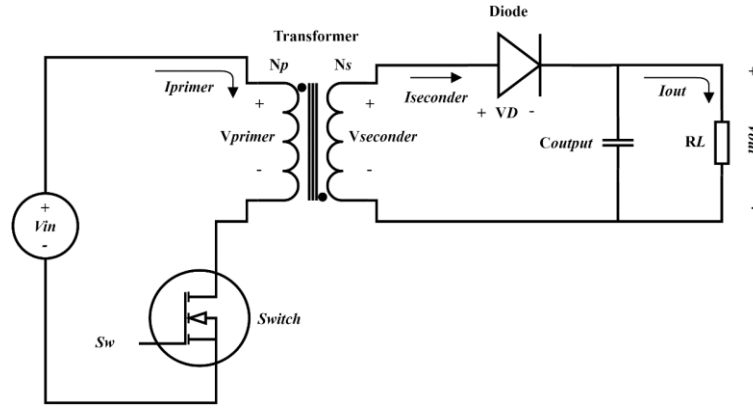


Figure 1. Conventional Flyback Converter Circuit Diagram

In the conventional flyback converter circuit diagram shown in Fig. 1, V_{in} supply voltage, I_{primer} transformer primary current, N_p primary winding number, N_s transformer secondary winding number, V_{primer} transformer primary voltage, S_w switching element MOSFET, $V_{seconder}$ transformer secondary voltage, $I_{seconder}$ transformer secondary current, V_D the output diode voltage, C_{output} the output filter capacitor, the output current of the I_{out} converter, the load at the output of the R_L converter, and the output voltage of the V_{out} represents converter.

Fig. 2 shows the waveforms generated by the switch of the Flyback converter in the turn on and off state.

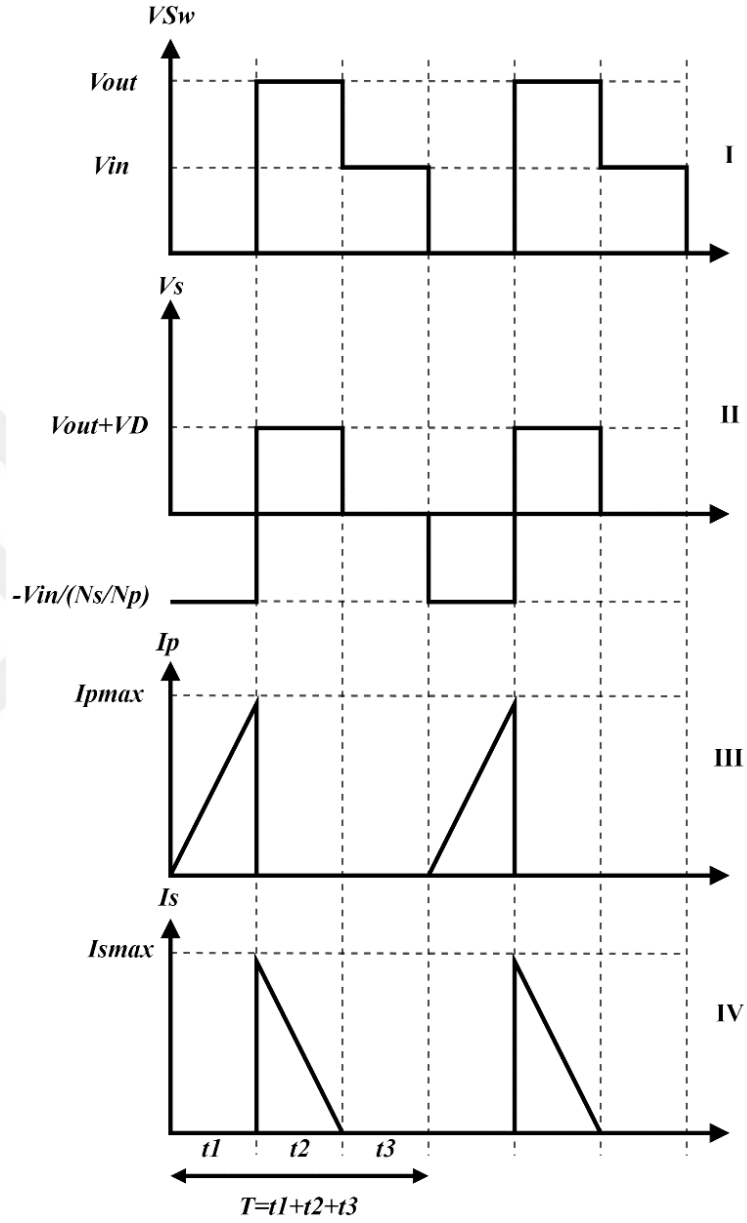


Figure 2. Flyback Converter Waveforms

In Fig. 2, waveform I indicates the voltage level of the switching element, waveform II V_s the secondary voltage, V_D symbol output diode voltage, waveform III

I_p the current in the primary winding of the transformer, I_s the current in the secondary winding of the transformer.

The current in the primary winding of the transformer and the magnetic flux increase when the switching element of the flyback converter switches to the conduction state. Fig. 3 shows the circuit diagram in which the switching element of the flyback converter circuit works in conduction.

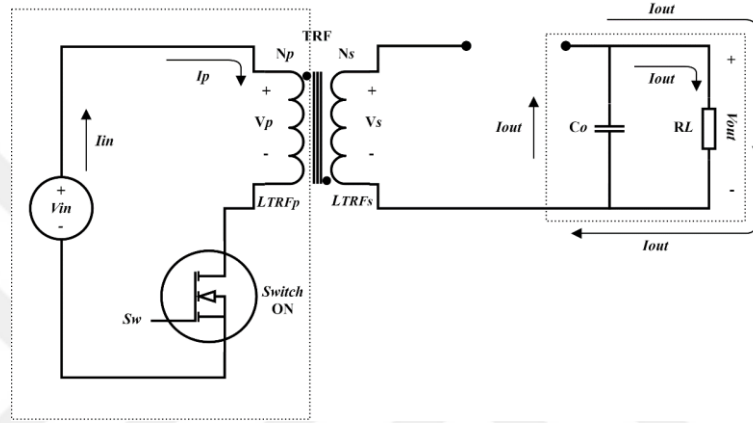


Figure 3. Switching Element Conduction Mode

As the magnetic flux in the primary winding increases, energy is stored in the transformer and the magnetizing inductance of the transformer is used to store the magnetic energy. Depending on the magnetization inductance (L_{TRFp}) in the primary winding of the transformer, the magnetizing current or the current occurring in the primary winding are shown in the Eq. 2.1 and the L_p indicates the magnetising inductance of the secondary winding of the transformer.

$$I_p = \frac{V_{in}}{L_p} (t_1 - 0) \quad 2.1$$

The maximum primary winding current on the transformer when the switching element is open is shown in Eq. 2.2 below and V_s indicates the secondary voltage of the transformer.

$$I_p^{max} = \frac{V_s}{L_p} t_1 \quad 2.2$$

Since the secondary winding of the transformer is in the opposite direction with respect to the primary winding, the diode used in the flyback circuit is reverse polarized and negative voltage accumulates on it. In addition, due to the reverse polarity of the diode, energy transfer is not provided from the primary to the secondary of the transformer of the flyback circuit. The current to flow over the load is provided by the output filter capacitor. (R. Kanthimathi & J. Kamala, 2015)

In order for the transformer core not to reach the saturation stage due to the stored energy, an air gap is left and the energy accumulated in the core is accumulated in the air gap. The primary and secondary windings of the transformer are in different directions. In this case, current passes through the primary winding while no current passes through the secondary winding.

As soon as the switching element turns off, the energy, current and magnetic flux levels in the primary winding decrease to zero. Fig. 4. shows the flyback converter circuit when the switching element is in the off state.

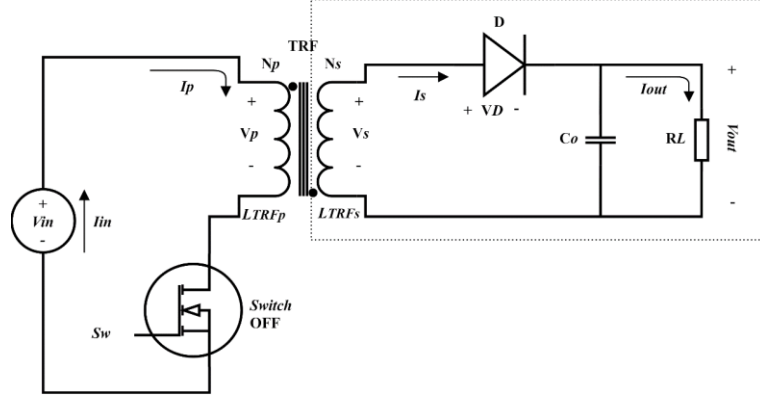


Figure 4. Switching Element Cutoff Mode

The polarity of the transformer changes due to the magnetic flux decrease and the energy in the primary winding is induced into the secondary winding and positive voltage is accumulated. The diode operates in conduction mode when a positive voltage is applied to the secondary winding and the maximum current in the secondary is generated and decreases when the diode is in conduction. The current from the primary part of the transformer is equal to the current supplied from the power source. As stated in Eq. 2.3, when the switching element is in cutoff mode, no current flows from the power supply, that is, from the input part of the circuit.

$$I_{in} = 0 \quad 2.3$$

The maximum current induced in the secondary winding is shown in Eq. 2.4.

$$I_s^{max} = \frac{N_s}{N_p} \frac{V_s}{L_p} t_1 \quad 2.4$$

The current in the secondary winding is shown in Eq. 2.5 while in Eq. 2.5 the symbol L_s indicates the magnetization inductance of the transformer secondary winding.

$$I_s^m = -I_s^{max} \frac{V_{out}}{L_s} (t_1 - 0) \quad 2.5$$

The energy accumulated in the air gap of the transformer is transferred to the secondary part of the circuit. In this case, the output filter capacitor is charged and the load current at the output is obtained. (Kamil, 2007) Capacitance, ESR, ESL and ripple voltage value across the capacitor are important to determine the output filter capacitor. Capacitance, ESR and ESL values of the filter capacitor control the peak-to-peak voltage at the frequency of the switching element.

2.2 Flyback Converter Operation

Switching power supplies are examined, there are continuous, discontinuous and limit modes while The flyback converter also has continuous, discontinuous and boundary modes. (Urgun et al, 2009) Transducer performance and waveforms differ for CCM and DCM modes. The flyback converter has two different operating modes, CCM and DCM, which occur when switching when the current of its transformer is at zero. In addition, the flyback converter has a boundary mode that occurs when the energy stored in the transformer reaches zero when the switching period ends. In other words, it is the boundary state of the switching element when it goes from open to closed, from closed to open. Fig. 5. shows the circuit diagram suitable for the operating modes of the flyback converter with clamp&snubber.

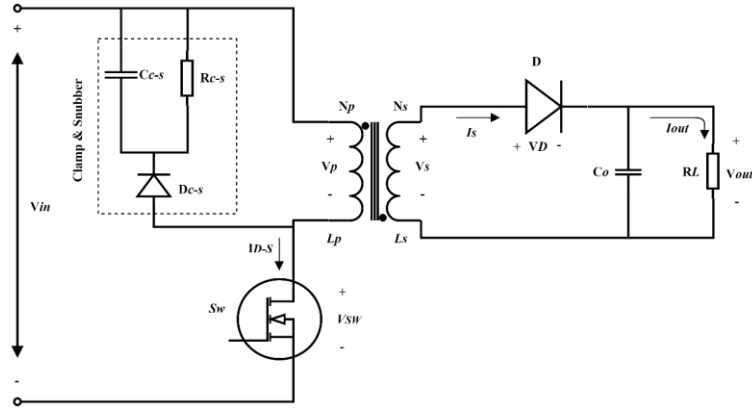


Figure 5. Circuit Diagram of Flyback Converter with Clamp&Snubber

DCM and CCM modes are determined according to the on-off state of the switching element when the currents in the primary and secondary parts of the transformer, which the converter has, fall to zero or change direction without falling to zero. The flyback converter also has Boundary Conduction Mode (BCM) mode different from DCM and CCM mode. The BCM is formed by the flyback converter operating at the boundary between CCM and DCM. The switching element of the converter starts to operate in conduction mode while energy is stored on the transformer as much as the load needs to be used. Then the switching element switches to the cut-off mode. When the load connected to the output side of the converter absorbs all the energy on the transformer, the switch starts to operate in transmission mode again. BCM mode operation is similar to DCM mode. A dead time occurs when the switching element changes from the conduction mode to the cut-off mode. In addition, enables the converter to work more efficiently and it creates a variable switching frequency. BCM mode is similar to CCM mode in AC voltage applications while it can operate more efficiently than CCM mode. In AC voltage applications, core losses play an important role in the design because the growth of the magnetic field increases the losses in the transformer core. (Picard, 2010) In the next section, DCM and CCM modes are explained in detail.

2.2 Flyback Converter Operating Modes

2.2.1 Continuous conduction mode. The CCM mode in the flyback converter occurs when the switching element of the flyback converter turns on again without reducing the current on the secondary side of the transformer to zero. The fact that the secondary current does not decrease to zero indicates that there is some more energy on the secondary side of the transformer. In other words, it is the mode in which the energy stored in the transformer of the flyback converter is in the transformer in the ON state of the MOSFET when it is in the next duty area. In this case, the energy transfer is less in CCM mode. In CCM mode, the secondary current on the transformer does not go down to zero before the duty cycle is over. The minimum primary current of the transformer in CCM mode is shown in Eq. 2.6 and primary waveforms in Fig. 6. In Eq. 2.6, the forward voltage of the V_f output diode defines the f_{sw} switching frequency value.

$$I_p^{min} = \left(\frac{(V_{out} + V_f)I_{out}}{V_{in}f_{sw}t_{ON}} \right) - \left(\frac{1}{2}I_p^{ripple} \right) \quad 2.6$$

Maximum primary current in Eq. 2.7,

$$I_p^{max} = I_p^{min} + I_p^{ripple} \quad 2.7$$

In Eq. 2.8, the symbol I_p^{ripple} indicates the ripple current formed in the primary, while the symbol L_p indicates the magnetization inductance of the primary winding of the transformer.

$$I_p^{ripple} = \frac{1}{L_p} V_{in} t_{ON} \quad 2.8$$

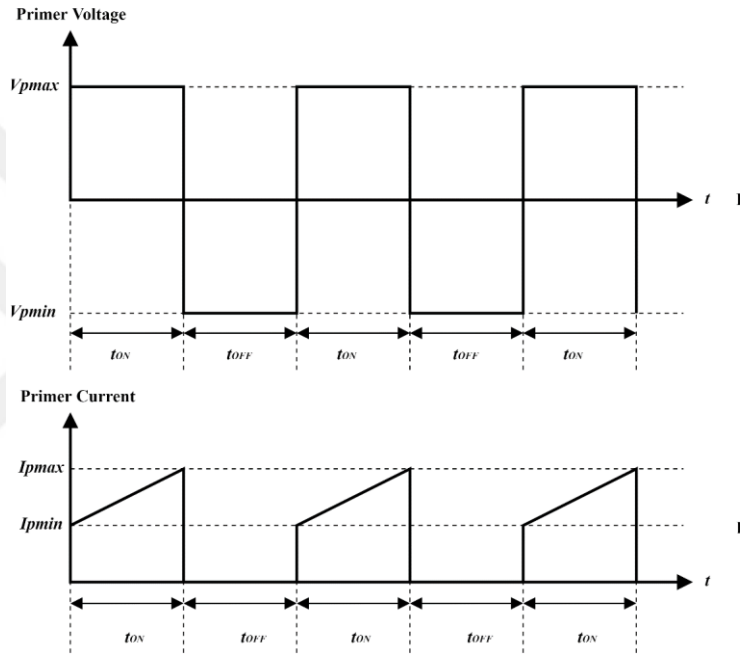


Figure 6. Primary Voltage and Current Waveforms of CCM Mode Transformer

In Fig. 2.6, waveform I indicates the maximum voltage in the primary winding of the transformer V_p^{max} , the minimum primary voltage V_p^{min} , in waveform II, the maximum primary current I_p^{max} and the minimum primary current I_p^{min} of the transformer. In Eq. 2.9, the minimum secondary current in CCM mode is,

$$I_s^{min} = I_p^{min} \frac{N_p}{N_s} \quad 2.9$$

Maximum secondary current in Eq. 2.10,

$$I_s^{max} = (I_p^{max} + I_p^{ripple}) \left(\frac{N_P}{N_S} \right) \quad 2.10$$

While the secondary waveforms are shown in Fig. 7, waveform I represents the maximum voltage in the secondary winding of the transformer V_s^{max} , the minimum primary voltage V_s^{min} , in waveform II, the maximum secondary current I_s^{max} and the minimum secondary current I_s^{min} indicate the transformer.

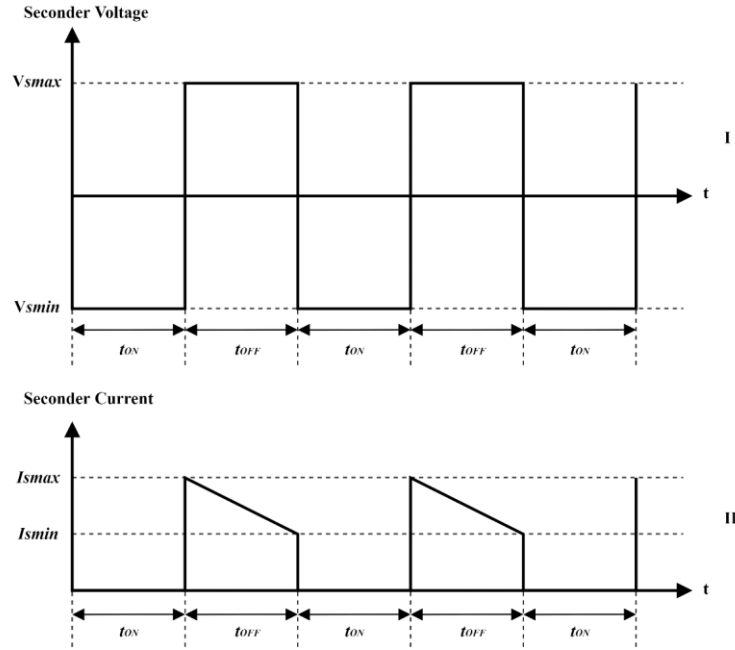


Figure 7. Secondary Voltage and Current Waveforms of CCM Mode Transformer

The current on the output diode is the same as the current on the secondary section of the transformer. The minimum current of the output diode is specified in Eq. 2.11 and its maximum current is specified in Eq. 2.12.

$$I_D^{min} = I_s^{min} \quad 2.11$$

$$I_D^{max} = I_s^{max} \quad 2.12$$

In the flyback converter, the CCM mode provides a high inductive capacity and large volume transformer due to its larger DC flux density. (R. Kanthimathi & J. Kamala, 2015)

Fig. 8 shows the waveforms of the switching element and output diode of the Flyback converter in CCM mode, waveform I shows the PWM signal applied for switching on and off of the switching element, in other words the duty cycle, in waveform II shows the voltage levels and ringing on the switching element, in waveform III shows the current change on the switching element indicated by the symbol ΔI_{sw} , and waveform IV shows the current change on the output diode.

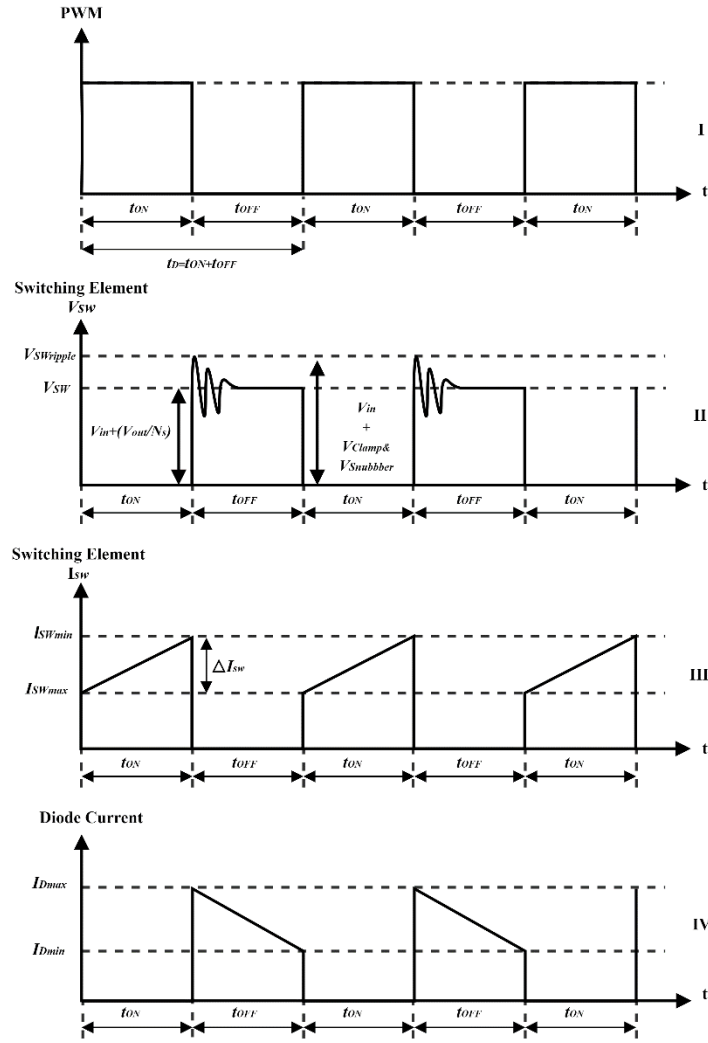


Figure 8. Switching Element and Output Diode Waveforms in CCM Mode

CCM mode is the low magnetization state of the core of the flyback converter transformer. The flux in the transformer core increases linearly from the zero order until the switch is closed during the open mode of the switching element. The flux in the core must be equal to the change in the switch open state time and the change in the switching element closed state time. In other words, the change in core flux during the duty cycle should be equal. When the flux change in the core is equal during the on state (t_{ON} time shown in waveform I in Figure 2.8) and off state (t_{OFF} time shown in waveform I in Fig. 8), the core is prevented from reaching saturation. Switching element open state is given in Eq. 2.13 and closed state is given in Eq. 2.14.

$$t_{ON} = \left(\frac{1}{f_{sw}} \right) (V_{out} + V_f) \left(\frac{\frac{N_P}{N_S}}{V_{in}(V_{out} + V_f) \frac{N_P}{N_S}} \right) \quad 2.13$$

$$t_{OFF} = \left(\frac{1}{f_{sw}} \right) - t_{ON} \quad 2.14$$

The load connected to the flyback converter in practice creates losses in the transformer and output diode of the converter. Ideal situation In CCM mode, the duty cycle of the flyback converter changes to keep the output voltage constant. The input and output voltages generated by the CCM mode flyback converter are shown in Eq. 2.15. The symbol D indicates the duty cycle.

$$\frac{V_{out}}{V_{in}} = \left(\frac{N_S}{N_P} \right) \cdot \left(\frac{D}{1 - D} \right) \quad 2.15$$

During the duty cycle, resonance and voltage stress occur on the energy stored in the transformer due to the switching element capacitance and the transformer leakage inductance. Ringing is reflected from the primary side of the transformer to the secondary winding while to eliminate the ringing, damping is done with snubber and clamp circuits. The maximum voltage stress of the switching element is shown in Eq. 2.16. The symbol V_{cs} denotes the clamp&snubber circuit (ringing) peak voltage, the leakage inductance in the primary winding of the V_p^{leak} transformer.

$$V_{sw}^{ripple} = V_{in} + V_{cs} + V_p^{leak} \quad 2.16$$

In CCM mode, the flyback converter contains lower peak currents on the transformer and a smaller output capacitor than in DCM mode. This makes EMI problems simpler. The equivalent series resistance (ESR) of the output filter capacitor plays an important role in reducing the output voltage ripple of the flyback converter. The voltage ripple variation (peak-to-peak) of the output filter capacitor is equal to the maximum value of the current in the secondary of the output diode and transformer. (Hart, 2010) In CCM mode, flyback converter output voltage fluctuation is in Eq. 2.17,

$$\frac{\Delta V_{out}}{V_{out}} = \frac{D}{R_L C_o f_{sw}} \quad 2.17$$

The voltage variation formula of the output filter capacitor equivalent resistance is given in Eq. 2.18.

$$\Delta V_{C_o}^{ESR} = (I_p^{max} + I_p^{ripple}) \left(\frac{N_p}{N_s} \right) R_C^{ESR} \quad 2.18$$

In CCM mode, the transformer consumes more energy than the DCM mode, as it has lower peak currents on it and reduces the transmission loss of the switching element.

The flyback converter with CCM mode simplifies the design since the input and output voltage gain are determined according to the duty cycle while convenient to use in designs with low voltage and high current requirements.

2.2.2 Discontinuous conduction mode. DCM mode in flyback converter occurs when the switching element of the flyback converter turns on again when the current on the secondary side of the transformer drops to zero. The current increases on the primary side of the transformer. When the switching element of the converter enters cutoff mode, the current in the secondary section of the transformer drops to zero when the switch turns on. The decrease of the secondary current to zero indicates that the energy level of the primary section of the transformer has decreased to zero and the secondary section has reached the maximum energy level. This indicates that there is maximum energy transfer on the transformer of the flyback converter in DCM mode. Since the diode in the output section of the converter is in the off state, the output current is supplied from the filter capacitor while the output diode in the cut-off state, the output section of the converter becomes isolated from the input section. In DCM mode, the secondary current on the transformer goes down to zero before the duty cycle ends. The primary voltage and current waveforms of the transformer are shown in Fig. 9 and the minimum primary current in Eq. 2.19.

$$I_p^{min} = 0 \quad 2.19$$

The maximum primary current of the transformer is shown in Eq. 2.20.

$$I_p^{max} = I_p^{ripple} = \frac{1}{L_p} V_{in} t_{ON} \quad 2.20$$

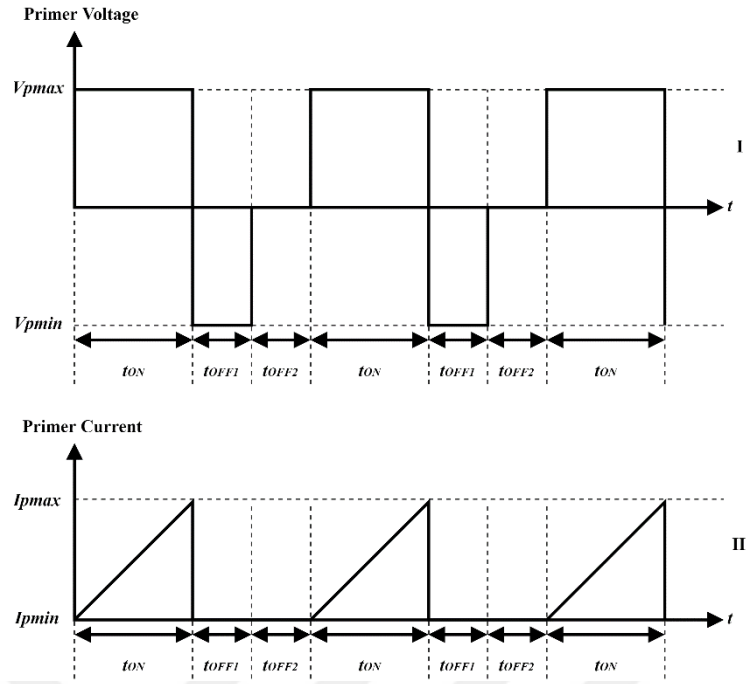


Figure 9. Primary Voltage and Current Waveforms of DCM Mode Transformer

The secondary voltage and current waveforms of the transformer are shown in Fig. 10 and the minimum secondary current in Eq. 2.21.

$$I_s^{min} = 0 \quad 2.21$$

The maximum secondary current of the transformer is shown in Eq.2.22.

$$I_s^{max} = \frac{1}{L_p} V_{in} t_{ON} \left(\frac{N_p}{N_s} \right) \quad 2.22$$

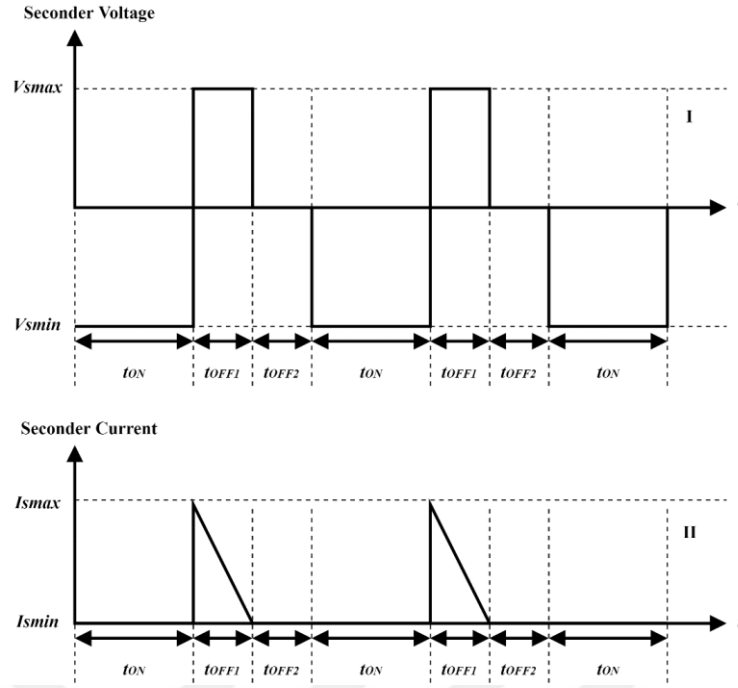


Figure 10. Primary Voltage and Current Waveforms of DCM Mode Transformer

As shown in Eq. 2.23, when the components used in the converter are considered ideal, the power supplied from the power supply is the same as the power drawn from the load of the circuit.

$$P_{in} = P_{out} \quad 2.23$$

In practice, in the DCM mode shown in Eq. 2.24, the value of the input power of the converter is found by dividing the output power value by the efficiency of the converter. In Eq. 2.24, the symbol η indicates the efficiency of the converter.

$$P_{in} = \frac{P_{out}}{\eta} \quad 2.24$$

In Fig. 11, the PWM signal applied for the opening and closing of waveform I switching element, in other words, the duty cycle, voltage levels and ringing on the switching element in waveform II, current levels on the switching element in waveform III, the current levels occurring on the output diode are indicated in waveform IV. In the switching element conduction mode shown in Fig. 11, the inductance value and the on-time of the switching element increase up to the value determined during t_{ON} time. During the switching element conduction time, the output current is provided through the output capacitor, since the output diode behaves in the cut-off state. The switching element cut-off status is indicated by t_{OFF} . When the switching element is in the cut-off state, the energy on the transformer creates a negative voltage on the secondary winding. This situation fills the output filter capacitor by sending the current on the output diode to the load part. Since the current on the output diode decreases during t_{OFF1} , the energy stored during the tone period is exhausted. When the stored energy is exhausted, a ringing occurs at t_{OFF2} time and no current flows during t_{OFF2} time. If the elements contained in the transducer are considered ideal, ringing does not occur, but in a practical situation, ringing does occur. Resonance occurs due to the capacitances of the switching element and the output diode, the magnetizing inductance of the transformer.

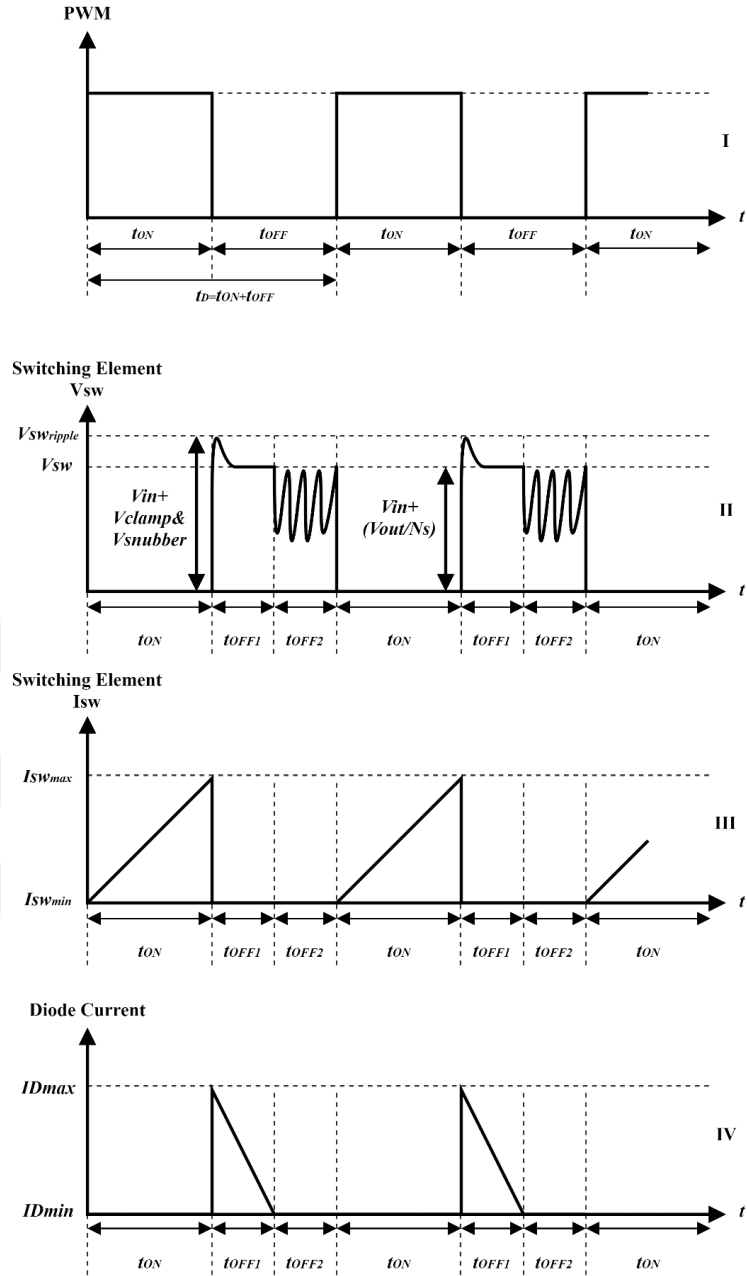


Figure 11 Switching Element and Output Diode Waveforms in DCM mode

The open t_{ON} state of the switching element in Eq. 2.15, in Eq. 2.16, the time of the switching element, number one closed state t_{OFF1} and number two closed state t_{OFF2} , are also indicated in Eq. 2.17.

$$t_{ON} = \sqrt{2I_{out}L_p \left(\frac{V_{out} + V_f}{f_{sw}V_{in}^2} \right)} \quad 2.15$$

$$t_{OFF1} = \left(t_{ON} \frac{V_{in} + (V_{out} + V_f) \frac{N_p}{N_s}}{(V_{out} + V_f) \frac{N_p}{N_s}} \right) - t_{ON} \quad 2.16$$

$$t_{OFF2} = \frac{1}{f_{sw}} - t_1 - t_2 \quad 2.17$$

The voltage formed on the switching element during the t_{OFF2} process is specified in Eq. 2.18.

$$V_{sw} = V_{in} \quad 2.18$$

The current generated on the output diode is the same as the current in the secondary section of the transformer as in the CCM mode. The minimum current of the output diode is specified in Eq. 2.11, and its maximum current is specified in Eq. 2.12. The voltage formed on the output diode at the time t_{OFF2} is given in Eq. 2.19.

$$V_D^{t_{OFF2}} = -V_{out} \quad 2.19$$

The current formed on the output filter capacitor at the time t_{ON} is specified in Eq. 2.20.

$$I_{C_o} = -I_{out} \quad 2.20$$

The minimum current formed on the output filter capacitor during t_{OFF1} and t_{OFF2} times is specified in Eq. 2.21, and maximum current is specified in Eq. 2.22.

$$I_{C_o}^{min} = I_{smin} - I_{out} \quad 2.21$$

$$I_{C_o}^{max} = I_s^{max} - I_{out} \quad 2.22$$

DCM mode in flyback converter has low efficiency and high leakage inductance of transformer. The leakage inductance of the transformer and the current in its primary winding create instantaneous resonance and voltage stress while this causes EMI problems and high cost of the switching element. (Urgun et al, 2009) Transducer efficiency can be increased when the ringing and voltage stress in the switching element is reduced. Snubber&clamp applications are used to reduce ringing and tension stress while rectifier diodes can be used to reduce switching element problems. (Choi, 2004) DCM mode is suitable for low current and high voltage applications. In DCM mode, low inductance transformer is required compared to CCM mode, due to the high current ripple. This reduces the size and cost of the transformer. (Picard, 2010) In CCM mode, the input and output voltage gain is determined by the duty cycle, while in DCM mode, the input and output voltage gain is determined by the duty cycle and load condition. (Choi, 2004)

2.3 Snubber Cell & Clamp

In switching power applications, the transformer or coil connected to the switching element generates ringing due to leakage inductances in the on and off transition states of the switching element. It causes switching losses due to current and voltage blockage in the on and off transition state (Pressman et al, 2009). As the switching element transitions from the on state to the off state, the current changes rapidly, utilising the leakage inductance of the transformer or coil, causing overvoltage amplification and voltage stress. When high current and voltage occur during the transition of the switching element from the off state to the on state, it takes the switching element out of the safe operating region (Trzynadlowski, 2015). This situation overstresses the circuit elements and causes the converter to fail or shorten its life. In addition, the resulting voltage stress is transmitted to the converter system as a noise problem. Ringing-induced, radiated, transmitted noise, voltage stress, EMI problems and switching losses. The switching element voltage stress of the converter has a high peak value. Clamp designs are recommended for the voltage stress on the switching element of the converter and snubber designs are recommended for ringing. (Ridley, 2005) The purpose of clamp and snubber designs is to reduce switching losses by reducing the ringing and voltage stress on the switching element, to prevent overvoltage and overcurrent spikes and to ensure linear and safe operation of the switched power system.

Snubber designs raise key criteria such as efficiency, reliability and performance of a switching power converter, while lowering criteria such as size and weight. This makes the converter more useful and convenient. In addition, switching converters with high switching frequency create higher ringing and EMI problems, which can be easily solved with snubber designs. (Todd, 1993) This makes the converter more useful and convenient. In addition, switching converters with high switching frequency create higher ringing and EMI problems, which can be easily solved with snubber designs. (Todd, 1993) Snubber designs help to distribute voltage between series connected switching elements and to protect currents between parallel connected switching elements. Snubber designs and structures vary according to

switching power supply topologies. (Trzynadlowski, 2015) The snubber circuit dampens the reactances created by leakage inductances and capacitances. RC snubber is used to minimise the power dissipation on the switching element. The resistor connected in parallel to the switching element reduces the voltage rise caused by the ringing and absorbs the energy. When a capacitor is connected in series, it dissipates the energy on the system while switching loss is reduced. When the resistor and capacitor values are selected appropriately in the design, the losses from the switching element are reduced by 40% (Todd, 1993).

Ringing occurs in flyback converters due to capacitance and inductance. In the flyback converter, the printed circuit board designed for the converter, the converter elements create capacitance, while the transformer creates leakage inductance. Ringing is seen on the diode and switching element. (Betten, 2016) In flyback converters, when the switching element is switched off, the leakage inductance of the transformer stops the current and the voltage on the switching element rises excessively while the capacitance formed in other components with the voltage rise creates ringing in the primary winding of the flyback transformer. The peak voltage value of the ringing is given in Eq. 2.23 while in Eq. 2.23, V_s^{ripple} is the peak value of the resonance, I_t^{OFF} is the current value of the switching element in the off state, L_p is the primary winding inductance of the transformer, C_{Lp} is the inductance in the primary winding of the transformer.

$$V_s^{ripple} = I_t^{OFF} \left(\sqrt{\left(\frac{L_p}{C_{Lp} + C_o} \right)} \right) + V_{in} + \left(\frac{V_{out}}{N_p} \right) \quad 2.23$$

Considering the cost issue, RC snubber design is used to damp the ringing. (Sonst, 2021) RC snubber does not need additional control and is called energy consuming snubber. Fig. 12 shows the RC snubber circuit of the Flyback converter topology.

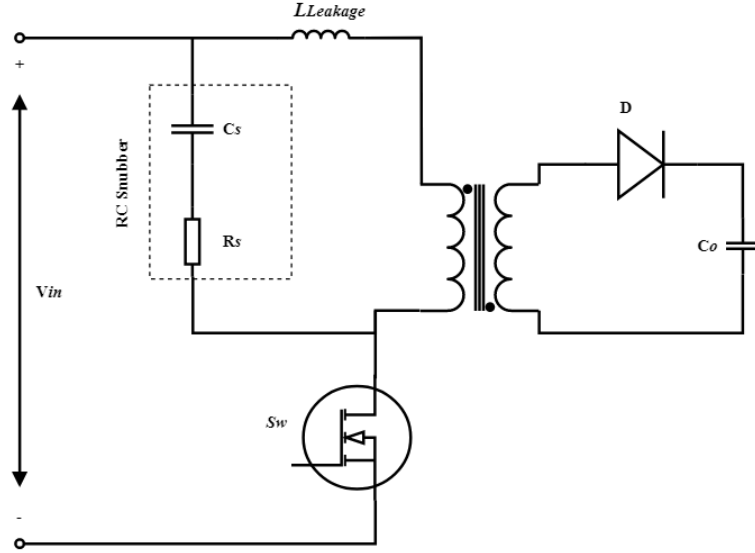


Figure 12. RC Snubber Circuit of the Flyback Topology

In the snubber circuit, a resistor is used to damp the resonance formed by the leakage inductances of the transformer, and a capacitor connected in series with the resistor is used to damp the voltages generated by the frequency generated by the switching element. (Ridley, 2005) If the characteristic impedance value of the resonance formed by the leakage inductances is equal to the snubber resistance, the ringing is significantly damped. In addition, the frequency of the resonance helps the damping significantly. The frequency of the resonance should be approximately twice the switching frequency (Ridley, 2005). The characteristic resonance value is given in Eq. 2.24. In Eq. 2.24, the symbol Z is the characteristic value of the resonance, f_s is the frequency of the resonance and L_p^{leak} is the leakage inductance of the transformer and Z_p^{leak} is the primary leakage inductance impedance of the transformer.

$$[Z_p^{leak} = 2\pi f_r L_p^{leak}] \quad 2.24$$

Since the inductance value of the resonance is equal to the leakage inductance in the primary section of the transformer, the snubber resistance can be easily estimated. When the value of the capacitor connected in series is chosen equal to the impedance of the ringing frequency, the ringing is stopped. The value of the capacitor is given in Eq. 2.25 while in the Eq. 2.25, the symbol C_s indicates the snubber capacitor value and the symbol R_s indicates the snubber resistance value.

$$C_s = \frac{1}{2\pi f_s R_s} \quad 2.25$$

In practice, it is observed experimentally by calculating the capacitor value using resistance estimation and the ringing can be optimised by increasing the capacitor value. The power dissipation on the snubber circuit is determined by the capacitor connected in series with the snubber resistance. Snubber power dissipation is given in Eq. 2.26 while in Eq. 2.26, the symbol P_s indicates the power dissipation value and V_s indicates the difference between the peak ringing voltage and the input voltage.

$$P_s = C_s (V_{in} + V_s)^2 f_{sw} \quad 2.26$$

Clamp design is used when the snubber is inadequate. Clamp is designed to reduce the ringing and power losses of the switching elements as in the snubber. The clamp design protects the power converter system by not allowing the voltage jump generated by the switching element to exceed the maximum voltage value of the switching element. EMI problems are solved more easily by controlling the peak voltage, current and power on the switching element. Generally, resistor, capacitor and diode components are used. Fig. 13 shows the RCD clamp circuit of the flyback converter topology.

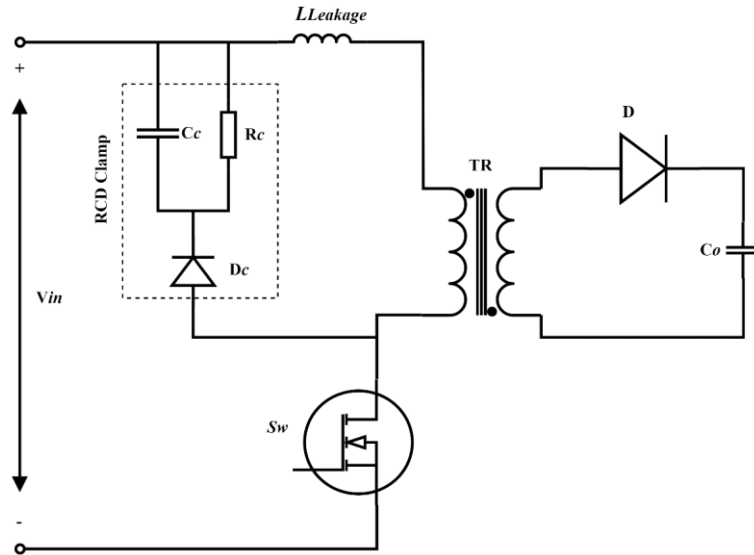


Figure 13. Clamp Circuit of the Flyback Converter

2.4 Flyback Converter Transformer

In SMPS converters, magnetic elements such as inductors and transformers are used according to topologies. Transformers are used for energy storage and energy transmission. In flyback converters, magnetic element transformer is used and transformer design is among the most important elements for the converter. Flyback transformer stores and transmits energy while isolating the input and output. Flyback transformer works like an inductor with two or more than two windings while its behaviour of storing and transmitting energy shows that it is different from a normal transformer. Due to the energy storage in the transformer, there is no need for an output filter compared to other isolated SMPS topologies, which reduces the number of components used in the converter and reduces the cost value. (Adams, 2001) For transformer design, it is necessary to determine the input voltage, output voltage, maximum power rating, duty cycle and operating frequency parameters. According to the determined parameters, the number of transformer windings, size and core selection can be made easily. Fig. 14 shows the flyback transformer structure.

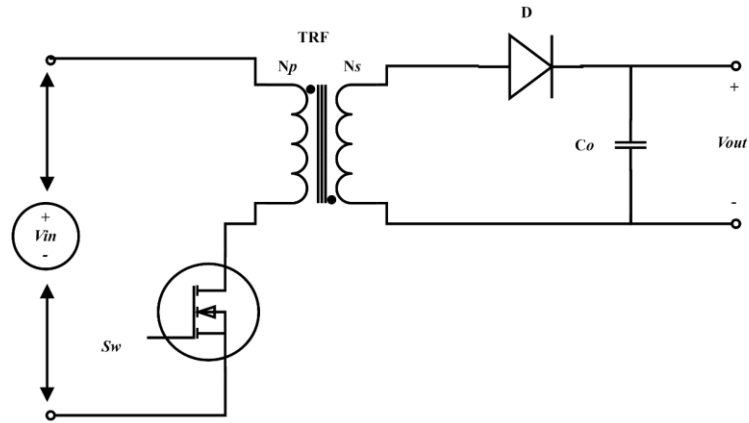


Figure 14. Flyback Transformer Structure

The primary winding of the transformer is magnetised when the switching element switches to the conduction state and stores energy in the air gap. When the switching element switches to the cut-off state, the energy stored in the air gap passes to the secondary winding. The secondary winding stops the magnetisation of the core and transmits the energy accumulated in the air gap to the load part. The secondary winding can provide isolation while it increases or decreases the input voltage according to the winding number ratio. When the number of secondary windings increases, more output voltage can be obtained with a single input voltage. No current flows in the primary and secondary windings of the transformer at the same time (Picard, 2010). (Picard, 2010) Fig. 15 shows the transformer in the switch conduction state and Fig. 16 shows the transformer in the switch off state.

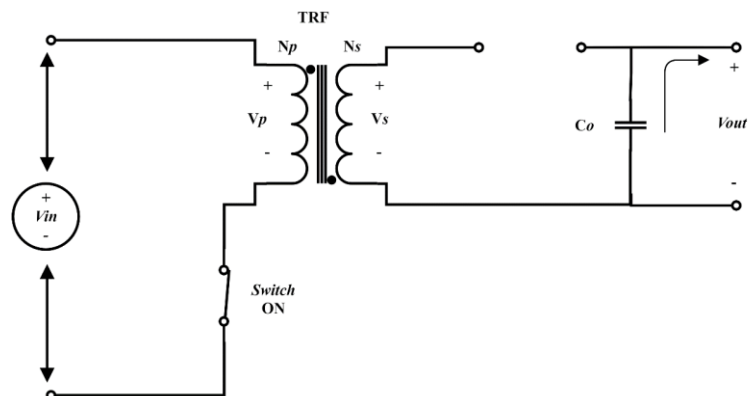


Figure 15. Transformer Current Flow with Switch On

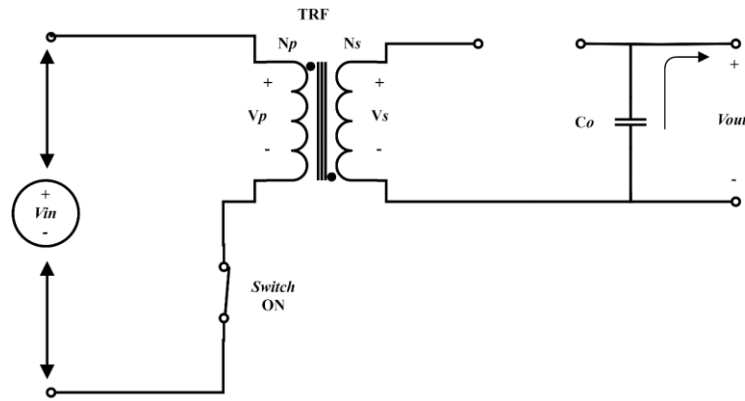


Figure 16. Transformer Current Flow with Switch Closed

In order for the flyback transformer to operate at high switching frequency, a core material with low losses should be used to store the energy of the transformer and to prevent saturation of the core. The core should be selected according to system parameters such as size, power rating, output voltage rating and cost. It is more appropriate to use large cores in low voltage value and multiple output voltage applications. There are different core materials used today. The core type commonly used in SMPS applications, including flyback, is ferrite. Choi, (2012) Ferrite has iron oxide compounds in its structure while it is baked at high temperature. Ferrite cores are used in many inductor designs while they are also preferred in transformers. In the physical design of inductors and transformers, ferrite is formed by wrapping wire around it. Ferrite cores have high permeability, high flux density, low core loss and limited shapes (Coilcraft, 2022).

There are E, EFD, EPC, EE, EF, ETD, EET and PQ core types. EFD and EPC cores are used due to their small size, EE and EF cores are used for vertical or horizontal use, ETD and EER cores are used if a larger area is required, and PQ type cores are used for lower winding number, winding surface area ratio and good EMI return. PQ type cores require less number of windings than EE type cores. PQ type cores have proportional round feet, large outer surface and better EMI spectrum than

E type cores. In addition, due to its round surface, it reduces the copper cost for production and provides heat dissipation for design. (TDK, 2013)

One of the most important criteria in ferrite selection is magnetic permeability, magnetic flux density and magnetic field strength. Magnetic permeability of ferrite cores is high and changes with temperature while the change in magnetic permeability with temperature leads to power loss. Magnetic permeability and magnetic flux density are analysed by examining the B-H graph in the data sheets of ferrites. Fig. 17 shows an ideal B-H graph, known as the magnetisation curve. Tesla (T) stands for induction in the core, B stands for magnetic flux density and H stands for magnetic field strength.

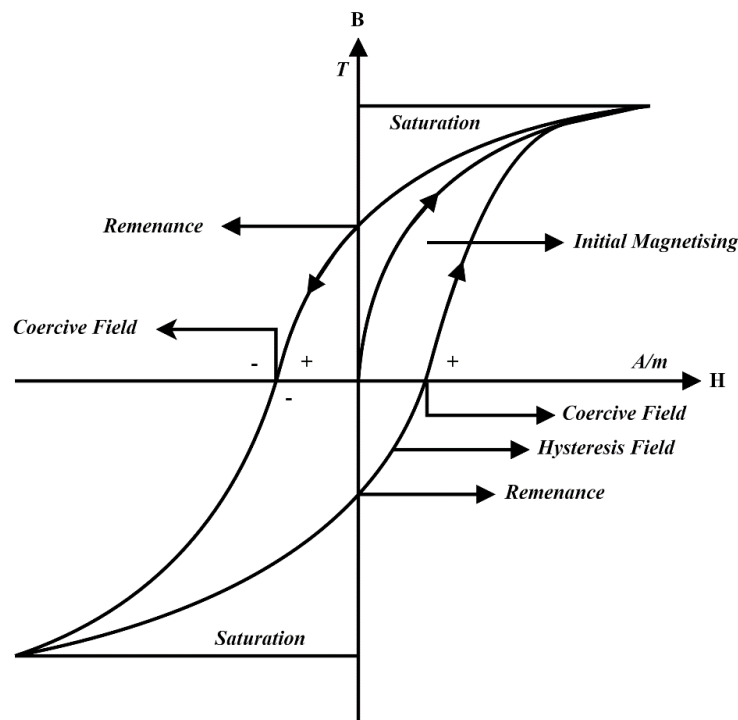


Figure 17. Typical Magnetization Graph

The curve indicated as the initial magnetising shows the initial magnetisation of a core that has never been magnetised. At small currents, the slope of the curve is low and the core increases the magnetisation at very small rates, which is referred to as the initial permeability. As the current is increased, the curve steepens and remains close to the line in a fairly large region in this region the permeability is high. At large

field intensities (currents), the slope of the curve starts to decrease again and eventually it is referred to as saturation, i.e. saturation. In the saturation region, the transformer now operates like a simple air coil without a core. If the current passing through a magnetised coil is reduced until saturation, the magnetisation in the core does not decrease by following the initial change curve backwards, it remains slightly above this curve and the magnetisation is not zero when the current is completely cut off. The remaining magnetisation is called remanence. In order to eliminate the remanence, it is necessary to flow some current in the opposite direction, which is called coercive current. If the current is continued to increase in the reverse direction, the saturation region in the negative direction is reached. By decreasing this current in the negative direction, the remanence and coercive field regions in this direction are passed with a change similar to that in the positive direction, and by increasing the current in the positive direction again, the saturation region on the positive side is reached again. As can be seen, the magnetisation characteristic is not a curved line, but a closed shape and this is called hysteresis curve. As a result of the B-H analysis, boundary values and behaviours are found for the core not to saturate. Eq. 2.27 specifies the magnetic field strength and Eq. 2.28 specifies the magnetic flux density.

$$H = \frac{I \cdot N}{l} \frac{A}{m} \quad 2.27$$

$$B = \frac{\Phi}{A} \frac{V_s}{m^2} \quad 2.28$$

The other important issues in the flyback transformer are the air gap, winding wires and core losses. In the flyback transformer, energy is stored in the air gap as mentioned before. The reason why the energy is stored in the air gap is that the core has a high permeability and cannot store excess energy without reaching saturation. When the air gap is added to the transformer, the hysteresis curve of the core is skewed and the magnetic field intensity is increased to take the core to the saturation region.

(Dinwoodie, 1999) In each period, the hysteresis curve will create energy loss by rotating once. In order to reduce the energy loss in the design, the hysteresis field is selected as small as possible. Cores are not placed in one piece for the windings. The core, which becomes conductive when placed in one piece, causes the primary winding to behave like a short circuit to the secondary winding. In this case, the primary winding flows current called Foucault currents and causes a large power loss. Hysteresis and Foucault losses are called iron losses and are considered as a resistance parallel to the primary. The resistances of transformer windings cause power and voltage losses. The losses that the winding currents spend on the winding wires are called iron losses. In order to reduce copper losses, it is reduced by using large cross-sections of winding wires. In converter design, parallel windings with more than one wire should be used for high efficiency and to reduce the skin effect.

Chapter 3

Proposed Design Details

3.1 Introduction

In this section, the parameters of the flyback converter are given and detailed calculations are made. Calculations related to the transformer and EMI filter designs are given and information about the integrated used for the converter is given. Together with the calculations and approximate estimates, the converter circuit diagram is created and the converter circuit is simulated with LTspice programme and the results are given.

3.2 Flyback Converter Design Calculation

The requirements of the flyback converter circuit designed within the scope of the thesis are given in Table 1. In the circuit designed for the converter, the LT8316 integrated produced by Analogue Devices was used due to the fact that it does not need a feedback circuit, does not need an optocoupler circuit for regulation and generally reduces the cost. MOSFET was chosen as the switching element due to its high switching frequency, high input voltage value, thermal performance and cost. FCB125N65S3 MOSFET manufactured by ON Semiconductor was selected in line with the requirements considered in the design. The transformer core CF139PQ2620 manufactured by Cosmo Ferrites was selected due to its small size, cost, ease of design and EMI response. The output diode is MBRD10100CT manufactured by ON Semiconductor due to cost and thermal performance. The design calculations for the flyback converter are shown below with formulas.

Table 1

Flyback Converter Requirements

Flyback Converter Requirements		
Parameter	Symbol	Value
Max. Input Voltage	V_{in}^{max}	137.5VDC
Nominal Input Voltage	V_{in}^{nom}	110VDC
Min. Input Voltage	V_{in}^{min}	77VDC
Output Current	I_{out}	1.5A
Output Voltage Ripple	V_{out}^{ripple}	$V_{out}^{ripple} < 100mV$
Output Power	P_{out}	36W
Switching Frequency	f_{sw}	45kHz

3.2.1 Design of flyback transformer. The magnetisation inductance suitable for the requirements of the flyback converter transformer converter is calculated by determining the minimum opening, closing time and output power of the converter specified in the datasheet page of the LT8316 integration. The current generated in the secondary winding of the transformer transfers the output voltage value to the tertiary winding of the transformer with a feedback pulse while LT8316 needs 800ns to adjust and sample the output voltage transferred with a margin of error. In other words, the secondary winding of the transformer transfers the voltage value to the tertiary winding of the transformer for a minimum of 800ns in order to generate the sampling value. In addition, it is necessary to determine the duty cycle (D) value in order to determine the magnetizing inductances of the transformer, the current values of the primary and secondary, and the SENSE(R_{sns}) resistance value of the switching element, which is known as the current limiting resistance. Eq. 3.1 specifies the duty cycle value while Eq. 3.2 calculates the duty cycle. The symbol N_{PS} represents the primary winding coefficient and its value is taken as 4.

$$D = \frac{(V_{out} + V_f)N_{PS}}{((V_{out} + V_f)N_{PS}) + V_{in}^{min}} \quad 3.1$$

$$D = \frac{(24 + 0.8) * 4}{((24 + 0.8) * 4) + 77} = 0,56 \quad 3.2$$

Eq. 3.3 specifies the resistance value R_{sns} . Eq. 3.4 calculates the resistance value R_{sns} while 80% is the approximate flyback converter efficiency. The LT8316 integration limits the current at maximum by creating a voltage of maximum 100mV and minimum 50mV on the SENSE pin to set the maximum current of the converter. In the worst case condition, that is, at the minimum input voltage, the current limitation is determined in the duty cycle.

$$R_{sns} = \frac{1 - D}{I_o} 50mV N_{PS} 80\% \quad 3.3$$

$$R_{sns} = \frac{1 - 0,56}{1.5} * 0.05 * 4 * 0,8 = 0,046 \cong 50m\Omega \quad 3.4$$

The approximate value of the minimum primary magnetisation inductance of the transformer is given in equation 3.5 while calculated in equation 3.6. In the 3.5 equation, L_p primary inductance value, V_{out} output voltage, V_f output diode forward voltage value, I_{sw}^{min} minimum switching current limit and t_{off}^{min} minimum switching off time value. In the datasheet of the LT8316 integration, t_{off}^{min} value is 800ns, I_{sw}^{min} value is $20mV/R_{sns}$. As stated in the datasheet of the LT8356 integration, the maximum input voltage is 600VDC. The flyback converter designed in this thesis has a maximum input voltage of 137.5VDC. In order for the converter to have a wider input voltage range and ease of production according to the requirements, the PCB designed is

designed to operate up to 600VDC above 137.5VDC by changing only the side components on it. In this case, in order for the primary winding of the transformer to withstand the maximum input voltage, the maximum voltage value on the transformer was set at 600VDC.

$$L_p \geq \frac{t_{off}^{min} N_{PS}(V_{out} + V_f)}{I_{sw}^{min}} \quad 3.5$$

$$L_p \geq \frac{800 * 10^{-9} * 4 * (24 + 0.8)}{0.4} = 198.4\mu H \cong 200\mu H \quad 3.6$$

As stated in the datasheet of the LT8316 integration, the current rises when the MOSFET is first switched on and The increase in current causes the switching element to operate unstable. In order to reduce the current rise, the integrated MOSFET does not switch the MOSFET into conduction in less than 300ns. Considering this situation, the value of the primary inductance is given in the Eq. 3.7 while calculated in Eq. 3.8. In the Eq. 3.7, t_{th} value is taken as 300ns.

$$L_p \geq \frac{t_{th} V_{in}^{max}}{I_{sw}^{min}} \quad 3.7$$

$$L_p \geq \frac{300 * 10^{-9} * 600}{0.4} = 450\mu H \quad 3.8$$

LT8316 integration is operating at maximum frequency, when the magnetisation inductance of the transformer must be the value that will provide power to the secondary of the transformer. For this reason, the magnetisation inductance is required to be frequency dependent. Eq. 3.9 specifies the frequency dependent primary

magnetisation inductance while calculated in Eq. 3.10. In Eq. 3.9, I_{sw}^{max} is the maximum switching current limit, f_{sw}^{max} is the maximum switching frequency of the LT8316 integrated and η is the efficiency. I_{sw}^{max} is set at $100mV/R_{sns}$, η is approximately 80% and f_{sw}^{max} is set at 140 kHz as stated in the datasheet of the LT8316 integration.

$$L_p \geq \frac{2(V_{out} + V_f)I_o}{\eta(I_{sw}^{max})^2 f_{sw}^{max}} \quad 3.9$$

$$L_p \geq \frac{2 * (24 + 0.8) * 1.5}{0.8 * 2^2 * 140 * 10^3} = 166\mu H \cong 170\mu H \quad 3.10$$

The maximum value of the magnetisation inductance depends on the high speed opening and closing of the output diode. Activates the 50us timer to bring the diode into conduction when the LT8316 integration cannot detect the output diode closing. In this case, the primary magnetisation inductance should not be large enough to exceed the 50us time required for the output diode to enter conduction and the primary magnetisation inductance needs to be dependent on the backup time(t_{bt}). The maximum value of the magnetisation inductance in the primary winding is given in Eq. 3.11. while in Eq. 3.11, t_{bt} is taken as 50 μs and calculated in Eq. 3.12.

$$L_p^{max} < \frac{0.8(V_{out} + V_f)N_{PS}t_{bt}}{I_{sw}^{max}} \quad 3.11$$

$$L_p^{max} < \frac{0.8 * (24 + 0.8) * 4 * 50 * 10^{-6}}{2} = 1.98mH \cong 2mH \quad 3.12$$

The values calculated in equations 3.6, 3.8 and 3.10 are the limit values for the primary magnetisation inductance of the transformer. In practice, the magnetisation inductance is determined by calculating approximately 20%-50% more than the calculated limit values for high voltage, current, thermal performance and not saturating the core. Calculate the nominal primary magnetisation inductance, the highest limit value calculated in equation 3.8 is $450\mu H$. In this case, the nominal primary magnetisation inductance value is stated in equation 3.13 using the result of equation 3.8 and calculated in Eq. 3.14. The maximum primary magnetisation inductance value is specified in Eq. 3.15 using the result of equation 3.12 and is calculated in Eq. 3.16.

$$L_p = L_p + (L_p * 50\%) \quad 3.13$$

$$L_p = 450 * 10^{-6} + (450 * 10^{-6} * 50\%) = 675\mu H \cong 670\mu H \quad 3.14$$

$$L_p^{max} = L_p^{max} + (L_p^{max} * 50\%) \quad 3.15$$

$$L_p^{max} = 2 * 10^{-3} + (2 * 10^{-3} * 50\%) = 3mH \quad 3.16$$

The secondary magnetisation inductance value depends on the primary magnetisation inductance value. The secondary magnetisation inductance value is stated in Eq. 3.17 while is calculated in equation 3.18.

$$L_p = L_s(N_{PS})^2 \quad 3.17$$

$$L_s = 41.8\mu H \cong 40\mu H \quad 3.18$$

The air gap of the transformer is determined by the nominal inductance (A_L) value of the core, the effective area of the core, the primary magnetisation inductance and the primary winding coefficient. The air gap of the transformer is given in Eq. 3.19 while calculated in Eq 3.20. In Eq. 3.19, L_{ag} is the air gap value, μ_0 is the magnetic permeability coefficient of the core, A_e is the effective area and A_L is the nominal inductance value of the core. The A_L value in the datasheet of the CF139PQ2620 core is 4500nH, the A_e value is 121mm² but is determined in m² and μ_0 is taken as $4\pi 10^{-7}$.

$$L_{ag} = \mu_0 A_e \left(\frac{N_{PS}^2}{L_p} - \frac{1}{A_L} \right) 10^3 \quad 3.19$$

$$L_{ag} = 4 * \pi * 10^{-7} * 121 * 10^{-6} * \left(\frac{4^2}{670 * 10^{-6}} - \frac{1}{4500 * 10^{-9}} \right) * 10^3 = 0.03mm \quad 3.20$$

The air gap value was determined as 0.2mm when the data sheet of the core was analysed and the new A_L value corresponding to the 0.2mm air gap specified in the data sheet was determined as 630nH.

The tertiary winding is fed from the Bias pin of the LT8316 integration. As stated in the data sheet of the LT8316 integration, the tertiary winding coefficient should be between 10VDC-30VDC values. The tertiary winding ratio is specified in equation 3.21 while calculated in Eq. 3.22.

$$\frac{10VDC}{V_{out}} < N_{TS} < \frac{30VDC}{V_{out}} \quad 3.21$$

$$0.41 < N_{TS} < 1.25 \quad 3.22$$

The tertiary winding N_{TS} ratio is determined as 0.5.

The number of primary and secondary windings is related to the coefficient specified in the datasheet of the LT8316 integration and the A_L value of the CF139PQ2620 core. In the datasheet of the core, the A_L value depending on the air gap calculation is 630nH. The number of turns depending on the A_L value of the core is specified in Eq. 3.23, the number of primary turns in Eq. 3.24 and the number of secondary turns in Eq. 3.25.

$$N = \sqrt{\frac{L}{A_L}} \quad 3.23$$

$$N_P = \sqrt{\frac{670 * 10^{-6}}{630 * 10^{-9}}} = 32.61 \cong 32 \quad 3.24$$

$$N_S = \sqrt{\frac{40 * 10^{-6}}{630 * 10^{-9}}} = 7.96 \cong 8 \quad 3.25$$

From the values found in the equations, the ratio of primary and secondary winding number is 4:1. In the numbered equation, the tertiary winding N_{TS} winding number coefficient is determined as 0.5, while the primary, secondary and tertiary winding number ratios are 4:1:0.5 respectively. Since the winding number ratios are determined, the number of primary windings is 8 times the number of tertiary windings and the number of secondary windings is 2 times the number of tertiary windings. In this case, the number of tertiary windings is stated in Eq. 3.26 while calculated in Eq. 3.27.

$$N_T = \frac{N_P}{8} = \frac{N_S}{2} \quad 3.26$$

$$N_T = \frac{33}{8} = 4.125 \cong 4 \quad 3.27$$

Since the tertiary winding winding number is 4, the magnetisation inductance is given in Eq. 3.28 using Eq. 3.23 and in Eq. 3.28, L_t is the tertiary winding magnetisation inductance.

$$4 = \sqrt{\frac{L_t}{630 * 10^{-9}}} = 10.08\mu H \cong 10\mu H, L_t = 10\mu H \quad 3.28$$

The input voltage is applied to the primary section of the transformer to accumulate energy when the switching element is in the conduction state. When the switching element is in the cut-off state, the energy accumulated in the primary section of the transformer is transferred to the secondary section, in other words, to the output of the flyback converter. This is related to the power generated at the input of the circuit. Using Eq. 2.24, the input power is given in Eq. 3.29. As stated in Eq. 3.29, the energy transfer in one second, that is, the energy transfer depending on the switching frequency, is stated in Eq. 3.30 while in Eq. 3.30, I_p^{peak} is the primary peak current and calculated in Eq. 3.31.

$$P_{in} = \frac{P_{out}}{\eta} = \frac{36}{0.8} = 45W \quad 3.29$$

$$P_{in} = \frac{1}{2} L_p I_p^{peak^2} f_{sw} \quad 3.30$$

$$45W = \frac{1}{2} * 670 * 10^{-9} * I_p^{peak^2} * 45 * 10^3, I_p^{peak} = 1.72A \quad 3.31$$

The average current value in the primary winding of the transformer is determined by the duty cycle and the minimum input voltage value. Eq. 3.32 specifies the average current value in the duty primary winding and Eq. 3.33 specifies the effective value.

$$I_p^{mean} = \frac{I_p^{peak} D}{2} = \frac{1.72 * 0.56}{2} = 0.48A \quad 3.32$$

$$I_p^{RMS} = \frac{1}{\sqrt{3}} I_p^{peak} \sqrt{D} = \frac{1}{\sqrt{3}} * 1.72 * \sqrt{0.56} = 0.74A_{RMS} \quad 3.33$$

The secondary current peak value of the transformer is given in Eq. 3.34 and the secondary effective value is given in Eq. 3.35.

$$I_s^{peak} = I_p^{peak} \frac{N_p}{N_s} = 1.72 * \frac{32}{8} = 6.88A \quad 3.34$$

$$I_s^{RMS} = \frac{1}{\sqrt{3}} I_p^{peak} \sqrt{1-D} = \frac{1}{\sqrt{3}} * 1.72 * \sqrt{1-0.5} = 0.65A_{RMS} \quad 3.35$$

Primary peak current and number of turns of the transformer are found, the flux density of the core is calculated when the primary magnetisation inductance. Eq. 3.36 specifies the flux density of the core while B denotes the flux density of the core and calculated in the Eq. 3.37.

$$L_p I_p^{peak} = N_{PS} B A_e \quad 3.36$$

$$670 * 10^{-6} * 1.72 = 32 * B * 121 * 10^{-6}, B = 0.29T \cong 0.3T \quad 3.37$$

Transformer winding wire American Wire Gauge table is taken as reference. Table 2 indicates the AWG table.

Table 2 (METERS UK, 2019)

American Wire Gauge

AWG	Diameter [inches]	Diameter [mm]	Area [mm ²]	Resistance [Ohms / 1000 ft]	Resistance [Ohms / km]	Max Current [Amperes]	Max Frequency for 100% skin depth
21	0.0285	0.7239	0.41	12.8	41.984	1.2	33 kHz
22	0.0254	0.64516	0.326	16.14	52.9392	0.92	42 kHz
23	0.0226	0.57404	0.258	20.36	66.7808	0.729	53 kHz

In this table, wires that can pass current over 0.48A minimum in cross-sectional mm² are determined. Copper wire with a cross-sectional area of 0.35mm² was selected for the primary winding due to cost and small carcass area. This wire passes a maximum current of 0.92A. The primary winding of the transformer receives an average current of 0.48A. When winding twice with a single wire, a maximum current of 1.84A can pass through the wire. For the secondary winding, a maximum current of 7.36A can pass through the wire when it is wound twice using 4 wires, in other words, bifilar winding. Transformer winding information is specified in Annex A.1.

According to the calculated parameters, transformer winding numbers, winding coefficients and magnetisation inductances are given in Table 3.

Table 3

Transformer Parameters

Transformer Parameters		
Parameter	Symbol	Value
Number of Primary Windings	N_P	32
Secondary Winding Number	N_S	8
Number of Tertiary Windings	N_T	4
Primary, Secondary, Tertiary Winding Ratio	$N_{PS}: N_{SP}: N_{TS}$	4:1:0.5
Primary Magnetisation Inductance	L_p	$670\mu H$
Secondary Magnetisation Inductance	L_s	$40\mu H$
Tertiary Magnetisation Inductance	L_t	$10\mu H$

3.2.2 MOSFET voltage, current and power value calculations. The MOSFET is selected by determining the V_{DS} voltage, internal resistance R_{DS}^{on} and power dissipation parameters. Since the maximum input voltage of the flyback converter is 137.5VDC, the V_{DS} voltage should be minimum 137.5VDC. R_{DS}^{on} resistance should be in the order of m Ω . The MOSFET releases heat energy during switching on and off and due to the current flowing through it. R_{DS}^{on} resistance value is important due to its thermal performance and less energy consumption. The maximum current across the MOSFET is the same as the current in the primary section of the transformer. When the MOSFET switches from conduction to cut-off state, the voltage induced in the secondary of the transformer is reflected to the primary of the transformer and the reflected voltage falls on the MOSFET. The value of the minimum V_{DS} voltage falling on the MOSFET is found by adding the maximum input voltage, the voltage reflected on the primary of the transformer, and the peak voltage of the MOSFET during switching. In practice, the resonant voltage should be approximately 50% more than the maximum input voltage. Eq. 3.38 specifies the minimum V_{DS} voltage while it is calculated in Eq. 3.39. Eq. 3.40 specifies the value of the voltage reflected from the secondary to the primary of the transformer while it is calculated in Eq. 3.41 and Eq. 3.42 specifies the ringing value on the MOSFET at maximum input voltage while it is calculated in Eq. 3.43.

$$V_{DS}^{min} = V_{in}^{max} + V_{rf} + V_M^{ring} \quad 3.38$$

$$V_{rf} = (V_{out} + V_f) \frac{N_P}{N_S} \quad 3.39$$

$$V_{rf} = (24 + 0.8) * 4 = 99.2VDC \quad 3.40$$

$$V_M^{ring} = (V_{in}^{max} + V_f) 50\% + V_{in}^{max} \quad 3.41$$

$$V_M^{ring} = ((137.5 + 0.8) * 50\%) + 137.5 = 206.65VDC \quad 3.42$$

$$V_{DS}^{min} = 137.5 + 99.2 + 206.65 = 443.35VDC \quad 3.43$$

For cost and rail standards reasons, the FCB125N65S3 MOSFET was specified in accordance with the calculated MOSFET values. V_{DS} voltage is 650VDC, R_{DS}^{on} is 125mΩ and I_{DS} current is 20A under normal conditions specified in the datasheet.

Power losses occur on the MOSFET in the conduction state and due to switching. The sum of the conduction and switching power losses gives the total power loss on the MOSFET. Eq. 3.44 shows the total power dissipation equation on the MOSFET, Equation 3.45 shows the power dissipation in the conduction state, Eq. 3.46 shows the calculation of 3.47, Eq. 3.47 shows the switching power dissipation, Eq. 3.48 shows the calculation of 3.47 and Eq. 3.49 shows the total power dissipation calculation. In Eq. 3.44 P_M^{loss} is the total MOSFET loss, P_{MC}^{loss} is the conduction power loss of the MOSFET, P_{MSW}^{loss} is the switching power loss. In Eq. 3.47, C_{OSS} is the output capacitance value of the MOSFET, t_r is the Turn-On Rise Time of the MOSFET, t_f is the Turn-Off Fall Time.

$$P_M^{loss} = P_{MC}^{loss} + P_{MSW}^{loss} \quad 3.44$$

$$P_{MC}^{loss} = I_{p_{peak}}^2 (R_{DS}^{on} + R_{sns}) \quad 3.45$$

$$P_{MC}^{loss} = 1.72^2 * (0.125 + 0.05) = 0.38W \quad 3.46$$

$$P_{Msw}^{loss} = \left(\frac{1}{2} C_{OSS} (V_{in} + V_f)^2 f_{sw} \right) + \left(\frac{1}{2} (V_{in} + V_f) I_p^{peak} (t_r + t_f) f_{sw} \right) \quad 3.47$$

$$P_{Msw}^{loss} = \left(\frac{1}{2} * 40 * 10^{-9} * (137.5 + 0.8)^2 * 45 * 10^3 \right) + \left(\frac{1}{2} * (137.5 + 0.8) * 1.72 * (26 * 10^{-9} + 17 * 10^{-9}) * 45 * 10^3 \right) = 0.34W \quad 3.48$$

$$P_M^{loss} = 0.38 + 0.34 = 0.72W \quad 3.49$$

In addition, since the maximum input voltage can operate at 600VDC due to the LT8316 integration, a MOSFET suitable for the MOSFET pad on the PCB is selected by calculating the equations under section 3.2.2 according to the request, and a flyback converter with 24VDC output voltage and 1.5A output current can be easily designed by calculating the side components without designing a separate transformer. In this case, without working on a new PCB board for the flyback converter to be designed according to the new requirement, the cost is reduced and the product variety increases.

3.2.3 Snubber design calculations. The leakage inductances in the transformer cause ringing on the MOSFET, in other words, voltage spikes when the MOSFET switches to the cut-off state. If the leakage inductances in the transformer are high, the voltage ringing will be high. In this case, the leakage inductances in the transformer must be at minimum level. The voltage spike causes the MOSFET not to operate in the safe region, EMI problem, overloading the transformer, switching loss, output voltage fluctuation of the converter and decreases the efficiency. The voltage rise on the MOSFET is reduced by converting it into heat energy with a resistor and a capacitor and this is called passive snubber. In order to design the snubber on the circuit, it is necessary to know the frequency or period of the ringing. The period value of the ringing was found by measuring with FLUKE brand 199B model numbered oscilloscope and T5100 probe. As stated in the data sheet of the LT8316 integration, 100pF value is first given for the snubber and the required capacitor value is found by increasing the 100pF value according to the oscilloscope measurement. Equation 3.50 shows the snubber capacitor C_s while calculated in Eq. 3.51. T_{ring}^{sn} 250ns, T_{ring} 185ns were measured with oscilloscope and C_{sw} 200pF was determined.

$$C_s = \frac{C_{sw}}{\left(\frac{T_{ring}^{sn}}{T_{ring}}\right)^2 - 1} \quad 3.50$$

$$C_s = \frac{200 * 10^{-12}}{\left(\frac{250 * 10^{-9}}{185 * 10^{-9}}\right)^2 - 1} = 242pF \cong 300pF \quad 3.51$$

Snubber resistance is calculated with the leakage inductance value at the transformer primary. In Eq. 3.52, the leakage inductance at the transformer primary is calculated while the symbol L_p^{leak} indicates the leakage inductance at the transformer primary.

$$L_p^{leak} = \left(\frac{T_{ring}}{2\pi} \right)^2 \frac{1}{C_s} = \left(\frac{185 * 10^{-9}}{2\pi} \right)^2 \frac{1}{300 * 10^{-12}} = 2.8 \mu H \quad 3.52$$

Eq. 3.53 shows the snubber resistance value R_s .

$$R_s = \sqrt{\frac{L_p^{leak}}{C_s}} = \sqrt{\frac{2.8 * 10^{-6}}{300 * 10^{-12}}} = 96.6 \Omega \cong 100 \Omega \quad 3.53$$

The resistance R_s is selected as 120Ω .

Using Eq. 3.54, the power dissipation value on the snubber is shown in Eq. 3.55.

$$P_s = C_s (V_{in}^{max} + V_f + V_{DS}^{min})^2 f_{sw} \quad 3.54$$

$$P_s = 300 * 10^{-12} * (137.5 + 0.8 + 443.35)^2 * 40 * 10^3 = 4.05 W \quad 3.55$$

3.2.4 Output capacitor value and output ripple calculations. The output capacitor absorbs the voltage fluctuations on the rectified voltage at the output of the converter and stabilises the converter when instantaneous fluctuations occur in the load of the converter. It absorbs output voltage and current fluctuations when the input voltage of the converter changes. The output capacitor should have low ESR and the output voltage ripple of the converter should be kept below 100mV. Eq. 3.56 shows the output ripple voltage V_{out}^{ripple} and Eq. 3.58 shows the calculation of Eq. 3.57. The value of C_o is chosen to be 220μF because it is easily available and inexpensive.

$$V_{out}^{ripple} = \frac{L_p (I_p^{max})^2}{2C_o V_{out}} \quad 3.56$$

$$I_p^{max} = \frac{100mV}{R_{SNS}} = \frac{100mV}{50m\Omega} = 2 \quad 3.57$$

$$V_{out}^{ripple} = \frac{670 * 10^{-6} * 2^2}{2 * 220 * 10^{-6} * 24} = 253mV \quad 3.58$$

Since V_{out}^{ripple} value is not below 100mV, V_{out}^{ripple} will decrease below 100mV when the capacitor value is increased. The large capacitor value brings the cost problem while increasing the capacitor volume. In this case, due to the cost and size, the output voltage ripple is reduced below 100mV with a lower value, simple design and passive low-pass LC filter design. The LC filter reduces the voltage fluctuations caused by the material tolerance and transformer leakage inductances. In addition, the LC filter reduces the losses and increases the efficiency of the converter by absorbing the ringing on the diode without the need for a snubber circuit on the output diode. Eq. 3.58 shows the LC filter value and Eq. 3.60 shows the calculation of Eq. 3.59. In Eq. 3.59 f_{co}^{LC} is the filter cut-off frequency, L_{LC} is the filter inductance value and C_{LC} is the filter capacitance value. The value of L_{LC} is determined as 33μH.

$$f_{co}^{LC} = \frac{1}{2\pi\sqrt{L_{LC}C_{LC}}} \quad 3.59$$

$$f_{co}^{LC} = \frac{1}{2 * \pi * \sqrt{33 * 10^{-6} * 440 * 10^{-6}}} 1.8kHz \quad 3.60$$

The output capacitor value must be at least 220μF and if a capacitor higher than 220μF is used, it will reduce the cut-off frequency. Since the switching frequency of the converter is 45kHz, the corner frequency is 4.5kHz and the output filter frequency below 4.5kHz does not pose a problem, but it will further reduce the output ripple voltages.

3.2.5 Output diode calculation. The output diode is used to rectify the AC voltage in the secondary section of the transformer. The output diode affects the efficiency, performance and reliability of the converter due to thermal performance. The output diode should be selected from schottky diodes with low internal resistance. Eq. 3.61 shows the minimum voltage value of the output diode and Eq. 3.62 shows the calculation of Eq. 3.61.

$$V_D^{min} = V_{out} + \left(V_{in}^{min} + \frac{N_S}{N_P} \right) \quad 3.61$$

$$V_{D_{min}} = 24 + \left(137.5 * \frac{1}{4} \right) = 58.375\text{VDC} \quad 3.62$$

In practice, the rated current of the output diode should be 50% higher than the output current value. Eq. 3.63 shows the output diode current value.

$$I_D = I_{ut} + (I_{out} 50\%) = 1.5 + (1.5 * 50\%) = 2.25\text{A} \quad 3.63$$

For reasons of cost and rail standards, MBRD10100CT schottky diode was determined in accordance with the calculated output diode values. V_f voltage is 0.8VDC, 125m Ω and maximum current is 10A under normal conditions specified in the datasheet.

Power losses occur on the output diode in the conduction state. Eq. 3.64 shows the power loss on the output diode and Eq. 3.65 shows the calculation value of 3.64.

$$P_D^{loss} = V_f * I_{out} \quad 3.64$$

$$P_D^{loss} = 0.8 * 1.5 = 1.2W \quad 3.65$$

3.3 Control Detail of LT8316

Within the scope of the thesis, information about the LT8316 integration used in the flyback converter circuit is given in this section.

LT8316 is a high voltage controller that can provide up to 100W power without the need for an opto-isolator. It is capable of constant current and voltage regulation that can operate with minimum 16VDC and maximum 600VDC wide supply voltages. Designs suitable for constant current and constant voltage requirements can be made using LT8316. (Analog Devices, 2019) This reduces the number of products in stock of a company and reduces warehouse costs. Fig. 18 shows the internal structure of the LT8316 integration and Fig. 19 shows the pin configuration.

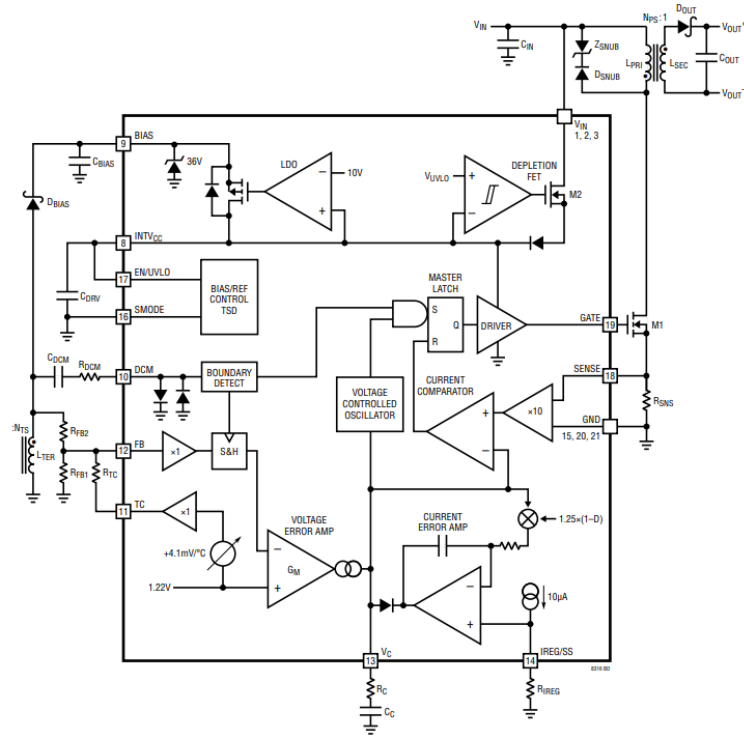


Figure 18. LT8316 Internal Structure (Analog Devices, 2019)

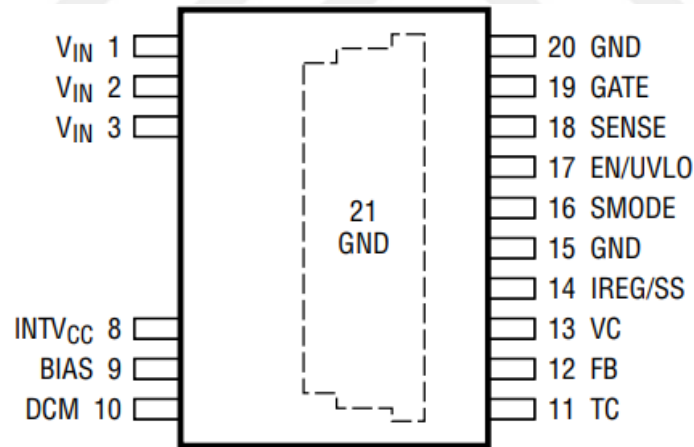


Figure 19. LT8316 Pin Configuration (Analog Devices, 2019)

Pins 1, 2 and 3 provide the supply voltage and supply the integration. Pin 8, Internal Gate Driver Bias Voltage, charges up to 12V using the current generated by pins 1, 2 and 3 when the controller starts the process. When the controller is running, BIAS pin 9 reduces the charging voltage to 10V. BIAS pin 9 provides power to the

INTVcc pin while absorbing the irregular input voltages to the controller. DCM pin 10 samples the output voltage by taking the derivative of the controller's switching frequency waveform and determines the mode. Pin 11, Temperature Compensation, regulates the waveforms and operation of the output diode and prevents temperature dependent variations. 12 numbered Feedback pin determines the output voltage according to the 1.22V reference voltage. Pin 13, Loop Compensation, uses the feedback voltage to set the limit current while adjusting the switching frequency for voltage and power distribution. Pin 14, Current Regulation/Soft-Start, enables the controller to be switched on gradually with voltage to consume less energy. By flowing 10uA current through it, it compares it with the voltage on it and adjusts the output current. Pin 16 Standby Mode Pin reduces the switching frequency for low power mode. Pin 17, Enable/Undervoltage Lockout, determines the input voltage range. The controller can be switched on and off in the desired range. Pin 18 Current Sense is used for current mode and limitation. Pin 19 Gate Driver Output sends a signal to the switching element to switch it on and off. Pins 15, 20 and 21 are connected to ground. (Analog Devices, 2019)

3.4 Flyback Converter Circuit Simulation

In this part of the thesis, simulation is performed in LTspice programme according to the calculated parameters and simulation results are shown.

Fig. 20 shows the simulation circuit diagram.

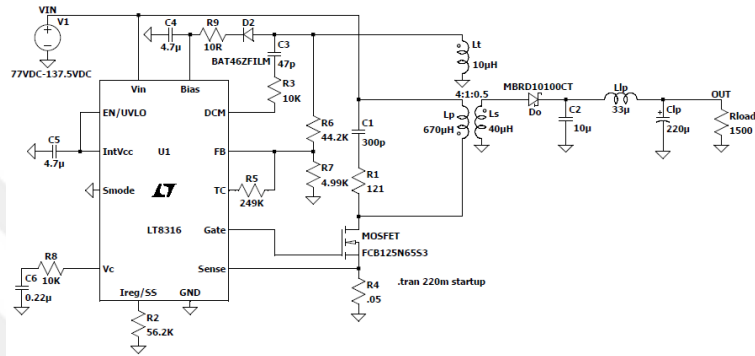


Figure 20. LT8316 LTspice Circuit Diagram

Fig. 21 shows the output voltage. V_{out} is observed as 24.2VDC in the input voltage range of 77VDC-137.5VDC in simulation.

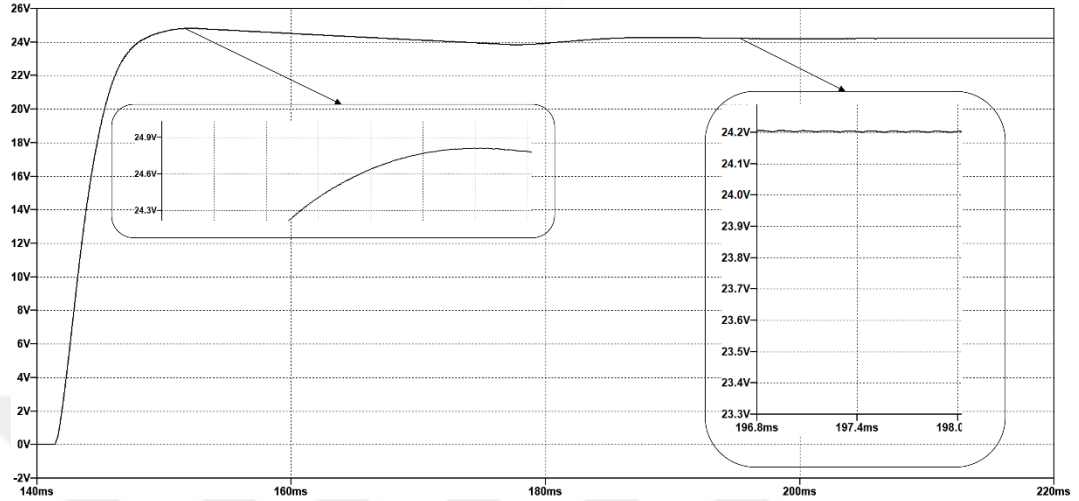


Figure 21. Ltspice Flyback Converter Output Voltage

Fig. 22 shows the MOSFET Drain-Source voltage. The V_{DS} voltage of the MOSFET is observed as 180VDC.

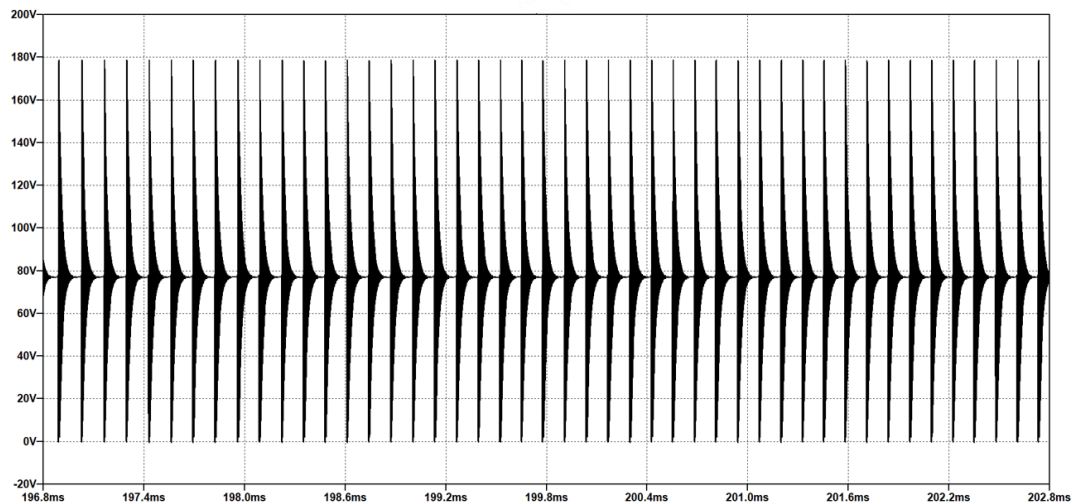


Figure 22. Ltspice MOSFET Drain-Source Voltage

Fig. 23 shows the MOSFET Drain-Source current is observed as 1.8A.

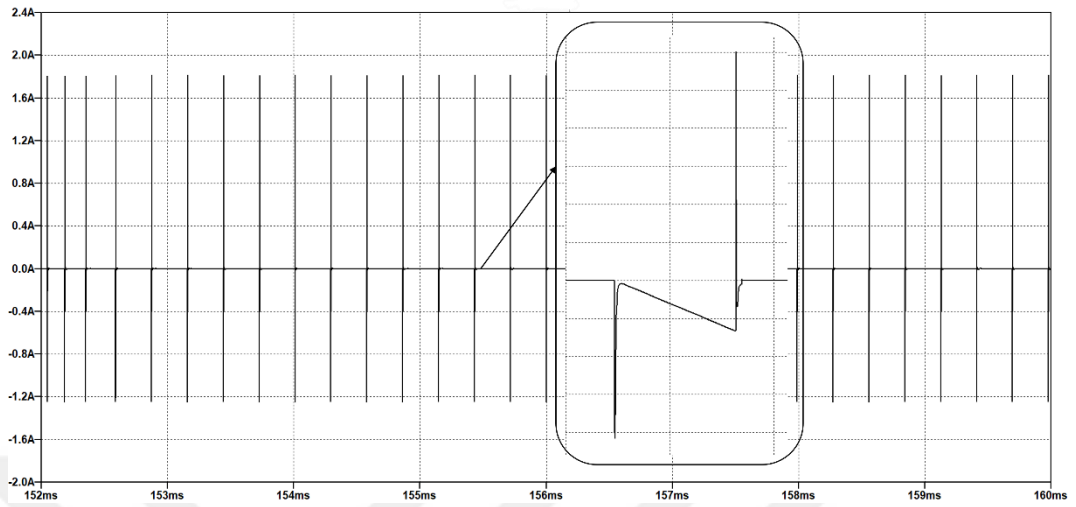


Figure 23. Ltspice MOSFET Drain-Source Current

Fig. 24 shows the primary current of the transformer. The value of I_p^{max} is observed as 1.5A.

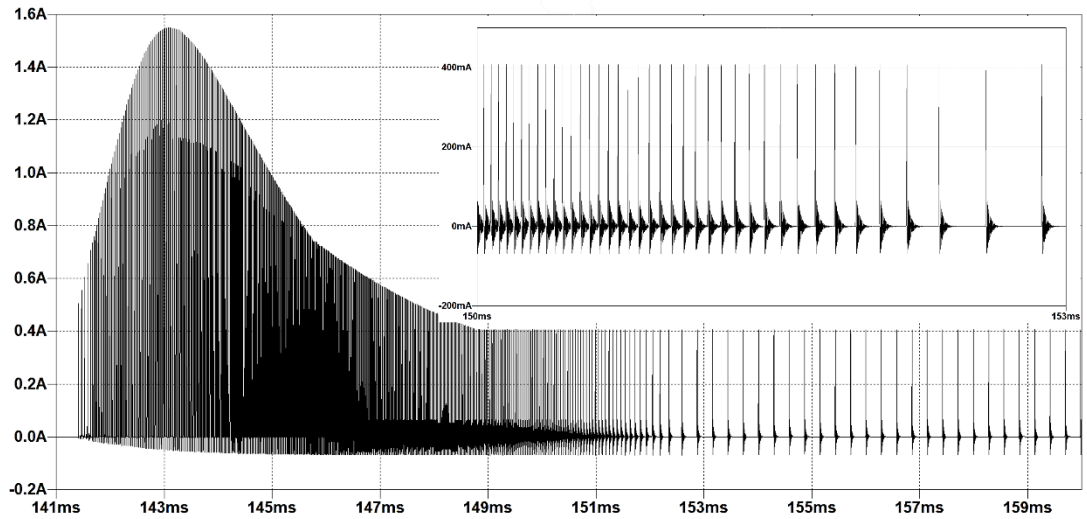


Figure 24. Ltspice Transformer Primer Current

Fig. 25 shows the voltage graph on the output diode. The voltage V_D on the output diode is observed as 24.9VDC and minimum -18VDC.

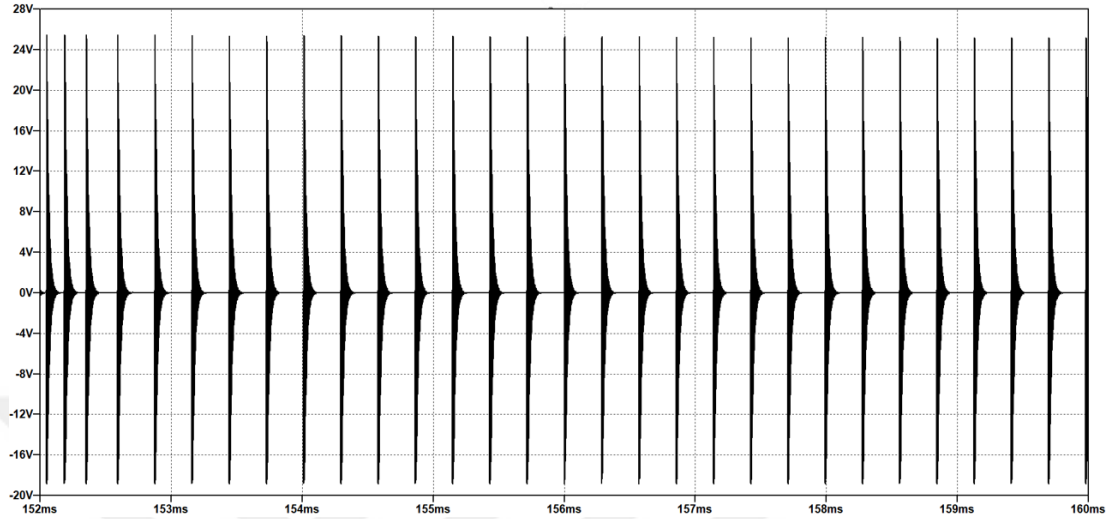


Figure 25. Ltpice Output Diode Voltage

Fig. 26 shows the current graph on the output diode. The current I_D^{max} on the output diode is observed as 6.09A.

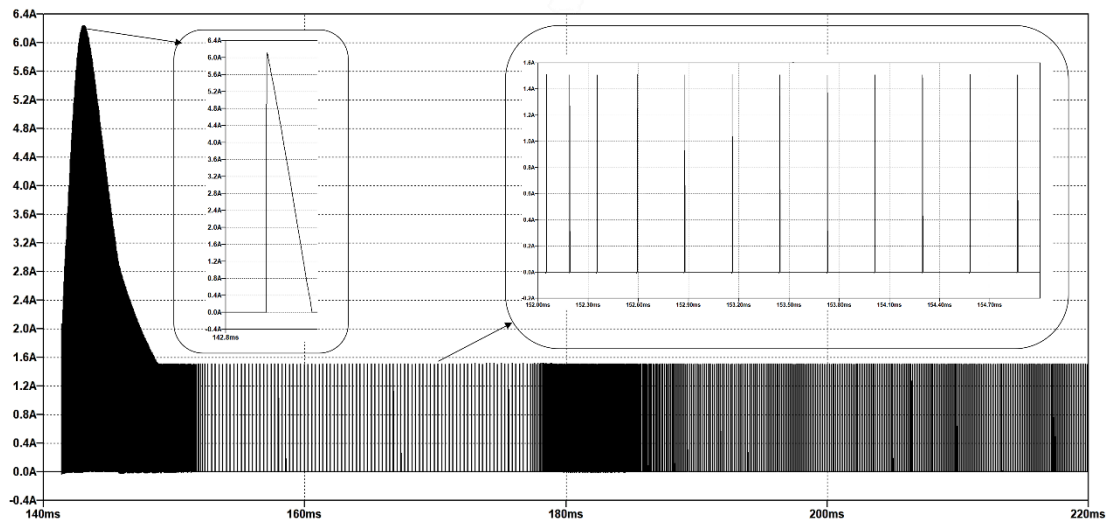


Figure 26 Ltpice Flyback Converter Output Diode Current

3.5 EMC Filter Design

In this section, an EMC filter in accordance with railway standards is designed, tested and verified. In addition, EN 50155 Railways Applications, TS EN 50121-3-2 standards are mentioned.

EN 50155 standard railway power converter 110VDC applications;

- 77VDC-137.5VDC input voltage range,
- At nominal 110VDC input voltage,
- 24VDC or 12VDC output voltage

power supplies are used in the conditions of -40°C / $+70^{\circ}\text{C}$.

The power supply temperature range should be -40°C / $+70^{\circ}\text{C}$.

TS EN 50121-3-2 EMC requirements in DC power applications;

- At frequencies between 150kHz-500kHz, there should be no noise at db higher than 99 dB μ V peak value.
- At frequencies between 500kHz-30MHz, there should be no noise at db higher than 93 dB μ V peak value.

EMC filter should be designed according to the standards while the line impedance for the EMC filter and the input impedance of the designed converter must be determined. The impedance of the EMC filter should not be greater than the impedance of the converter. If the filter impedance is greater than the converter impedance, the converter will not operate correctly because the frequencies will be over-cut. In the filter, two filter designs are made as differential mode and common mode. Differential mode is used for low frequencies and common mode is used for high frequencies.

In order to find the input impedance of the flyback converter, it is necessary to calculate the input current. Eq. 3.66 specifies the input current of the flyback converter while in Eq. 3.67, the input current is calculated and η is set as 80%.

$$I_{in} = \frac{V_{out} I_{out}}{V_{in}^{max} \eta} \quad 3.66$$

$$I_{in} = \frac{24 * 1.5}{137.5 * 0.8} = 0.32\Omega \quad 3.67$$

Eq. 3.68 shows the input impedance of the circuit while in calculated. Eq. 3.69. In Eq. 3.68, Z_{in} indicates the input impedance of the converter.

$$Z_{in} = \frac{V_{in}^{max}}{I_{in}} \quad 3.68$$

$$Z_{in} = \frac{137.5}{0.32} = 429.6875\Omega \cong 430\Omega \quad 3.69$$

Eq. 3.70 shows the impedance of the EMC filter circuit while calculated in equation 3.71. In Eq. 3.70, Z_{EMC} is the impedance of the EMC filter.

$$Z_{EMC} = \frac{1}{10} * Z_{in} \quad 3.70$$

$$Z_{EMC} = \frac{1}{10} * 430 = 43\Omega \cong 40\Omega \quad 3.71$$

The filter impedance cannot be higher than 40Ω . If it is higher than 40Ω , the switching frequency of the circuit changes and the converter does not work and gives different responses. After the impedance values are determined, the minimum capacitance value is found on the impedance paper and the maximum inductance value is determined with the low pass filter structure. The minimum capacitance value for the EMC filter C_{DM} is $1.5\mu F$. For differential mode, the maximum L_{DM} value is determined by selecting C_{DM} value $1.5\mu F$ in Eq. 3.72. It is calculated in Eq. 3.73. The cut-off frequency of the EMC filter is one tenth of the switching frequency. In Eq. 3.75, f_{co}^{EMC} is the cut-off frequency of the EMC filter, L_{DM} is the differential mode inductance value and C_{DM} is the differential mode capacitance value.

$$f_{co}^{EMC} = \frac{1}{2\pi\sqrt{L_{DM}C_{DM}}} \quad 3.72$$

$$4.5 * 10^3 = \frac{1}{2\pi\sqrt{L_{DM}^{max} * 1.5 * 10^{-6}}}, L_{DM}^{max} = 833\mu H \quad 3.73$$

Capacitors with the calculated capacitance value are not available due to cost, stock and capacitance value. In order to avoid this problem, the filter impedance can be reduced by increasing the capacitor value. Thus, C_{DM} is set as $10\mu F$ and L_{DM} value is set as $150\mu H$. In order to better dampen the noise in the EMC filter, an electrolytic capacitor of at least 5 times the value of C_{DM} is added. Eq. 3.74 shows the minimum C_D value while in Eq. 3.75 calculated and in Eq. 3.74 C_D refers to the dump capacitor.

$$C_D > 5C_{DM} \quad 3.74$$

$$C_D > 50\mu F \quad 3.75$$

C_D is determined to be 150uF.

The magnitude of the switching frequency is important when designing the common mode. If the switching frequency is small, the common mode inductance value should also be small. Within the scope of the thesis, 744822301 coded Common Mode Choke produced by Würth company was selected. By looking at the datasheet, the minimum common mode capacitor value is determined by using Eq. 3.76 since it has a value of 1mH and calculated in the Eq. 3.77. In Eq. 3.76, L_{CM} indicates the common mode inductance value and C_{CM}^{min} indicates the common mode minimum capacitance value.

$$f_{co}^{EMC} = \frac{1}{2\pi\sqrt{L_{CM}C_{CM}^{min}}} \quad 3.76$$

$$4.5 * 10^3 = \frac{1}{2\pi\sqrt{1 * 10^3 * C_{CM}^{min}}}, C_{CM}^{min} = 1.25pF \quad 3.77$$

C_{CM} value is selected with a value of 1uF.

Fig. 27 shows the EMC filter circuit designed according to the capacitance and inductance values determined.

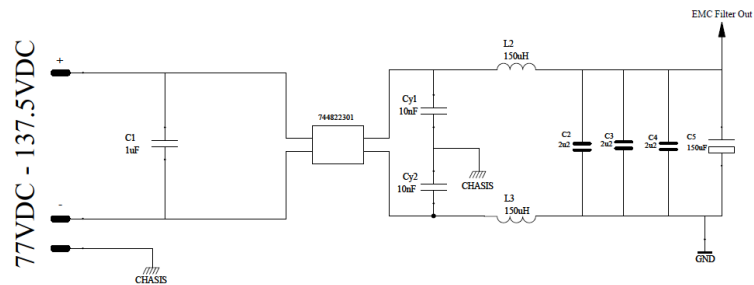


Figure 27 EMC Filter Circuit

Chapter 4

Experimental Results

4.1 Introduction

In this part of the thesis, the converter simulated in LTspice programme with theoretical calculations has been verified and a two-layer PCB has been drawn in EASY-PC programme for practical application. Fig. 28 shows the prototype two-layer PCB. The PCB was printed on a copper plate on a Promat200 machine and the components were soldered and assembled on the PCB. Full load, low load, low temperature and high temperature (-40°C and $+70^{\circ}\text{C}$) tests were performed on the PCB to verify the converter in a laboratory environment. Thermal images of the prototype circuit were measured with FLUKE brand Ti9 model thermal camera and oscilloscope images were measured with FLUKE brand 199B model oscilloscope and Twintex brand T5100 model two probes. The results of the prototype circuit without EMC filter and with EMC filter are also given below. The current measured regions were calculated using ohm's law with the determined resistance value.

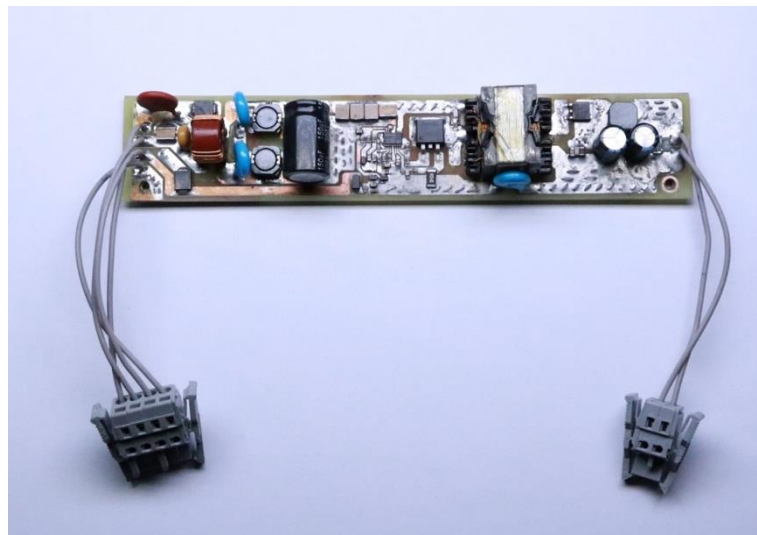


Figure 28. Prototype Flyback Converter

4.2 Input and Output Voltages at 100% (1.5A) Load

The input and output voltages of the designed converter at 77VDC, 110VDC and 137.5VDC input voltages at 1.5A 100% load are shown with oscilloscope as indicated in Table 1.

Fig. 29 shows the output voltage at 100% load when 77VDC minimum input voltage is applied. In Fig. 29, symbol A indicates the 77VDC input voltage value and symbol B indicates the output voltage. Fig. 29 is analysed, 24.4VDC output voltage is generated when 77VDC minimum input voltage is applied to the flyback converter at 100% load.

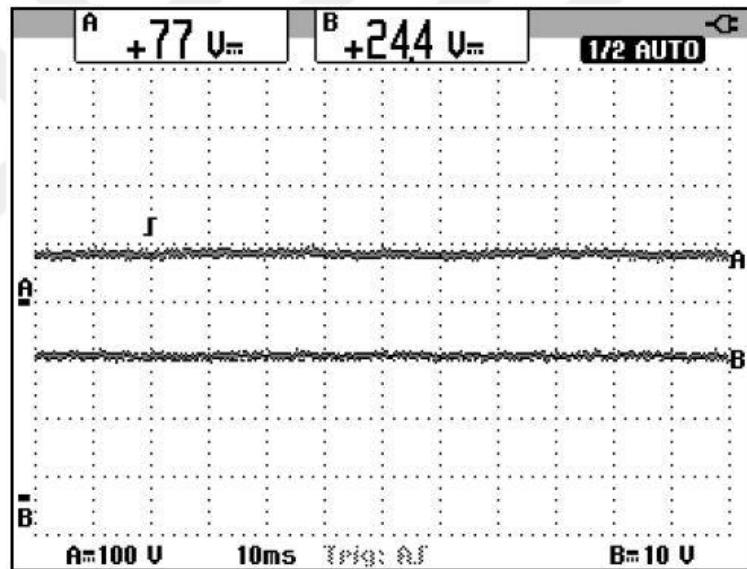


Figure 29. Prototype Flyback Converter Output Voltage at 100% Load (1.5A) at Minimum Input Voltage (77VDC)

Fig. 30 shows the output voltage at 100% load when 110VDC nominal input voltage is applied to the converter. In Fig. 30, symbol A indicates the 110VDC input voltage value and symbol B indicates the output voltage value. Fig. 30 is analysed, 24.4VDC output voltage is generated when 110VDC nominal input voltage is applied to the flyback converter at 100% load.

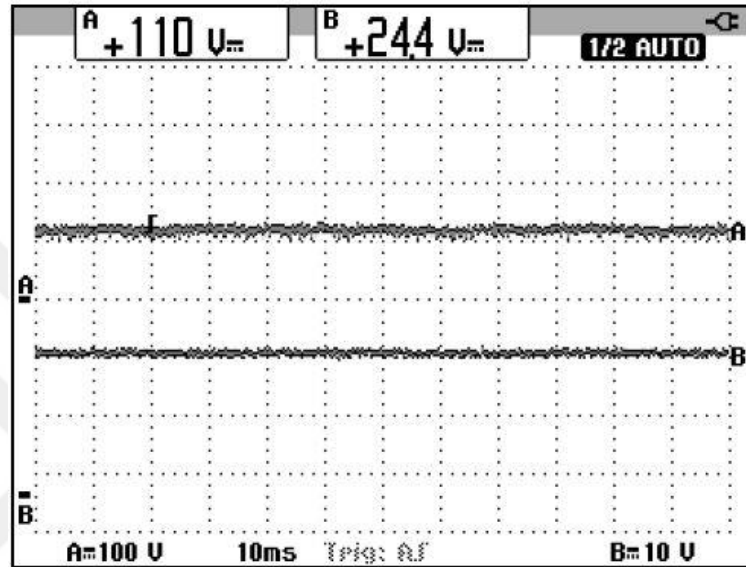


Figure 30. Prototype Flyback Converter Output Voltage at 100% Load (1.5A) at Nominal (110VDC) Input Voltage

Fig. 31 shows the output voltage at 100% load when 137.5VDC maximum input voltage is applied to the converter. In Fig. 31, symbol A indicates the input voltage value of 37.5VDC and symbol B indicates the output voltage value. Fig. 31 is analysed, 24.4VDC output voltage is generated when 137.5VDC maximum input voltage is applied to the flyback converter at 100% load.

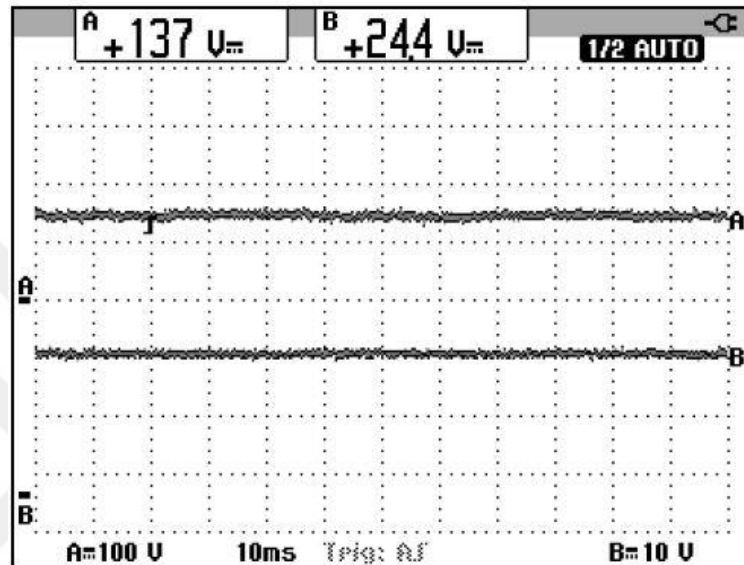


Figure 31 Prototype Flyback Converter Output Voltage at 100% Load (1.5A) at Maximum (137.5VDC) Input Voltage

4.3 Output Ripple Voltages at 100% Load (1.5A)

The output ripple voltage value specified in Table 1 must be less than 100mV. Fig. 32 shows the output ripple voltage at 77VDC minimum input voltage and 100% (1.5A) load. In Fig. 32, symbol A indicates the 77VDC minimum input voltage value applied to the converter and symbol B indicates the output ripple voltage value. Fig. 32 is analysed, it is observed that the minimum peak value of output ripple is -12.4mV, the maximum peak value is 8.80mV and the peak-to-peak value is 21.2mV.

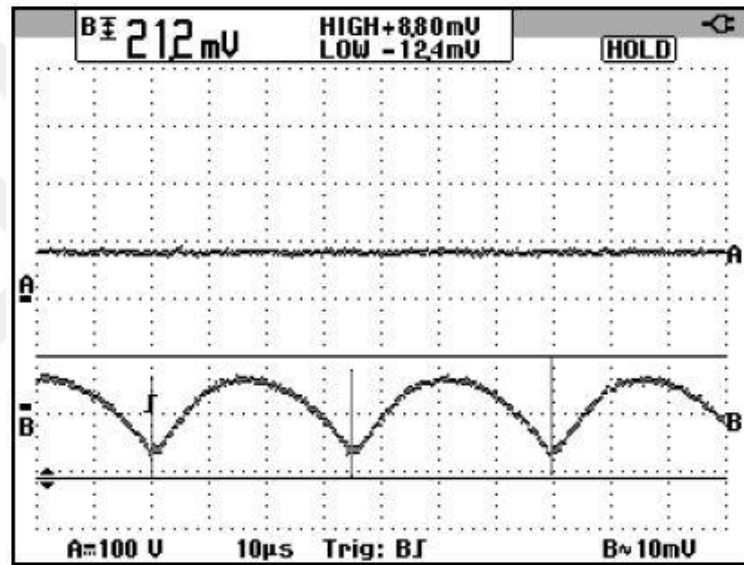


Figure 32 Prototype Flyback Converter Output Ripple Voltage at 100% Load (1.5A) at Minimum (77VDC) Input Voltage

Fig. 33 shows the output ripple voltage at 110VDC nominal input voltage and 100% load. In Fig. 33, symbol A indicates the 110VDC nominal input voltage value applied to the converter and symbol B indicates the output ripple voltage value. Fig. 33 is analysed, it is observed that the minimum peak value of output ripple is -12.8mV, the maximum peak value is 8.40mV and the peak-to-peak value is 21.2mV.

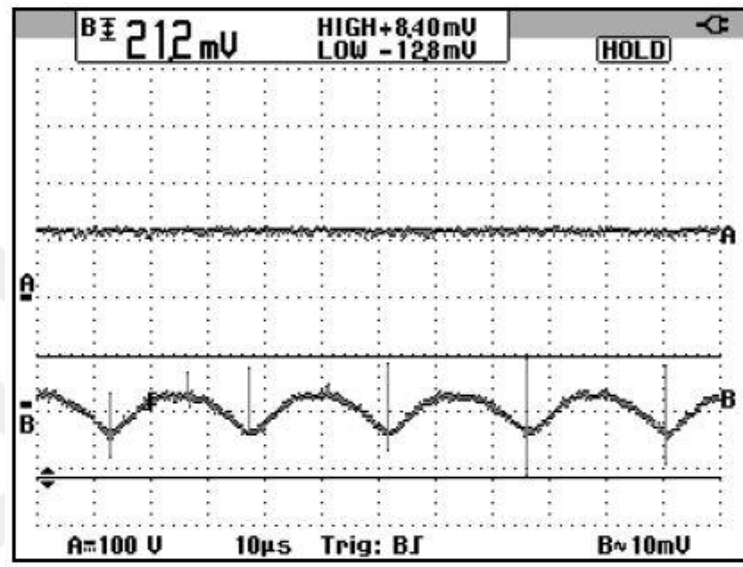


Figure 33. Prototype Flyback Converter Output Ripple Voltage at 100% Load (1.5A) at Nominal (110VDC) Input Voltage

Fig. 34 shows the output ripple voltage at 137.5VDC maximum input voltage and 100% load. In Fig. 34, symbol A indicates the maximum input voltage value of 137.5VDC applied to the converter and symbol B indicates the output ripple voltage value. Fig. 34 is analysed, it is observed that the minimum value of output ripple is -9.60mV, the maximum value is 11.6mV and the peak-to-peak value is 21.2mV.

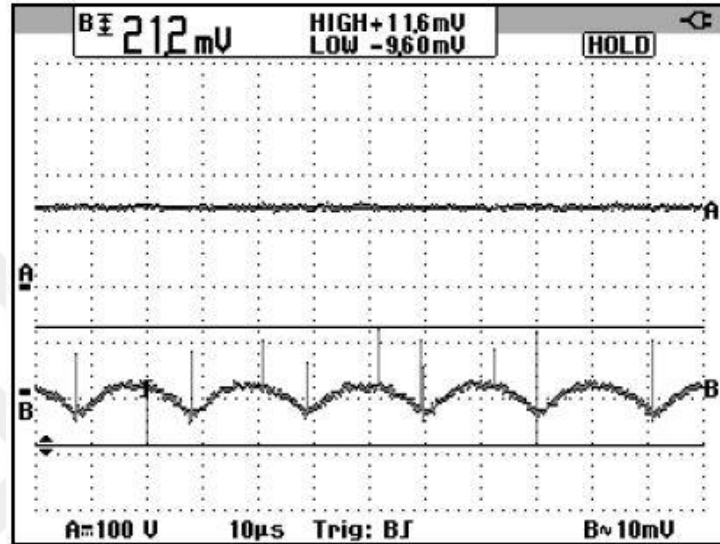


Figure 34. Prototype Flyback Converter Output Ripple Voltage at 100% Load (1.5A) at Maximum (137.5VDC) Input Voltage

4.4 Input and Output Voltages at 20% Load (300mA)

The input and output voltages of the designed converter are displayed on the oscilloscope at 77VDC, 110VDC and 137.5VDC input voltages at 300mA low load, in other words at 20% load. Fig. 35 shows the output voltage at 20% (300mA) load when 77VDC minimum input voltage is applied. In Fig. 35, symbol A indicates the 77VDC input voltage value and symbol B indicates the output voltage. Fig. 35 is analysed, 24.4VDC output voltage is generated when 77VDC minimum input voltage is applied to the flyback converter at 20% load.

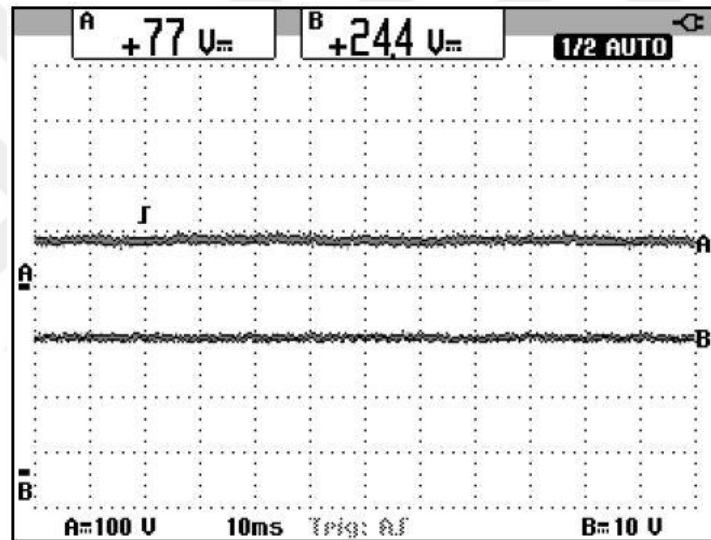


Figure 35. Prototype Flyback Converter Output Voltage at 20% Load (300mA) at Minimum (77VDC) Input Voltage

Fig. 36 shows the output voltage at 20% load when 110VDC nominal input voltage is applied to the converter. In Fig. 36, symbol A indicates the 110VDC input voltage value and symbol B indicates the output voltage. Fig. 36 is analysed, 24.4VDC output voltage is generated when 110VDC nominal input voltage is applied to the flyback converter at 20% load.

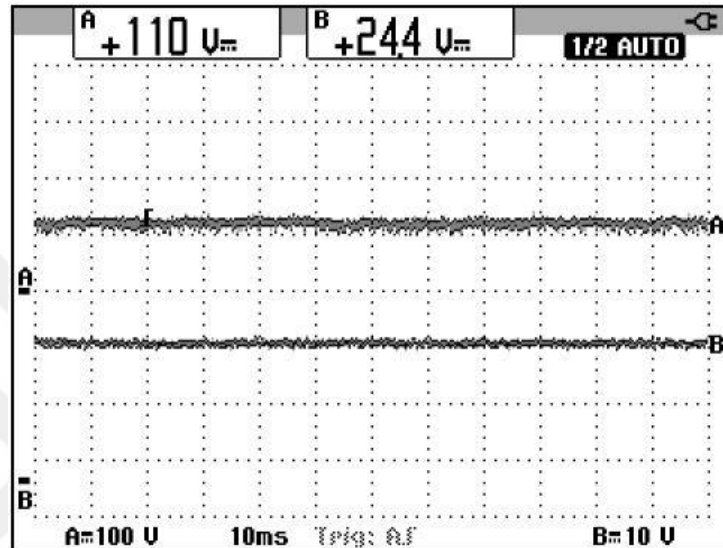


Figure 36. Prototype Flyback Converter Output Voltage at 20% Load (300mA) at Nominal (110VDC) Input Voltage

Fig. 37 shows the output voltage at 20% load when a maximum input voltage of 137.5VDC is applied to the converter. In Fig. 37, symbol A indicates 137.5VDC input voltage value and symbol B indicates output voltage. Fig. 37 is analysed, 24.4VDC output voltage is generated when 137.5VDC nominal input voltage is applied to the flyback converter at 20% load.

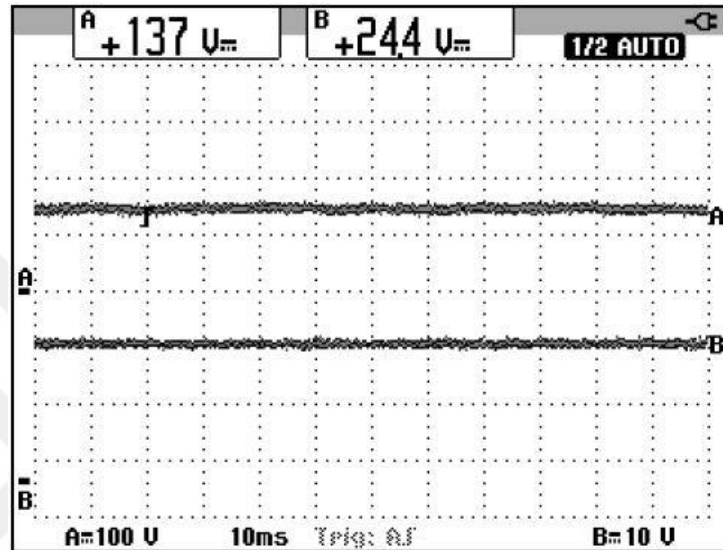


Figure 37. Prototype Flyback Converter Output Voltage at 20% Load (300mA) at Maximum (137.5VDC) Input Voltage

4.5 Output Ripple Voltages at 20% Load (300mA)

The output ripple voltage value specified in Table 1 should be less than 100mV even when the converter is operating at specified input voltages and low load. Fig. 38 shows the output ripple voltage at 77VDC minimum input voltage and 20% (300mA) load. In Fig. 38, symbol A indicates the 77VDC minimum input voltage value applied to the converter and symbol B indicates the output ripple voltage value. Fig. 38 is analysed, it is observed that the minimum value of output ripple is -3.20mV, the maximum value is 3.60mV and the peak-to-peak value is 6.80mV.

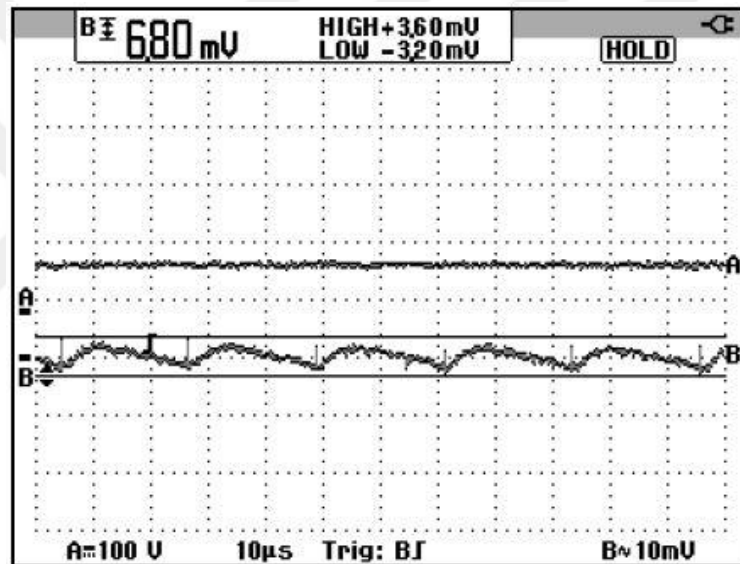


Figure 38. Prototype Flyback Converter Output Ripple Voltage at 20% Load (300mA) at Minimum (77VDC) Input Voltage

Fig. 39 shows the output ripple voltage at 110VDC nominal input voltage and 20% load. In Fig. 39, symbol A indicates the 110VDC nominal input voltage value applied to the converter and symbol B indicates the output ripple voltage value. Fig. 39 is analysed, it is observed that the minimum value of output ripple is -4.80mV, the maximum value is 6.40mV and the peak to peak value is 11.2mV.

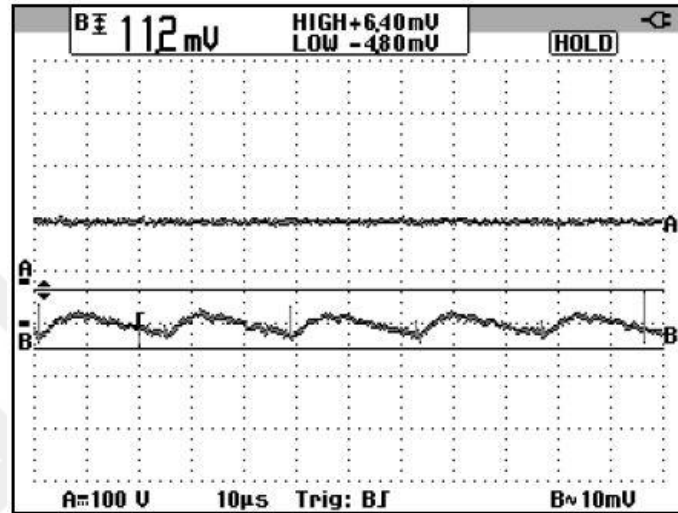


Figure 39. Prototype Flyback Converter Output Ripple Voltage at 20% Load (300mA) at Nominal (110VDC) Input Voltage

Fig. 40 shows the output ripple voltage at 137.5VDC maximum input voltage and 20% load. In Fig. 40, symbol A indicates the maximum input voltage value of 137.5VDC applied to the converter and symbol B indicates the output ripple voltage value. Fig. 40 is analysed, it is observed that the minimum value of output ripple is -9.60mV, the maximum value is 11.6mV and the peak to peak value is 21.2mV.

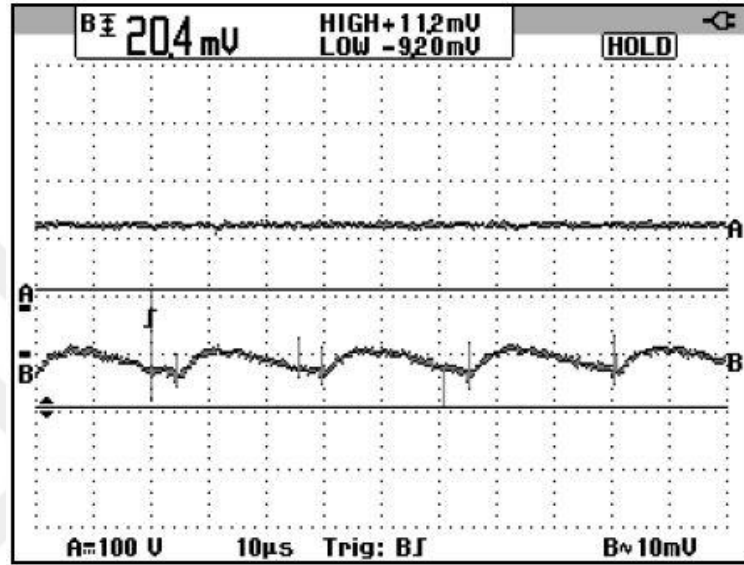


Figure 40. Prototype Flyback Converter Output Ripple Voltage at 20% Load (300mA) at Maximum (137.5VDC) Input Voltage

4.6 Transformer Secondary and Tertiary Winding Voltages at 100% Load (1.5A)

The current and voltage generated in the secondary and tertiary windings of the transformer at 100% load at 77VDC, 110VDC and 137.5VDC input voltages are displayed on the oscilloscope.

Fig. 41 shows the secondary voltage of the transformer at 100% load when 77VDC minimum input voltage is applied. In Fig. 41, symbol A indicates the 77VDC input voltage value and symbol B indicates the secondary voltage. Fig. 41 shows that the secondary voltage of the transformer has a maximum peak value of 31.2VDC, minimum peak value of -24.8VDC, peak-to-peak 56VDC, 25VRMS and a wave period of 35.20 μ s when 77VDC minimum input voltage is applied to the flyback converter at 100% load.

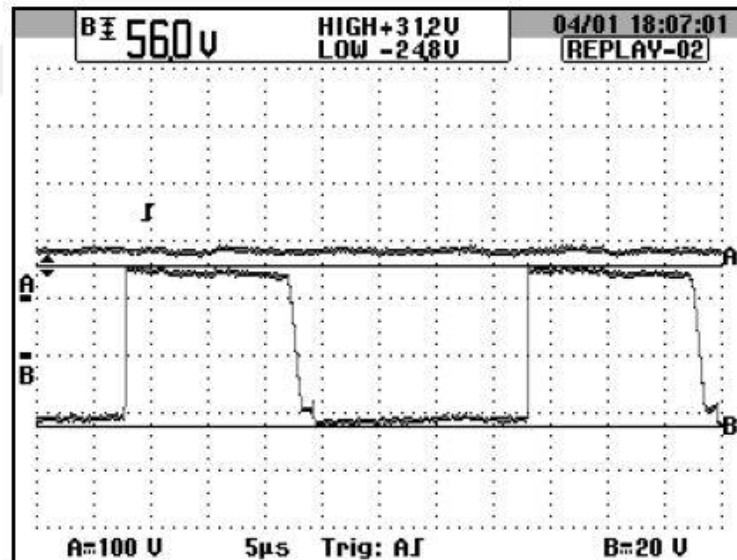


Figure 41. Prototype Flyback Converter Transformer Secondary Voltage at 100% Load (1.5A) at Minimum (77VDC) Input Voltage

Fig. 42 shows the secondary voltage of the transformer at 100% load when 110VDC nominal input voltage is applied. In Fig. 42, symbol A indicates the 110VDC input voltage value and symbol B indicates the secondary voltage. Fig. 42 shows that the secondary voltage of the transformer has a maximum peak value of 31.2VDC, minimum peak value of -35.2VDC, peak to peak 66.4VDC, 29.7VRMS and a wave period of 25.6 μ s when 110VDC nominal input voltage is applied to the flyback converter at 100% load.

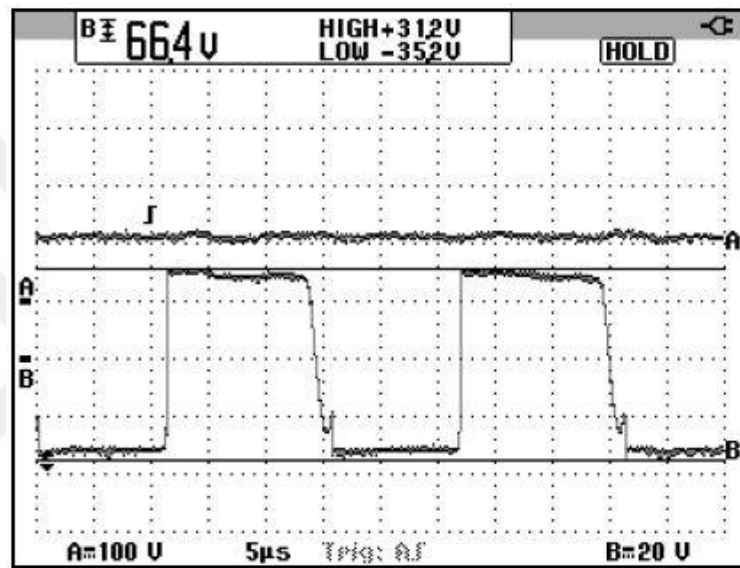


Figure 42. Prototype Flyback Converter Transformer Secondary Voltage at 100% Load (1.5A) at Nominal (110VDC) Input Voltage

Fig. 43 shows the secondary voltage of the transformer at 100% load when 137.5VDC maximum input voltage is applied. In Fig. 43, symbol A indicates the 137.5VDC input voltage value and symbol B indicates the secondary voltage. Fig. 43 shows that the secondary voltage of the transformer has a maximum peak value of 31.2VDC, minimum peak value of -42.4VDC, peak to peak 73.6VDC, 32.7VRMS and a wave period of 22 μ s when 110VDC nominal input voltage is applied to the flyback converter at 100% load.

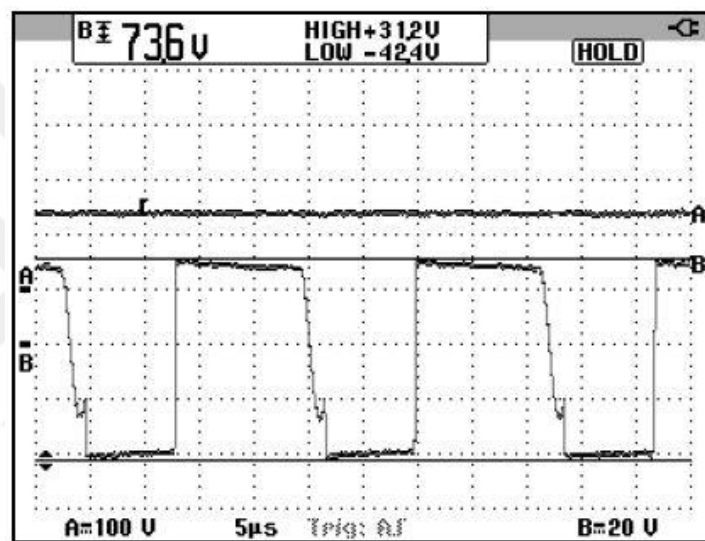


Figure 43. Prototype Flyback Converter Transformer Secondary Voltage at 100% Load (1.5A) at Maximum (137.5VDC) Input Voltage

Fig. 44 shows the tertiary winding voltage of the transformer at 100% load when 77VDC minimum input voltage is applied. In Fig. 44, symbol A indicates the 77VDC input voltage value and symbol B indicates the tertiary winding voltage. Fig. 44 shows the tertiary winding voltage of the transformer has a maximum peak value of 24VDC, minimum peak value of -12.8VDC, peak to peak 36.8VDC, 12.5VRMS and a wave period of 34.60 μ s when 77VDC minimum input voltage is applied to the flyback converter at 100% load,

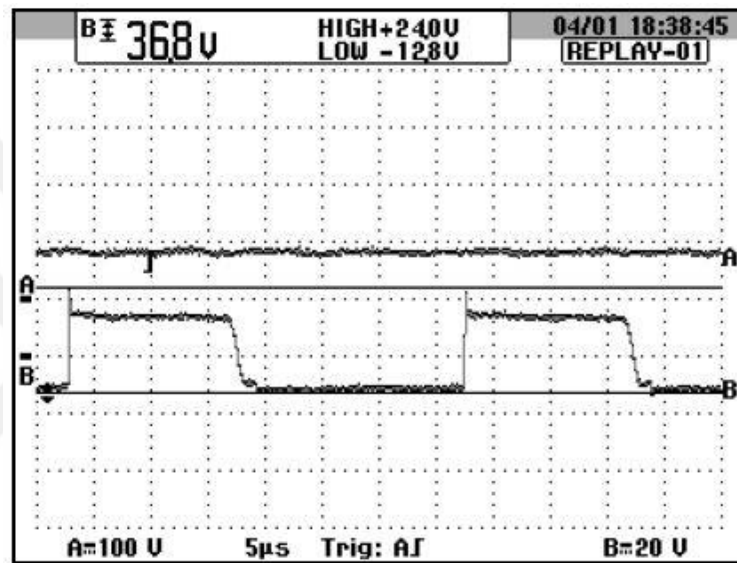


Figure 44. Transformer Tertiary Winding Voltage at 100% Load (1.5A) at Prototype Flyback Converter Minimum (77VDC) Input Voltage

Fig. 45 shows the tertiary winding voltage of the transformer at 100% load when 110VDC nominal input voltage is applied. In Fig. 45, symbol A indicates 110VDC input voltage value and symbol B indicates tertiary winding voltage. Fig. 45 shows that the tertiary winding voltage of the transformer has a maximum peak value of 14VDC, minimum peak value of -18.8VDC, peak to peak 32.8VDC, 12.1VRMS and a wave period of 25 μ s when 110VDC nominal input voltage is applied to the flyback converter at 100% load.

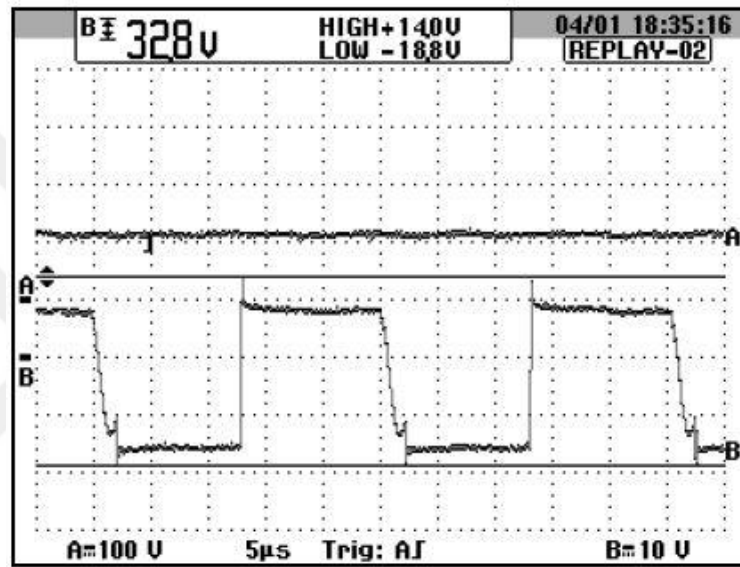


Figure 45. Transformer Tertiary Winding Voltage at 100% Load (1.5A) at Prototype Flyback Converter Nominal (110VDC) Input Voltage

Fig. 46 shows the tertiary winding voltage of the transformer at 100% load when a maximum input voltage of 137.5VDC is applied. In Fig. 46, symbol A indicates 137.5VDC input voltage value and symbol B indicates tertiary winding voltage. Fig. 46 shows that the tertiary winding voltage of the transformer has a maximum peak value of 23.2VDC, minimum peak value of -29.6VDC, peak to peak 52.8VDC, 16.4VRMS and a wave period of 21.40 μ s when a maximum input voltage of 137.5VDC is applied to the flyback converter at 100% load.

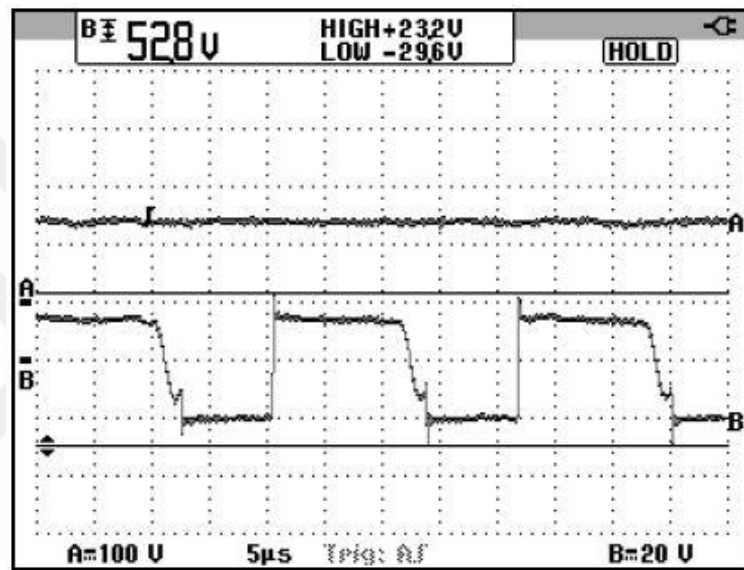


Figure 46. Transformer Tertiary Winding Voltage at 100% Load (1.5A) at Prototype Flyback Converter Maximum (137.5VDC) Input Voltage

4.7 Voltage in Secondary and Tertiary Winding of Transformer at 20% Load (300mA)

The voltages on the secondary and tertiary windings of the transformer at 20% load at 77VDC, 110VDC and 137.5VDC input voltages are displayed on the oscilloscope.

Fig. 47 shows the secondary voltage of the transformer at 20% load when 77VDC minimum input voltage is applied. In Fig. 47, symbol A indicates the 77VDC input voltage value and symbol B indicates the secondary voltage. Fig. 47 analysed, the secondary voltage of the transformer has a maximum peak value of 30.4VDC, minimum peak value of -24VDC, peak to peak 54.4VDC, 21.6VRMS and a wave period of 20.60 μ s when 77VDC minimum input voltage is applied to the flyback converter at 20% load.

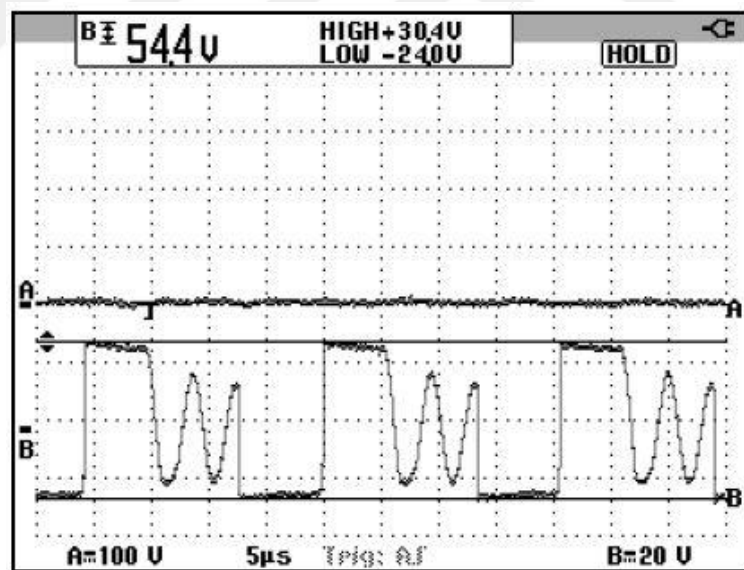


Figure 47. Prototype Flyback Converter Transformer Secondary Voltage at 20% Load (300mA) at Minimum (77VDC) Input Voltage

Fig. 48 shows the secondary voltage of the transformer at 20% load when 110VDC nominal input voltage is applied. In Fig. 48, symbol A indicates the 110VDC input voltage value and symbol B indicates the secondary voltage. Fig. 48 analysed, the secondary voltage of the transformer has a maximum peak value of 31.2VDC, minimum peak value of -33.6VDC, peak to peak 64.8VDC, 24VRMS and a wave period of 20.60 μ s when 77VDC minimum input voltage is applied to the flyback converter at 20% load.

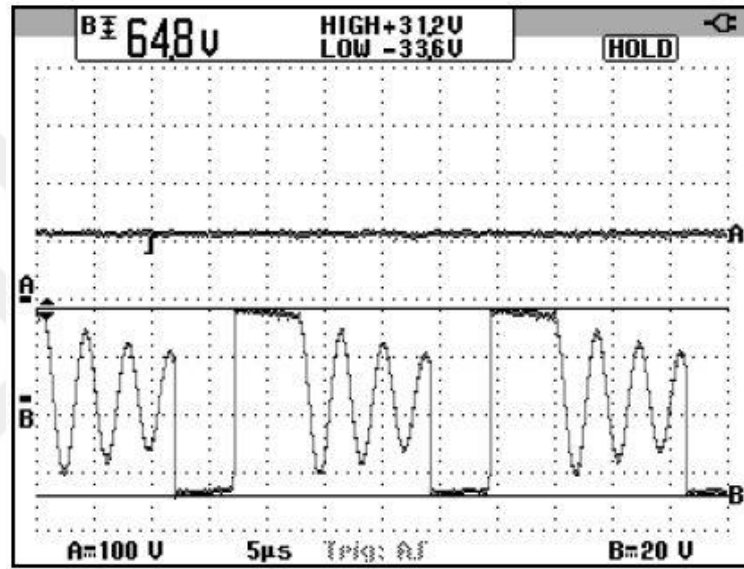


Figure 48. Transformer Secondary Voltage at 20% Load (300mA) at Prototype Flyback Converter Nominal (110VDC) Input Voltage

Fig. 49 shows the secondary voltage of the transformer at 20% load when a maximum input voltage of 137.5VDC is applied. In Fig. 49, symbol A indicates the 137.5VDC input voltage value and symbol B indicates the secondary voltage. Fig. 49 is analysed, when 137.5VDC minimum input voltage is applied to the flyback converter at 20% load, the secondary voltage of the transformer has a maximum peak value of 30.4VDC, minimum peak value of -41.6VDC, peak to peak 72VDC, 24.4VRMS and a wave period of 23.60 μ s.

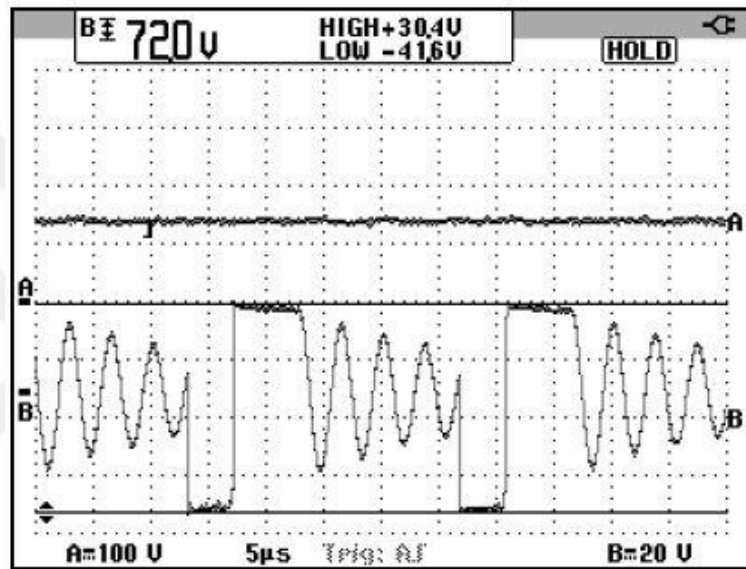


Figure 49. Transformer Secondary Voltage at 20% Load (300mA) at Maximum (137.5VDC) Input Voltage of Prototype Flyback Converter

Fig. 50 shows the tertiary winding voltage of the transformer at 20% load when 77VDC minimum input voltage is applied. In Fig. 50, symbol A indicates the 77VDC input voltage value and symbol B indicates the tertiary winding voltage. Fig. 50 analysed, the tertiary winding voltage of the transformer has a maximum peak value of 16VDC, minimum peak value of -24VDC, peak-to-peak 40VDC, 11VRMS and a wave period of 21.20 μ s when 77VDC minimum input voltage is applied to the flyback converter at 20% load.

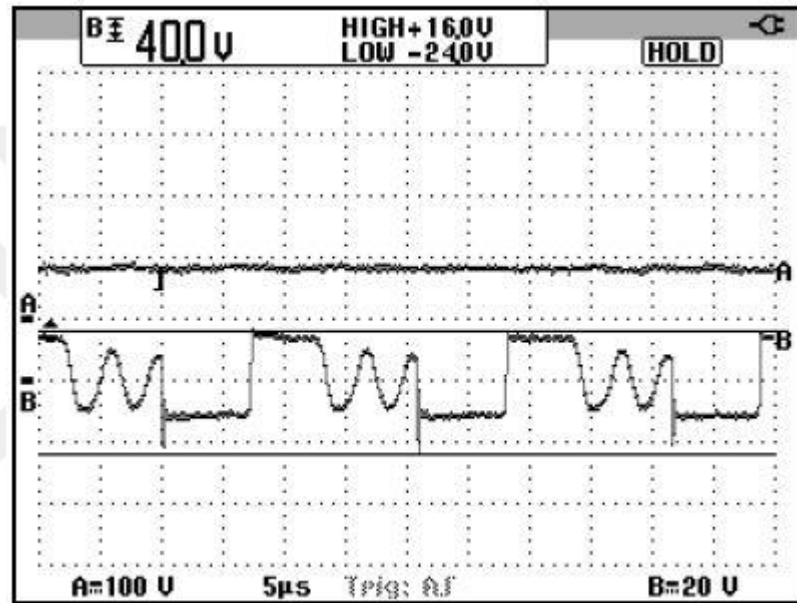


Figure 50. Prototype Flyback Converter Tertiary Winding Voltage of Transformer at 20% Load (300mA) at Minimum (77VDC) Input Voltage

Fig. 51 shows the tertiary winding voltage of the transformer at 20% load when 110VDC nominal input voltage is applied. In Fig. 51, symbol A indicates 110VDC input voltage value and symbol B indicates tertiary winding voltage. Fig. 51 analysed, the tertiary winding voltage of the transformer has a maximum peak value of 17.6VDC, minimum peak value of -33.6VDC, peak to peak 51.2VDC, 12VRMS and a wave period of 22.20 μ s when 110VDC nominal input voltage is applied to the flyback converter at 20% load.

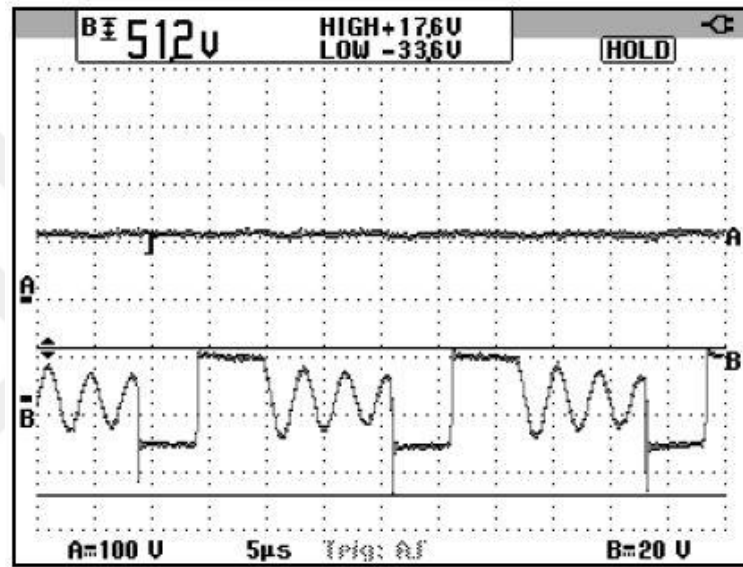


Figure 51. Tertiary Winding Voltage of Transformer at 20% Load (300mA) at Nominal (110VDC) Input Voltage of Prototype Flyback Converter

Fig. 52 shows the tertiary winding voltage of the transformer at 20% load when a maximum input voltage of 137.5VDC is applied. In Fig. 52, symbol A indicates 137.5VDC input voltage value and symbol B indicates tertiary winding voltage. Fig. 52 analysed, the tertiary winding voltage of the transformer has a maximum peak value of 18.4VDC, minimum peak value of -35.2VDC, peak to peak 53.6VDC, 12.1VRMS and a wave period of 23.60 μ s when a maximum input voltage of 137.5VDC is applied to the flyback converter at 20% load.

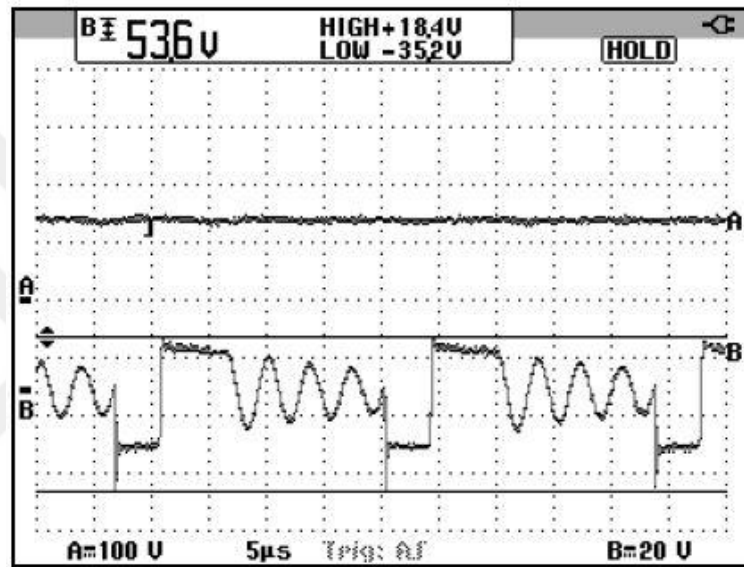


Figure 52. Tertiary Winding Voltage of Transformer at 20% Load (300mA) at Maximum (137.5VDC) Input Voltage of Prototype Flyback Converter

4.8 PWM Signal at 100% Load (1.5A)

The PWM signal receives feedback according to the input voltage value and sends it to the Drain end of the MOSFET from the Gate pin of the LT8316 integration. The MOSFET switches on and off according to the fill rate of the signal coming to the Drain terminal. The PWM signal generated at the GATE terminal of the LT8316 integration at 100% load at 77VDC, 110VDC and 137.5VDC input voltages is displayed with an oscilloscope.

Fig. 53 shows the PWM signal at 100% load when 77VDC minimum input voltage is applied and symbol A indicates the 77VDC input voltage value and symbol B indicates the PWM signal.

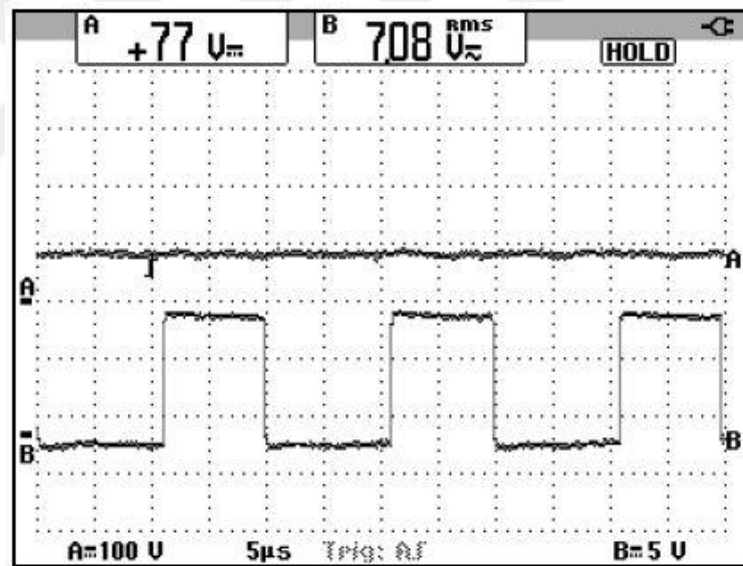


Figure 53. Prototype Flyback Converter PWM Signal at 100% Load (1.5A) at Minimum (77VDC) Input Voltage

Fig. 54 shows the PWM signal at 100% load when 110VDC nominal input voltage is applied and symbol A indicates the 110VDC input voltage value and symbol B indicates the PWM signal.

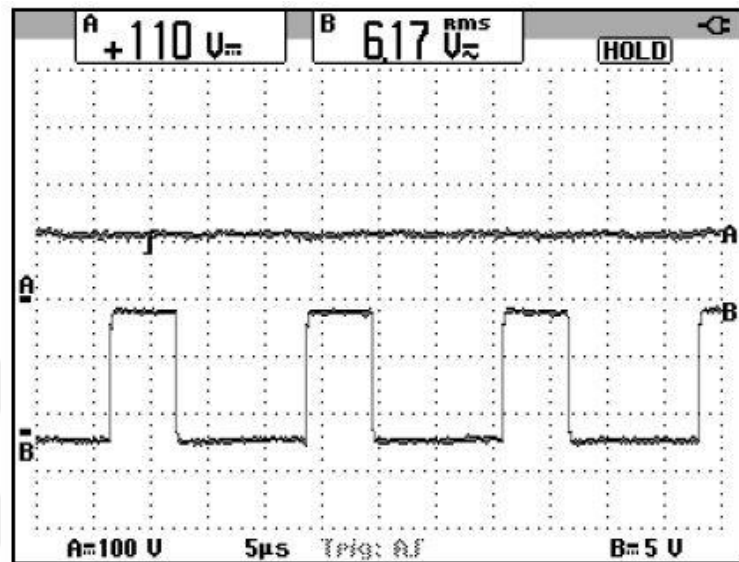


Figure 54. Prototype Flyback Converter PWM Signal at 100% Load (1.5A) at Nominal (110VDC) Input Voltage

Fig. 55 shows the PWM signal at 100% load when 137.5VDC maximum input voltage is applied and symbol A indicates 137.5VDC input voltage value and symbol B indicates PWM signal.

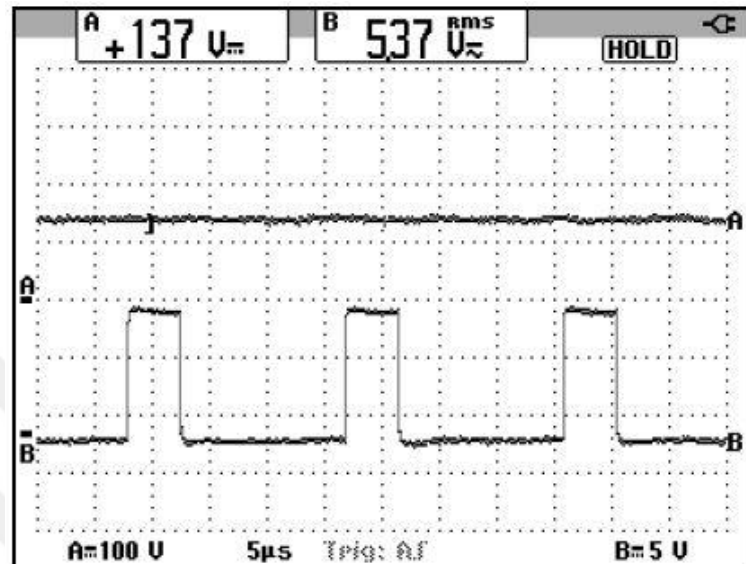


Figure 55. Prototype Flyback Converter PWM Signal at 100% Load (1.5A) at Maximum (137.5VDC) Input Voltage

4.9 PWM Signal at 20% Load (300mA)

The PWM signal generated at the GATE end of the LT8316 integration at 20% load at 77VDC, 110VDC and 137.5VDC input voltages is displayed with an oscilloscope.

Fig. 56 shows the PWM signal at 20% load when 77VDC minimum input voltage is applied and the symbol A indicates the 77VDC input voltage value and the symbol B indicates the PWM signal.

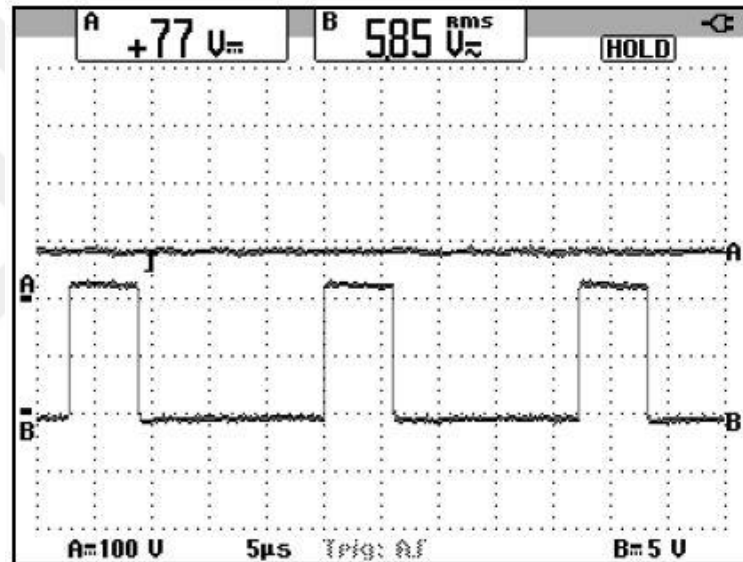


Figure 56. Prototype Flyback Converter PWM Signal at 20% Load (300mA) at Minimum (77VDC) Input Voltage

Fig. 57 shows the PWM signal at 20% load when 110VDC nominal input voltage is applied and symbol A indicates the 110VDC input voltage value and symbol B indicates the PWM signal.

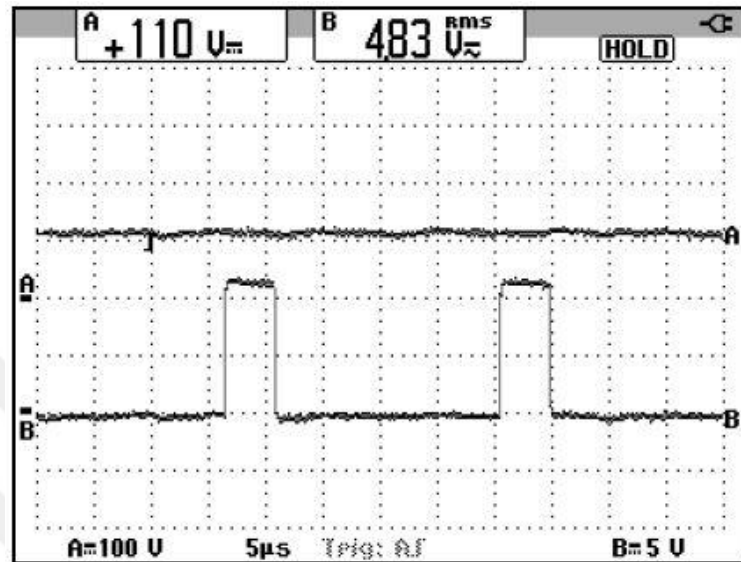


Figure 57. Prototype Flyback Converter PWM Signal at 20% Load (300mA) at Nominal (110VDC) Input Voltage

Fig. 58 shows the PWM signal at 20% load when a maximum input voltage of 137.5VDC is applied and the symbol A indicates the 137.5VDC input voltage value and the symbol B indicates the PWM signal.

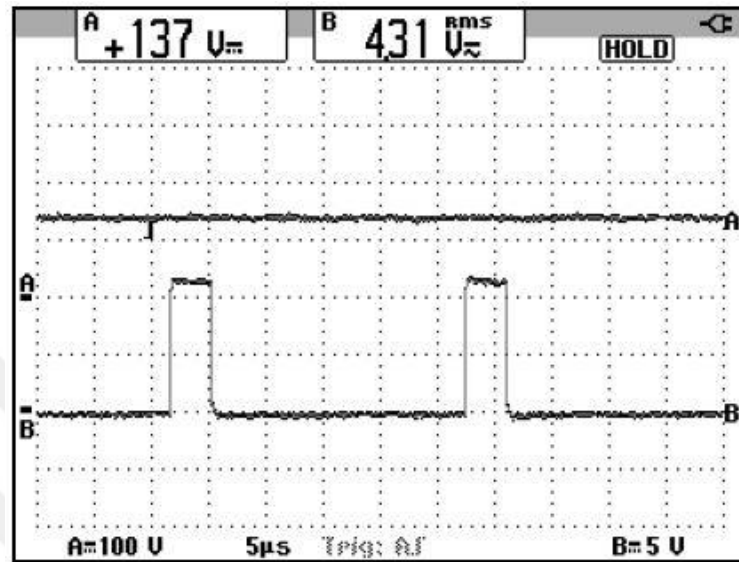


Figure 58. Prototype Flyback Converter PWM Signal at 20% Load (300mA) at Maximum (137.5VDC) Input Voltage

4.10 MOSFET Voltage, Current and Snubber

The drain pin of the MOSFET, the switching element of the designed flyback converter, switches by sending a PWM signal to the primary end of the flyback transformer. In addition, an input voltage signal is applied to the primary of the transformer. The input voltage signal and PWM signal are collected on the drain pin of the MOSFET. The PWM signal and the input voltage signal collected at the drain pin of the MOSFET generate ringing when the MOSFET is switched on and off. A snubber circuit was designed to dampen the ringing and applied to the prototype circuit. When the MOSFET is switched on, the current flowing through it passes through the drain-source region. The current on the MOSFET was found by the voltage falling on $40\text{m}\Omega$ connected to the source pin of the MOSFET using ohm's law. The designed converter is shown with oscilloscope at 77VDC, 110VDC and 137.5VDC input voltages without snubber, with snubber and currents on MOSFET at 100% load. At 100% load, the ringing is at maximum level and if the ringing at 100% load is damped, the ringing at 20% load is improved. For this reason, currents at 20% load are not shown.

4.10.1 MOSFET voltage, current and snubber at 100% load (1.5A). Fig. 59 shows the voltage signal without adding snubber design on the MOSFET at 100% load when 77VDC minimum input voltage is applied and in Fig. 59 symbol A indicates the 77VDC input voltage value and symbol B indicates the voltage signal without snubbers. Fig. 59 is analysed, the maximum peak value of the voltage on the MOSFET without snubber design is 88VDC, the minimum peak value is -364VDC, the peak-to-peak value is 452VDC, the period of a wave formed by ringing is 34.80 μ s and the ringing period is 330ns when 77VDC minimum input voltage is applied to the flyback converter at 100% load.

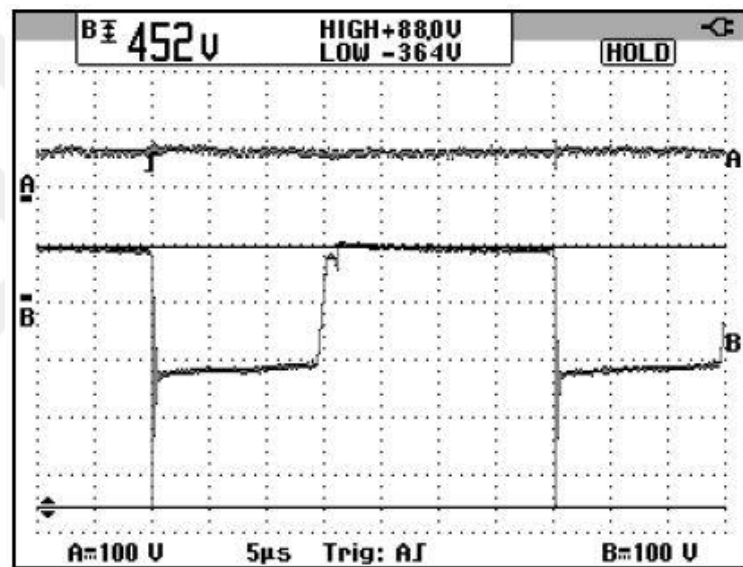


Figure 59. Prototype Flyback Converter MOSFET Voltage without Snubbers at 100% Load (1.5A) at Minimum (77VDC) Input Voltage

Fig. 61 shows the voltage signal by adding the snubber design on the MOSFET at 100% load when 77VDC minimum input voltage is applied and in Fig. 61 the symbol A indicates the 77VDC input voltage value and the symbol B indicates the voltage signal generated by the snubber ringing. The maximum peak value of the voltage on the MOSFET is 108VDC, the minimum peak value is -168VDC, the peak-to-peak value is 276VDC, the wave period with ringing is 21.60 μ s and the ringing period is 150ns when 77VDC minimum input voltage is applied to the flyback converter with snubber design at 100% load. When the snubber design is added, the wavelength decreases from 34.80 μ s to 21.60 μ s, the ringing period decreases from 330ns to 150ns, the ringing is damped and the total voltage on the MOSFET decreases from 452VDC to approximately half of 276VDC.

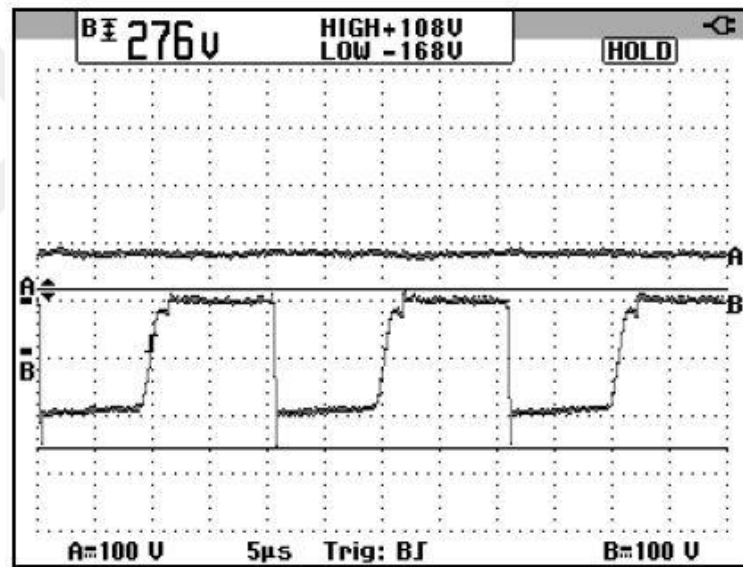


Figure 60. Prototype Flyback Converter Minimum (77VDC) Input Voltage at 100% Load (1.5A) MOSFET Voltage with Snubber Design

Fig. 61 shows the signal passing through the drain-source section when the MOSFET is switched on at a minimum input voltage of 77VDC. As mentioned before, due to the lack of a current probe, the current value of 2.05A was found by calculating the signal on the 40mΩ resistor connected to the source pin of the MOSFET using ohm's law in Eq. 4.1. In Fig. 61, peak currents occurred at a time less than 10ns. Since the duration value is very small, only positive directional currents were calculated by ignoring it.

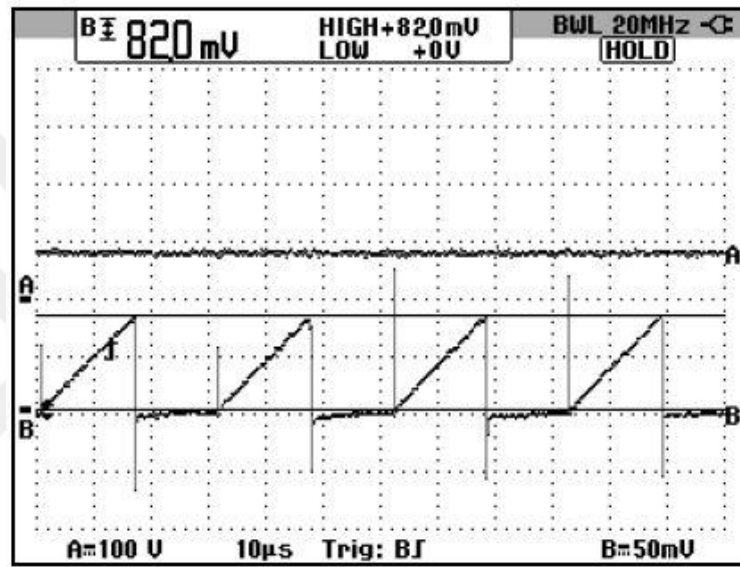


Figure 61. Prototype Flyback Converter Voltage Over 40mΩ at 100% Load (1.5A) at Minimum (77VDC) Input Voltage

$$I_{DS} = V/I = \frac{82 * 10^{-3}}{40 * 10^{-3}} = 2.05A \quad 4.1$$

Fig. 61 shows the voltage signal without adding snubber design on the MOSFET at 100% load when 110VDC nominal input voltage is applied and in Fig. 62, symbol A indicates the 110VDC input voltage value and symbol B indicates the voltage signal without snubbers. Fig. 61 is analysed, the maximum peak value of the voltage on the MOSFET without snubber design is 132VDC, the minimum peak value is -304VDC, the peak-to-peak is 436VDC, a wave period of 24.90 μ s and the ringing period is 250ns when 110VDC nominal input voltage is applied to the flyback converter at 100% load.

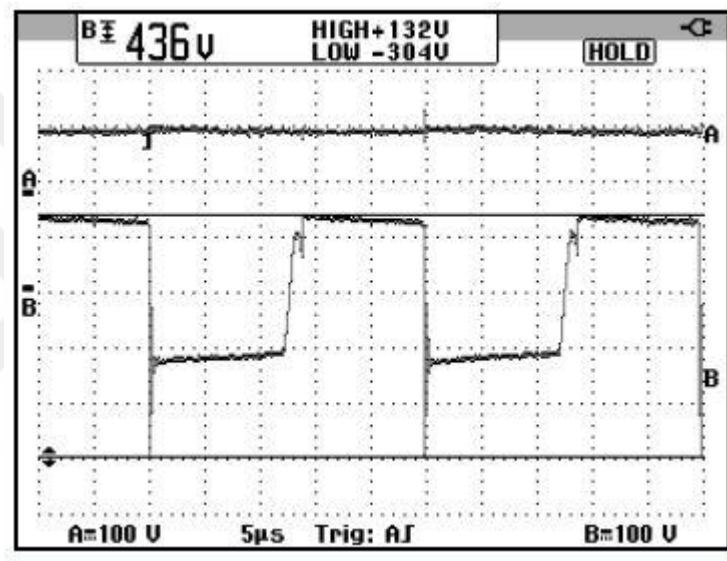


Figure 62. Prototype Flyback Converter MOSFET Voltage without Snubbers at 100% Load (1.5A) at Nominal (110VDC) Input Voltage

Fig. 63 shows the voltage signal by adding snubber design on the MOSFET at 100% load when 110VDC nominal input voltage is applied and in Fig. 63, symbol A indicates the 110VDC input voltage value and symbol B indicates the voltage signal generated by snubberised ringing. Fig. 63 analysed, the maximum peak value of the voltage on the MOSFET is 324VDC, the minimum peak value is -28VDC, the peak-to-peak voltage is 352VDC, the wave period is 14.80 μ s and the ringing period is 180ns when 110VDC nominal input voltage is applied to the flyback converter with snubber design at 100% load. When the snubber design was added, the wavelength decreased from 24.90 μ s to 14.80 μ s, the resonance decreased from 250ns to 180ns, the resonance was damped and the total voltage on the MOSFET decreased from 436VDC to 352VDC.

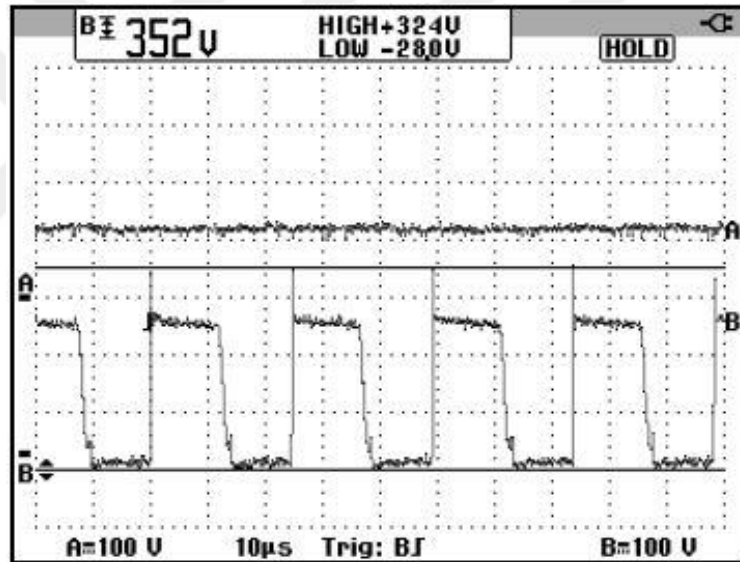


Figure 63. Prototype Flyback Converter MOSFET Voltage with Snubber Design at 100% Load (1.5A) at Nominal (110VDC) Input Voltage

Fig. 64 shows the signal passing through the drain-source section when the MOSFET is switched on at 110VDC nominal input voltage. The current value was found to be 1.75A by calculating the signal on the 40mΩ resistor using ohm's law in Eq. 4.2. In Fig. 64, peak currents occur in less than 10ns. Since the time value is very small, only positive directional currents are calculated by ignoring it.

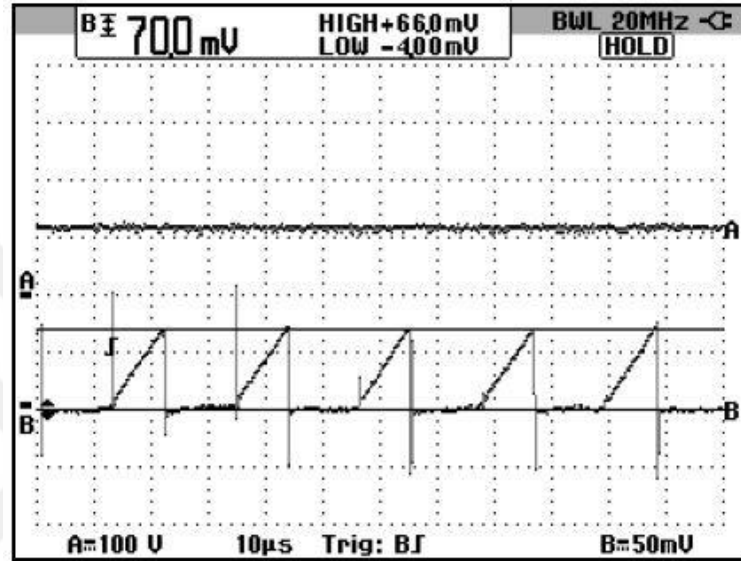


Figure 64. Prototype Flyback Converter Nominal (110VDC) Input Voltage at 100% Load (1.5A) Over 40mΩ

$$I_{DS} = V/R = \frac{70 * 10^{-3}}{40 * 10^{-3}} = 1.75A \quad 4.2$$

Fig. 65 shows the voltage signal without adding the snubber design on the MOSFET at 100% load when the maximum input voltage of 137.5VDC is applied and in Fig. 65, symbol A indicates the 137.5VDC input voltage value and symbol B indicates the voltage signal generated by ringing without snubber. Fig. 65 analysed, the maximum peak value of the voltage on the MOSFET without snubber design is 168VDC, the minimum peak value is -292VDC, the peak-to-peak value is 460VDC, the period of a wave formed by ringing is 20.40 μ s and the ringing period is 330ns when a maximum input voltage of 137.5VDC is applied to the flyback converter at 100% load.

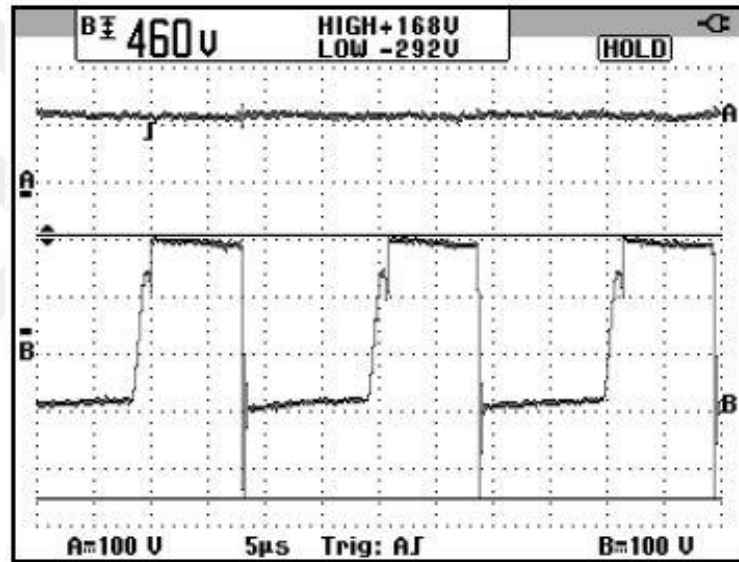


Figure 65. Prototype Flyback Converter MOSFET Voltage without Snubbers at 100% Load (1.5A) at Maximum (137.5VDC) Input Voltage

Fig. 66 shows the voltage signal by adding snubber design on the MOSFET at 100% load when 137.5VDC nominal input voltage is applied. In Fig. 66, symbol A indicates the 137.5VDC input voltage value and symbol B indicates the voltage signal generated by snubberised ringing. Fig. 66 analysed, the maximum peak value of the voltage on the MOSFET is 172VDC, the minimum peak value is -228VDC, the peak-to-peak is 400VDC, a wave period of 22 μ s and a ringing period of 260ns when 110VDC nominal input voltage is applied to the flyback converter with snubber design at 100% load. When the snubber design is added, the wavelength increases from 24.90 μ s to 14.80 μ s. The ringing is reduced from 250ns to 180ns and the total voltage across the MOSFET is reduced from 460VDC to 400VDC.

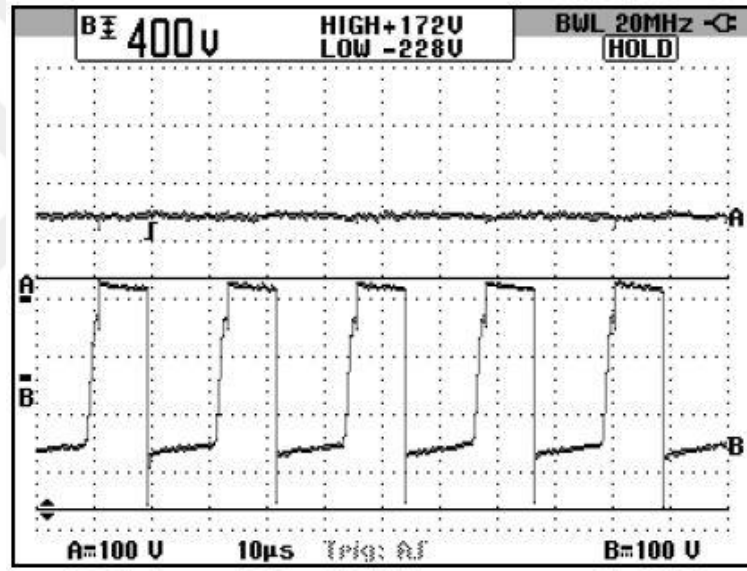


Figure 66. Prototype Flyback Converter MOSFET Voltage with Snubber Design at 100% Load (1.5A) at Maximum (137.5VDC) Input Voltage

Fig. 67 shows the signal passing through the drain-source section when the MOSFET is switched on at a maximum input voltage of 137.5VDC. The current value was found to be 1.6A by calculating the signal on the 40mΩ resistor using ohm's law in Eq. 4.3. In Fig. 67, peak currents occur at times less than 10ns. Since the time value is very small, only positive directional currents are calculated by ignoring it.

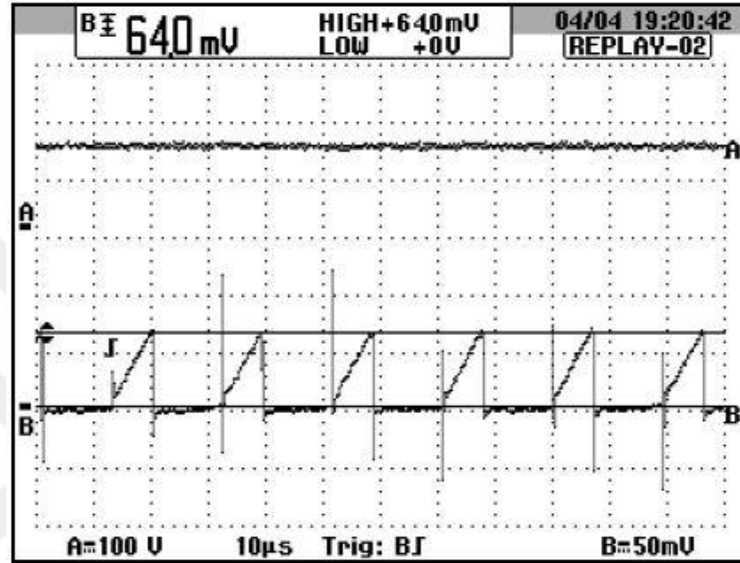


Figure 67. Prototype Flyback Converter Maximum (137.5VDC) Input Voltage at 100% Load (1.5A) Voltage Over 40mΩ

$$I_{DS} = V/R = \frac{64 * 10^{-3}}{40 * 10^{-3}} = 1.6A \quad 4.3$$

4.10.2 MOSFET current at 20% load (300mA). Fig. 68 shows the signal passing through the drain-source section when the MOSFET is switched on at 77VDC minimum input voltage and 20% load. The current value of the MOSFET at low load was calculated in Eq. 4.4 using ohm's law on the 40mΩ resistor connected to the source pin of the MOSFET and the current value was found to be 1A. In Fig. 68, peak currents occur at a time less than 10ns. Since the duration value is very small, only positive directional currents are calculated by ignoring it.

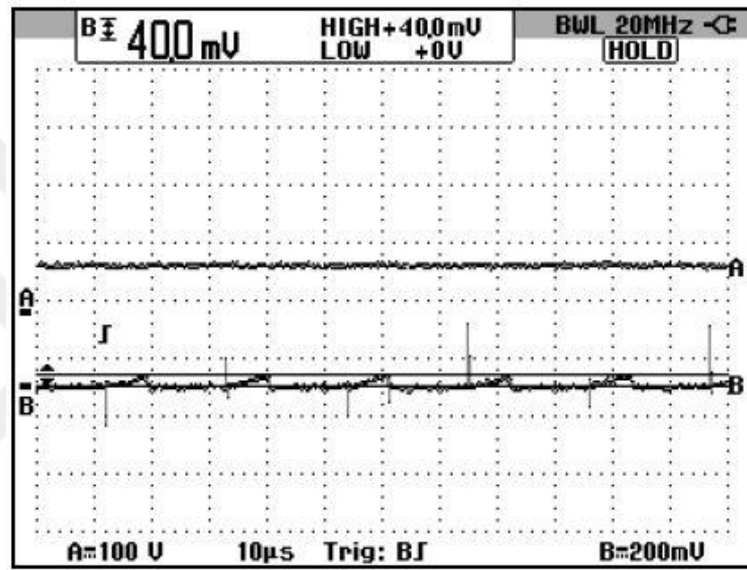


Figure 68. Prototype Flyback Converter Voltage Minimum (77VDC) Input Voltage at 20% Load (300mA) Voltage Over 40mΩ

$$I_{DS} = V/R = \frac{40 * 10^{-3}}{40 * 10^{-3}} = 1A \quad 4.4$$

Fig. 69 shows the signal passing through the drain-source section when the MOSFET is switched on at 110VDC nominal input voltage and 100% load. The current value was found to be 1.2A by calculating the signal across a 40mΩ resistor using ohm's law in Eq. 4.5. In Fig. 69, peak currents occur at times less than 10ns. Since the time value is very small, only positive directional currents are calculated by ignoring it.

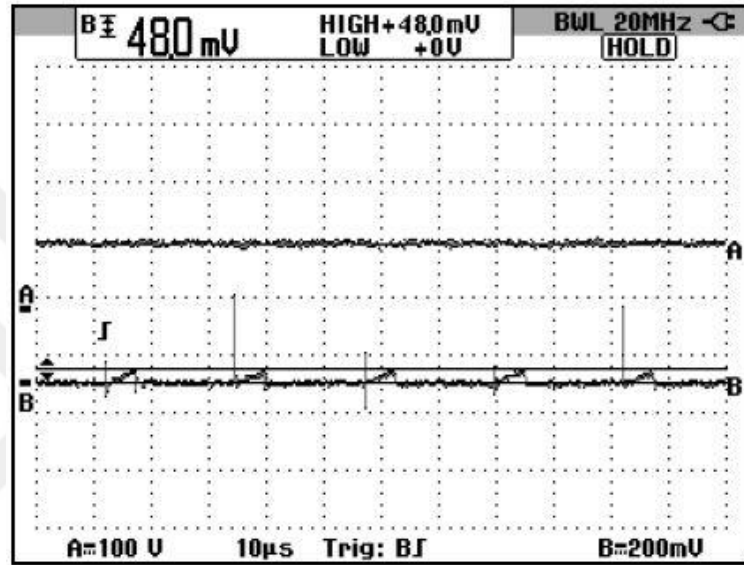


Figure 69. Prototype Flyback Converter at Nominal (110VDC) Input Voltage at 20% Load (300mA) Voltage over 40mΩ

$$I_{DS} = V/R = \frac{48 * 10^{-3}}{40 * 10^{-3}} = 1.2A \quad 4.5$$

Fig. 70 shows the signal passing through the drain-source section when the MOSFET is switched on at a maximum input voltage of 137.5VDC and 100% load. The current value was found to be 1A by calculating the signal across a 40mΩ resistor using ohm's law in Eq. 4.5. In Fig. 70, peak currents occur at times less than 10ns. Since the duration value is very small, only positive directional currents are calculated by ignoring it.

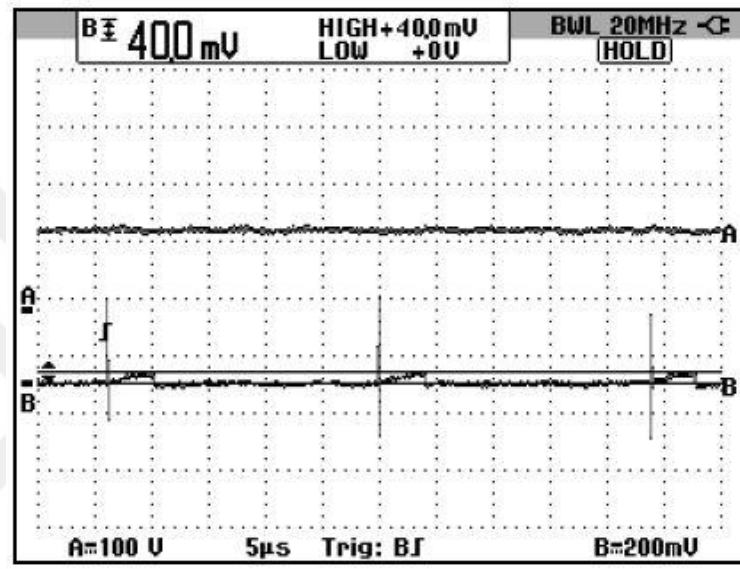


Figure 70. Prototype Flyback Converter Maximum (137.5VDC) Input Voltage at 20% Load (300mA) Voltage over 40mΩ

$$I_{DS} = V/R = \frac{40 * 10^{-3}}{40 * 10^{-3}} = 1A \quad 4.5$$

4.11 Output Diode Voltage and Current

The anode pin of the output diode is connected to the secondary of the transformer. For this reason, the current and voltage in the secondary of the transformer pass through the output diode. The current on the output diode was found with the voltage falling on $40\text{m}\Omega$ using ohm's law. The voltage and current signals on the output diode at 77VDC, 110VDC and 137.5VDC input voltages at 20% and 100% load are shown with an oscilloscope.



4.11.1 Output diode voltage and current at 100% load (1.5A). Fig. 71 shows the voltage signal on the output diode at 100% load at 77VDC minimum input voltage. In Fig. 71, symbol A indicates the 77VDC input voltage value and symbol B indicates the output diode voltage. Fig. 71 analysed, the secondary voltage of the transformer, in other words the voltage on the output diode, has a maximum peak value of 5.60VDC, minimum peak value of -44.8VDC, peak to peak 50.4VDC, 30.5VRMS and a wave period of 20 μ s when 77VDC minimum input voltage is applied to the flyback converter at 100% load.

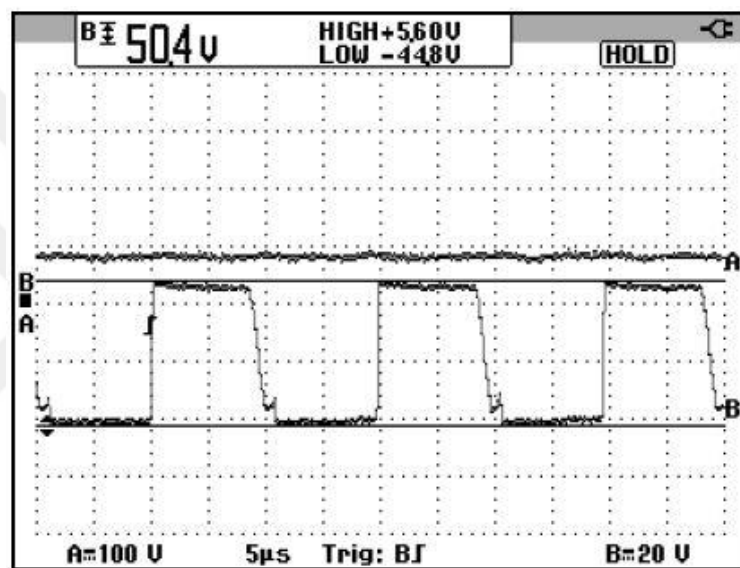


Figure 71. Prototype Flyback Converter Output Diode Voltage at 100% Load (1.5A) at Minimum (77VDC) Input Voltage

Fig. 72 shows the voltage signal on the 40mΩ resistor connected to the secondary of the transformer at 100% load at 77VDC minimum input voltage. In Eq. 4.6, the current on the output diode is calculated as 7.6A. In Fig. 72, symbol A indicates the 77VDC input voltage value and symbol B indicates the output diode voltage.

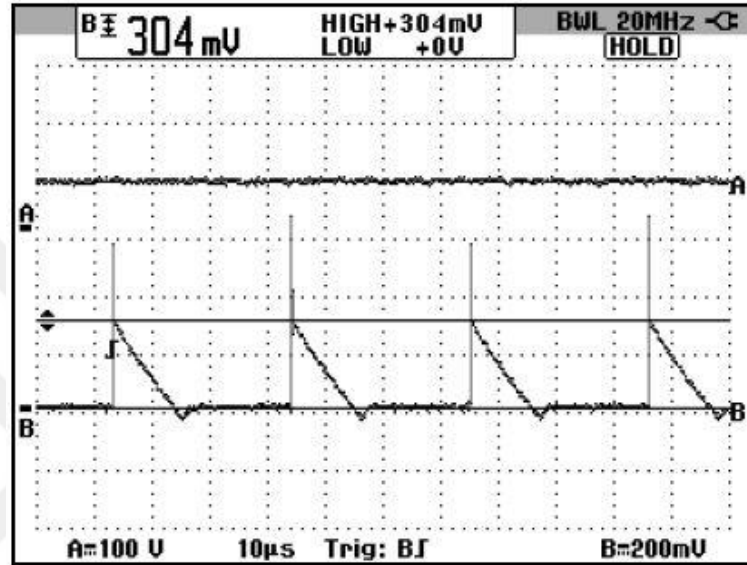


Figure 72. Prototype Flyback Converter Voltage Minimum (77VDC) Input Voltage at 100% Load (1.5A) Voltage Over 40mΩ

$$I_D = \frac{V}{R} = \frac{304 * 10^{-3}}{40 * 10^{-3}} = 7.6A \quad 4.6$$

Fig. 73 shows the voltage signal on the output diode at 100% load at 110VDC nominal input voltage. In Fig. 73, symbol A indicates the 110VDC input voltage value and symbol B indicates the output diode voltage. Fig. 73 is analysed, the secondary voltage of the transformer, in other words, the voltage across the output diode has a maximum peak value of 5.60VDC, minimum peak value of -56.8VDC, peak to peak 62.4VDC, 33.10VRMS and a wave period of 17.20 μ s when 110VDC minimum input voltage is applied to the flyback converter at 100% load.

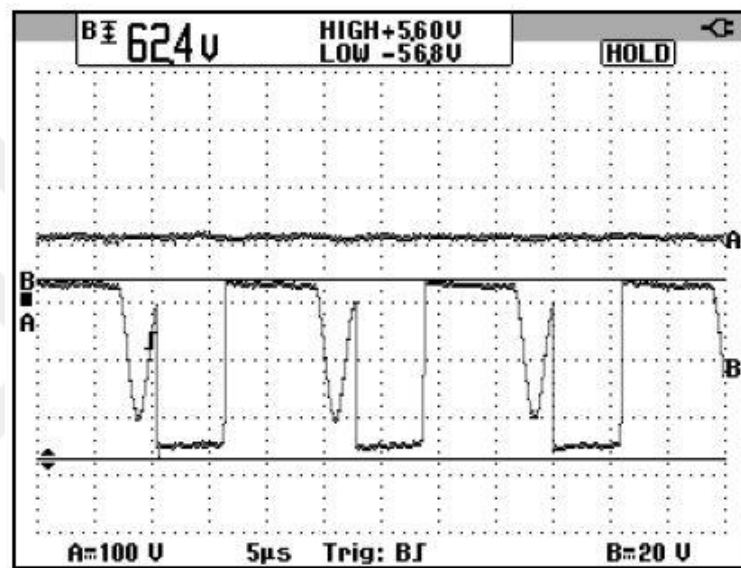


Figure 73. Prototype Flyback Converter Output Diode Voltage at 100% Load (1.5A) at Nominal (110VDC) Input Voltage

Fig. 74 shows the voltage signal on the 40mΩ resistor connected to the secondary of the transformer at 100% load at 110VDC nominal input voltage. In Eq. 4.7, the current on the output diode is calculated as 6A. In Fig. 74, symbol A indicates the 110VDC input voltage value and symbol B indicates the output diode voltage.

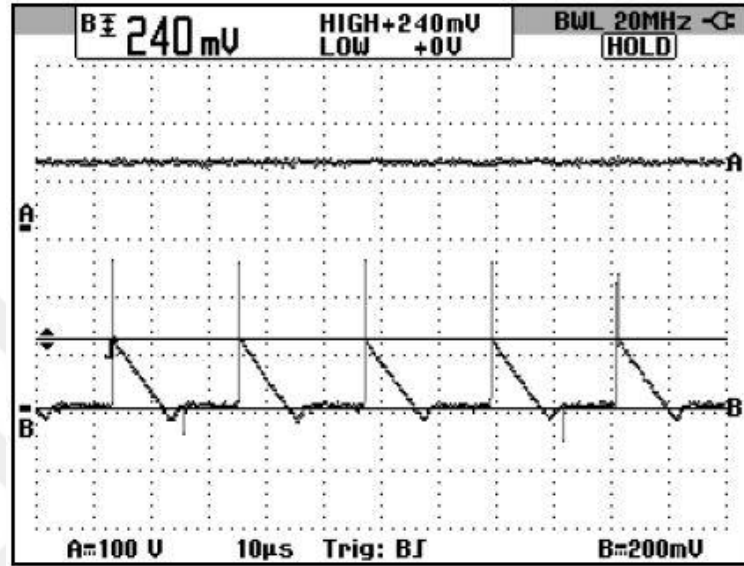


Figure 74. Prototype Flyback Converter Voltage Nominal (110VDC) Input Voltage at 100% Load (1.5A) Voltage Over 40mΩ

$$I_D = \frac{V}{R} = \frac{240 * 10^{-3}}{40 * 10^{-3}} = 6A \quad 4.7$$

Fig. 75 shows the voltage signal on the output diode at 100% load at 137.5VDC maximum input voltage. In Fig. 75, symbol A indicates the 137.5VDC input voltage value and symbol B indicates the output diode voltage. Fig. 75 is analysed, the secondary voltage of the transformer, in other words the voltage on the output diode, has a maximum peak value of 5.60VDC, minimum peak value of -61.6VDC, peak to peak 67.2VDC, 35.10VRMS and a wave period of 18.80 μ s when 137.5VDC maximum input voltage is applied to the flyback converter at 100% load.

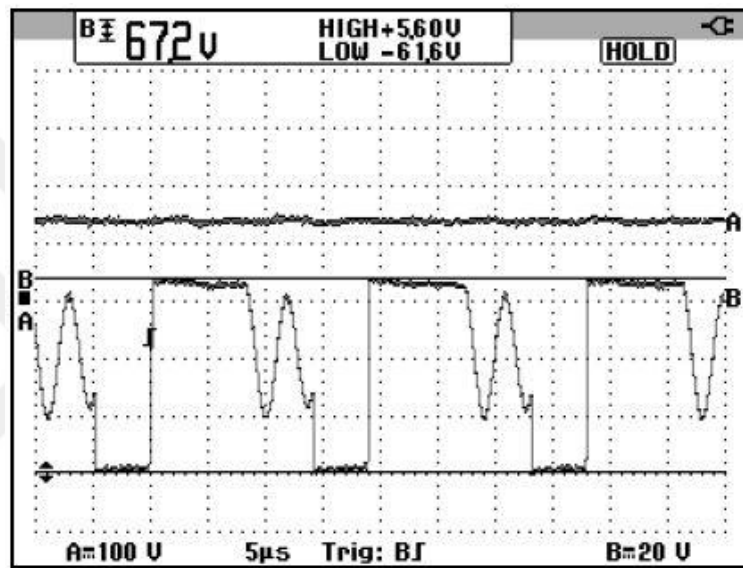


Figure 75. Prototype Flyback Converter Output Diode Voltage at 100% Load (1.5A) at Maximum (137.5VDC) Input Voltage

Fig. 76 shows the voltage signal across the 40mΩ resistor connected to the secondary of the transformer at 100% load at a maximum input voltage of 137.5VDC. In Eq. 4.8, the current on the output diode is calculated as 5A. In Fig. 76, symbol A indicates the 137.5VDC input voltage value and symbol B indicates the output diode voltage.

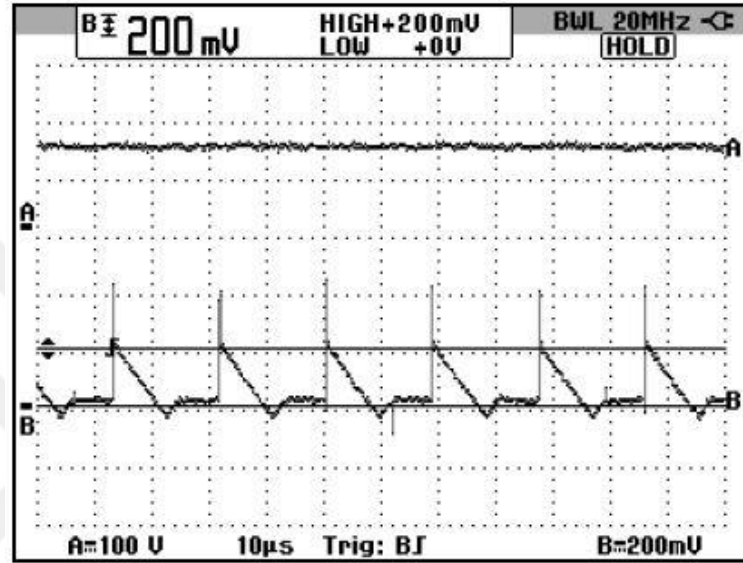


Figure 76. Prototype Flyback Converter Voltage Maximum (137.5VDC) Input Voltage at 100% Load (1.5A) Voltage Over 40mΩ

$$I_D = \frac{V}{R} = \frac{200 * 10^{-3}}{40 * 10^{-3}} = 5A \quad 4.8$$

4.11.2 Output diode voltage and current at 20% load (300mA). Fig. 77 shows the voltage signal on the output diode at 20% load at 77VDC minimum input voltage. In Fig. 77, symbol A indicates the 77VDC input voltage value and symbol B indicates the output diode voltage. Fig. 77 analysed, the secondary voltage of the transformer, in other words the voltage on the output diode, has a maximum peak value of 4.80VDC, minimum peak value of -48VDC, peak to peak 50.4VDC, 32.9VRMS and a wave period of 20.60 μ s when 77VDC minimum input voltage is applied to the flyback converter at 20% load.

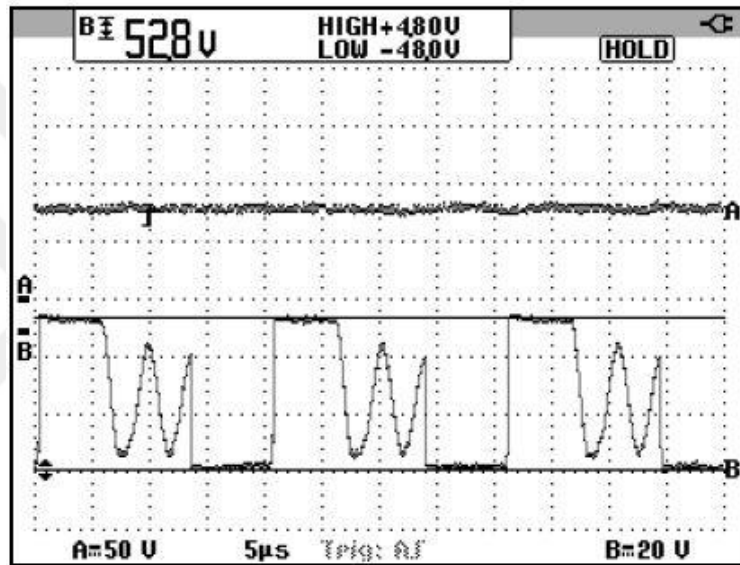


Figure 77. Prototype Flyback Converter Output Diode Voltage at 20% Load (300mA) at Minimum (77VDC) Input Voltage

Fig. 78 shows the voltage signal on the 40mΩ resistor connected to the secondary of the transformer at 20% load at 77VDC minimum input voltage. In Eq. 4.9, the current on the output diode is calculated as 2.4A. In Fig. 78, symbol A indicates the 77VDC input voltage value and symbol B indicates the output diode voltage.

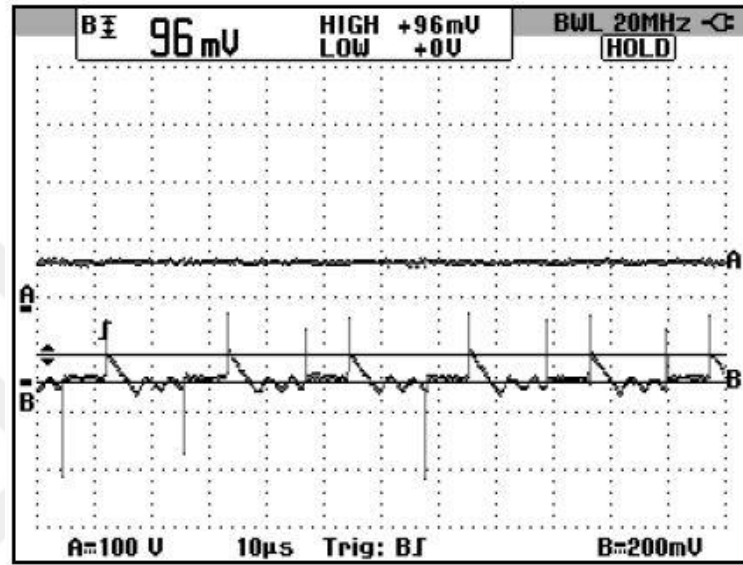
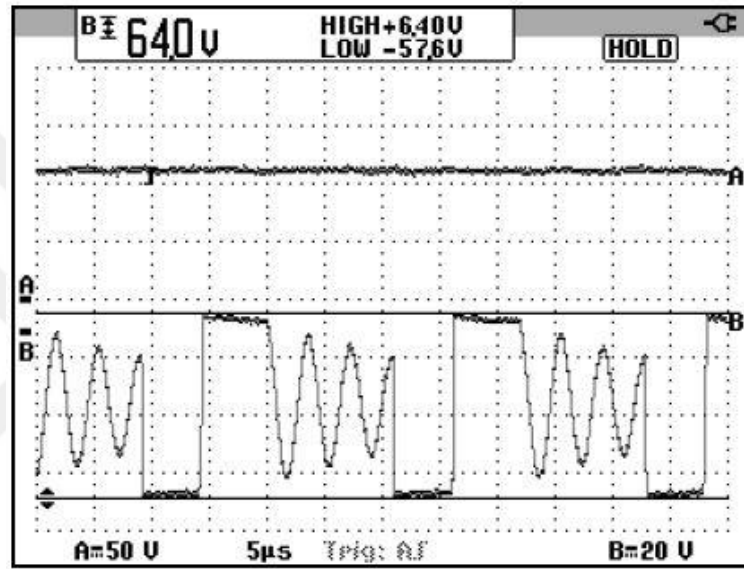


Figure 78. Prototype Flyback Converter Voltage Minimum (77VDC) Input Voltage at 20% Load (300mA) Voltage Over 40mΩ

$$I_D = \frac{V}{R} = \frac{96 * 10^{-3}}{40 * 10^{-3}} = 2.4A \quad 4.9$$

Fig. 79 shows the voltage signal on the output diode at 20% load at 110VDC nominal input voltage. In Fig. 79, symbol A indicates the 110VDC input voltage value and symbol B indicates the output diode voltage. Fig. 79 is analysed, the secondary voltage of the transformer, in other words the voltage on the output diode, has a maximum peak value of 6.40VDC, minimum peak value of -57.6VDC, peak to peak 65VDC, 34.4VRMS and a wave period of 20.60 μ s when 110VDC minimum input voltage is applied to the flyback converter at 20% load.



Şekil 79. Prototype Flyback Converter Output Diode Voltage at 20% Load (300mA) at Nominal (110VDC) Input Voltage

Fig. 80 shows the voltage signal on the 40mΩ resistor connected to the secondary of the transformer at 20% load at 110VDC nominal input voltage. In Eq. 4.10, the current on the output diode is calculated as 2.6A. In Fig. 80, symbol A indicates the 110VDC input voltage value and symbol B indicates the output diode voltage.

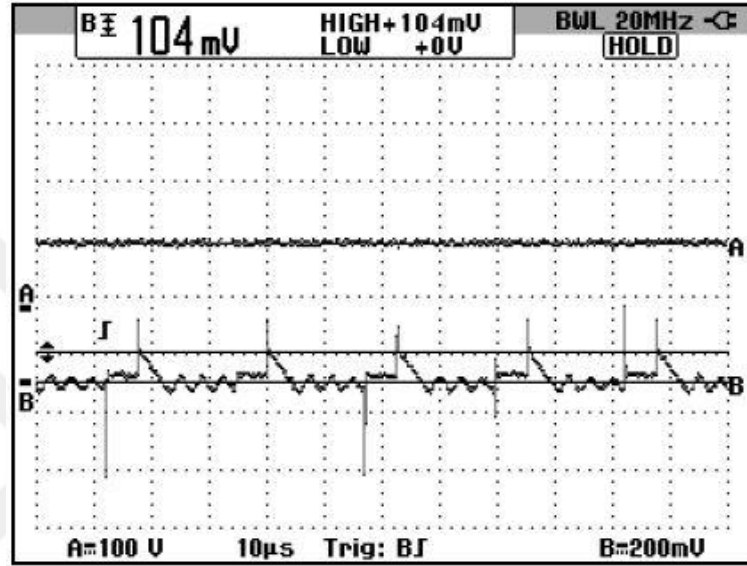
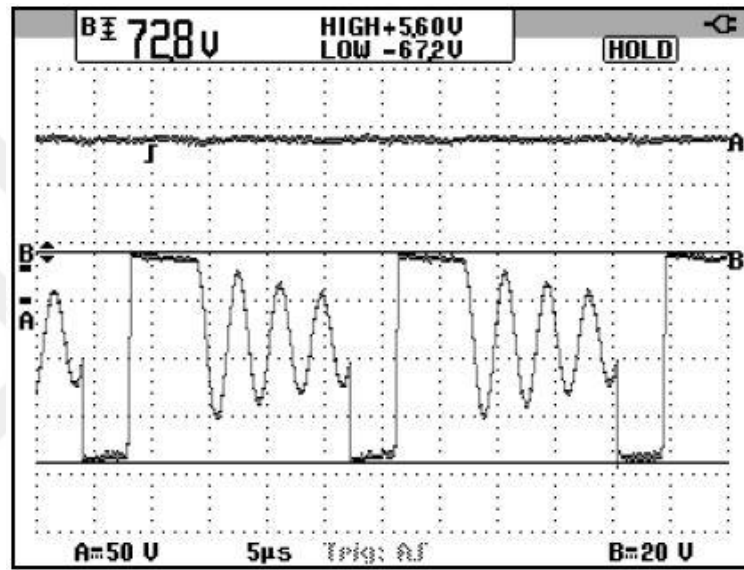


Figure 80. Prototype Flyback Converter Nominal (110VDC) Input Voltage at 20% Load (300mA) Voltage Over 40mΩ

$$I_D = \frac{V}{R} = \frac{104 * 10^{-3}}{40 * 10^{-3}} = 2.6A \quad 4.10$$

Fig. 81 shows the voltage signal on the output diode at 20% load at a maximum input voltage of 137.5VDC. In Fig. 81, symbol A indicates the 137.5VDC input voltage value and symbol B indicates the output diode voltage. Fig. 81 is analysed, the secondary voltage of the transformer, in other words the voltage on the output diode, has a maximum peak value of 6.40VDC, minimum peak value of -57.6VDC, peak to peak 65VDC, 34.4VRMS and a wave period of 20.60 μ s when 137.5VDC minimum input voltage is applied to the flyback converter at 20% load.



Şekil 81. Prototype Flyback Converter Output Diode Voltage at 20% Load (300mA) at Maximum (137.5VDC) Input Voltage

Fig. 82 shows the voltage signal on the 40mΩ resistor connected to the secondary of the transformer at 20% load at a maximum input voltage of 137.5VDC. In Eq. 4.11, the current on the output diode is calculated as 2.6A. In Fig. 82, symbol A indicates the 137.5VDC input voltage value and symbol B indicates the output diode voltage.

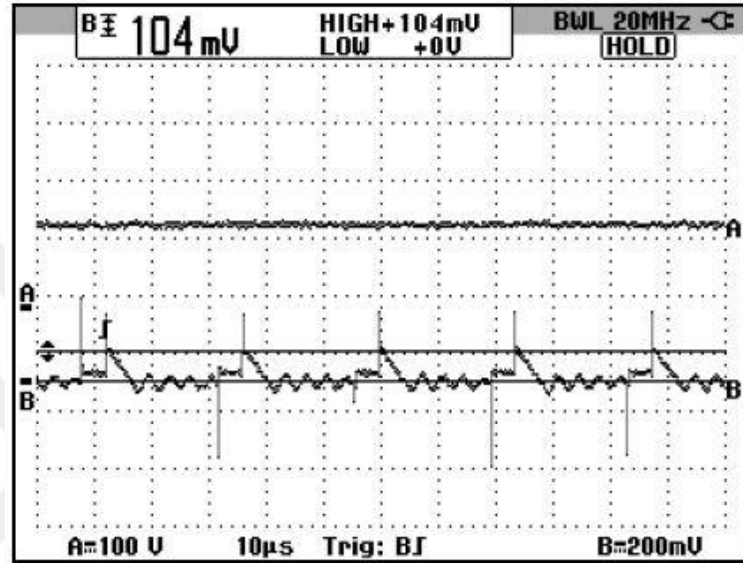


Figure 82. Prototype Flyback Converter Maximum (137.5VDC) Input Voltage at 20% Load (300mA) Voltage Over 40mΩ

$$I_D = \frac{V}{R} = \frac{104 * 10^{-3}}{40 * 10^{-3}} = 2.6A \quad 4.10$$

4.12 Input and Output Voltages at 100% (1.5A) Load in DIM Mode

The input and output voltages of the designed converter in DIM mode at 77VDC, 110VDC and 137.5VDC input voltages at 1.5A full load, in other words, at 100% load are shown. Fig. 83 shows the output voltage at 100% load when 77VDC minimum input voltage is applied in DIM mode. In Fig. 83, symbol A indicates the 77VDC input voltage value and symbol B indicates the output voltage. Fig. 83 analysed, 20.4VDC output voltage is generated when 77VDC minimum input voltage is applied to the flyback converter at 100% load.

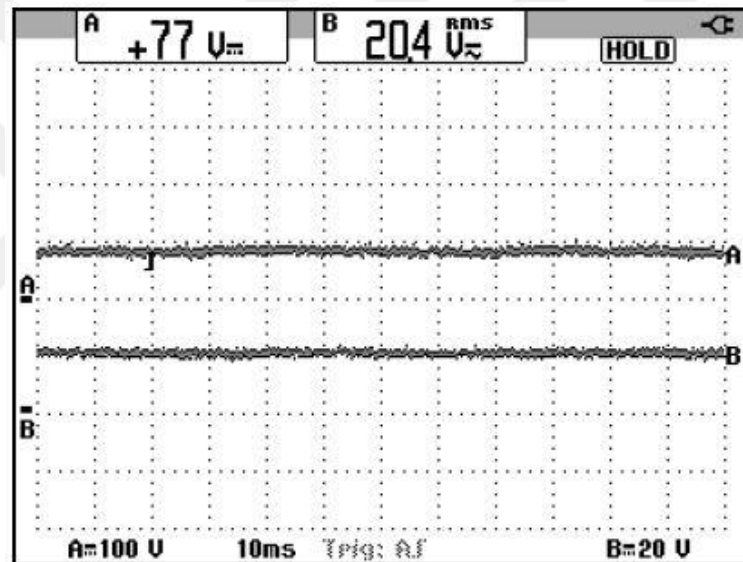


Figure 83. Prototype Flyback Converter Output Voltage at 100% Load (1.5A) at Minimum Input Voltage (77VDC) in DIM Mode

Fig. 84 shows the output voltage at 100% (1.5A) load when 110VDC nominal input voltage is applied in DIM mode. In Fig. 84, symbol A indicates the 110VDC input voltage value and symbol B indicates the output voltage. 20.4VDC output voltage is generated when 110VDC nominal input voltage is applied to the flyback converter at 100% load.

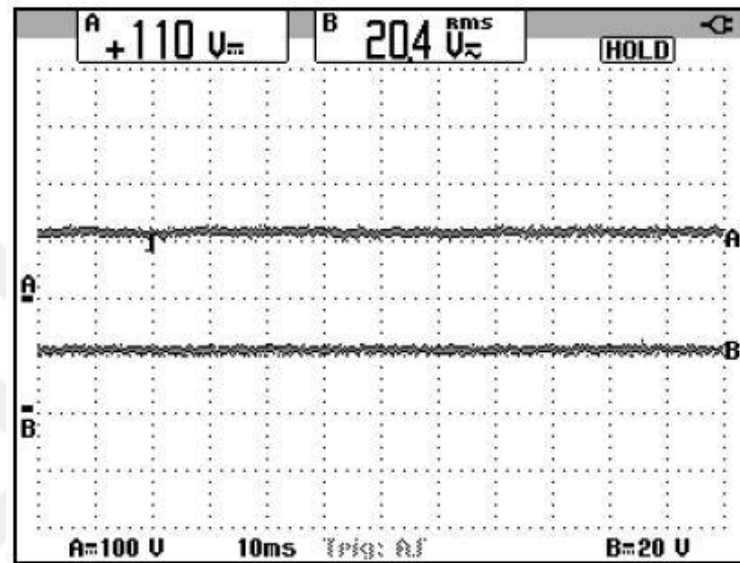


Figure 84. Prototype Flyback Converter Output Voltage at 100% Load (1.5A) at Nominal Input Voltage (110VDC) in DIM Mode

Fig. 85 shows the output voltage at 100% (1.5A) load when 137.5VDC maximum input voltage is applied in DIM mode. In Fig. 85, symbol A indicates the 137.5VDC input voltage value and symbol B indicates the output voltage. Fig. 85 analysed, 20.4VDC output voltage is generated when 137.5VDC maximum input voltage is applied to the flyback converter at 100% load.

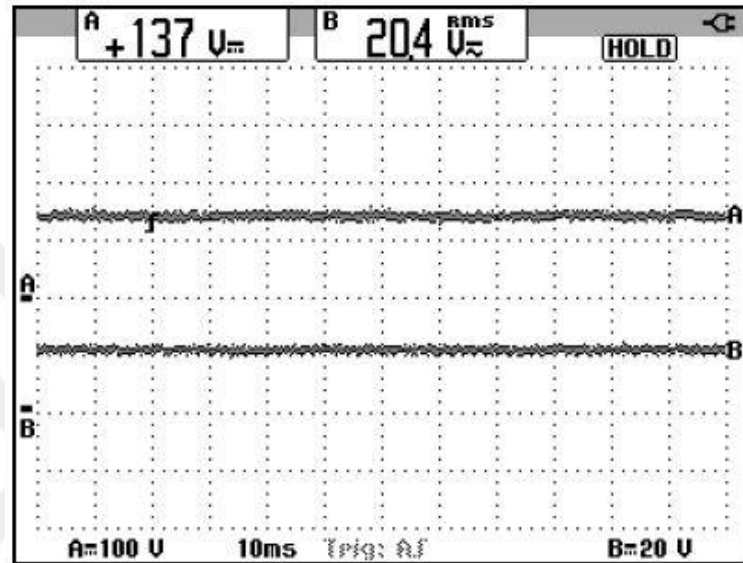


Figure 85. Prototype Flyback Converter Output Voltage at 100% Load (1.5A) at Maximum Input Voltage (137.5VDC) in DIM Mode

4.13 Input and Output Voltages at 20% (300mA) Load in DIM Mode

The input and output voltages of the designed converter are shown in DIM mode at 77VDC, 110VDC and 137.5VDC input voltages at 300mA low load, in other words at 100% load. Fig. 86 shows the output voltage at 20% (300mA) load when 77VDC minimum input voltage is applied in DIM mode. In Fig. 86, symbol A indicates the 77VDC input voltage value and symbol B indicates the output voltage. Fig. 86 analysed, 20.3VDC output voltage is generated when 77VDC minimum input voltage is applied to the flyback converter at 100% load.

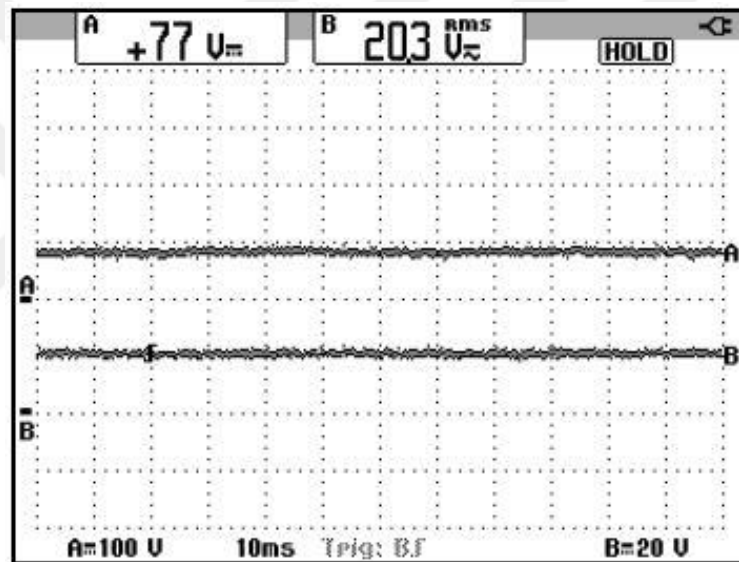


Figure 86. Prototype Flyback Converter Output Voltage at 20% Load (300mA) at Minimum Input Voltage (77VDC) in DIM Mode

Fig. 87 shows the output voltage at 20% (300mA) load when 110VDC nominal input voltage is applied in DIM mode. In Fig. 87, symbol A indicates the 110VDC input voltage value and symbol B indicates the output voltage. Fig. 87 analysed, 20.4VDC output voltage is generated when 110VDC nominal input voltage is applied to the flyback converter at 20% load.

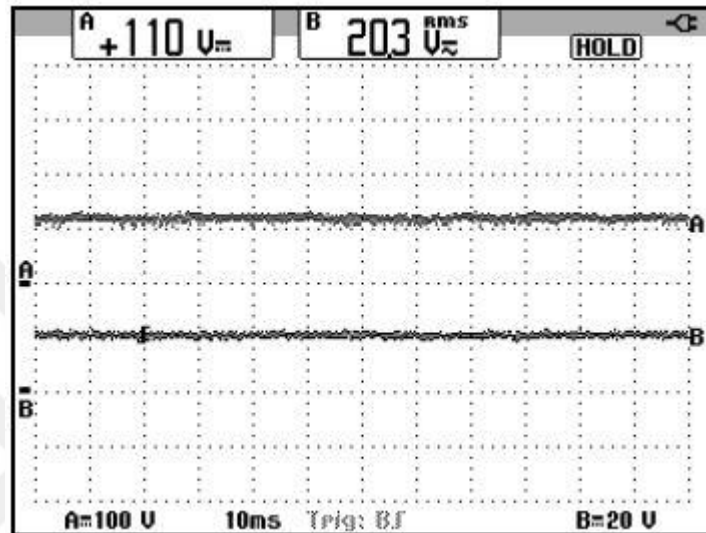


Figure 87. Output Voltage at 20% Load (300mA) at Nominal Input Voltage (110VDC) in Prototype Flyback Converter DIM Mode

Fig. 88 shows the output voltage at 20% (300mA) load when 137.5VDC maximum input voltage is applied in DIM mode. In Fig. 88, symbol A indicates the 137.5VDC input voltage value and symbol B indicates the output voltage. Fig. 88 analysed, 20.4VDC output voltage is generated when 137.5VDC maximum input voltage is applied to the flyback converter at 20% load.

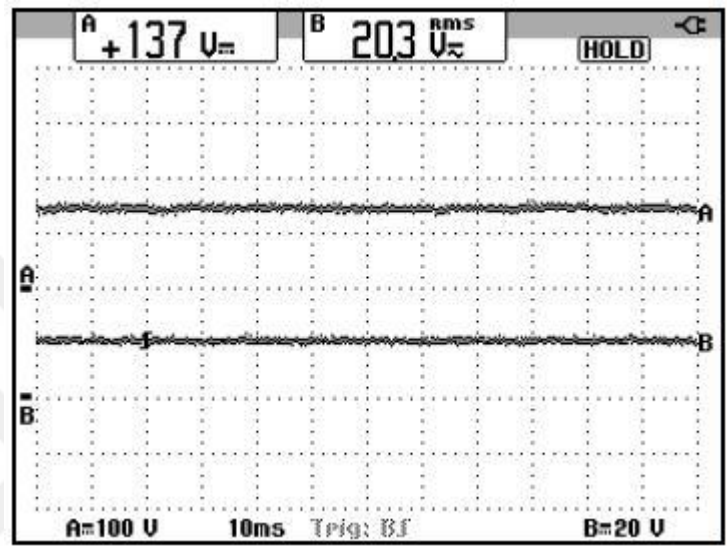


Figure 88. Prototype Flyback Converter Output Voltage at 100% Load (1.5A) at Maximum Input Voltage (137.5VDC) in DIM Mode

4.14 Voltage Regulations

The output voltage change of the flyback converter between 77VDC-137.5VDC voltages is shown with oscilloscope at 100% load, 20% and DIM mode.

4.14.1 Voltage regulation at 100% (1.5A) load. Fig. 89 shows the output voltage change at 100% (1.5A) load when 77VDC-137.5VDC maximum input voltage range is applied. In Fig. 89, symbol A indicates the change in input voltage value and symbol B indicates the output voltage. Fig. 89 analysed, a constant 24.4VDC output voltage occurs between 77VDC-137.5VDC input voltages at 100% load to the flyback converter.

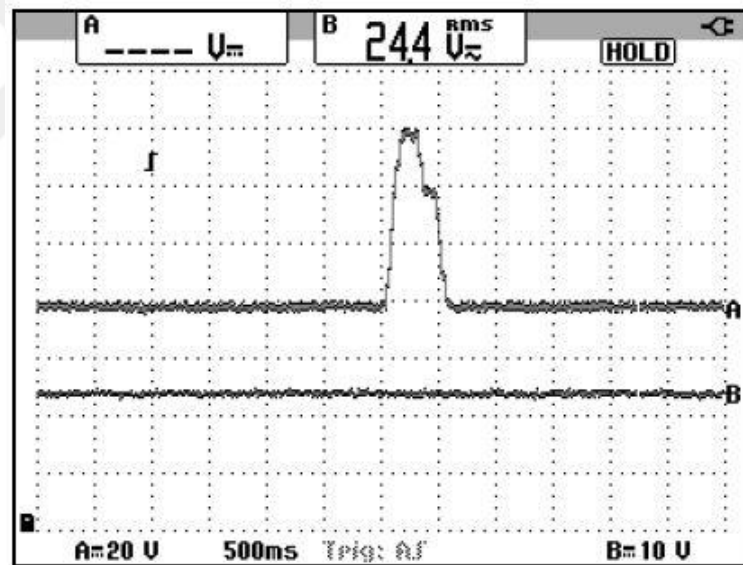


Figure 89. Prototype Flyback Converter Output Voltage Regulation at 100% Load (1.5A) between 77VDC-137.5VDC Input Voltage Range

4.14.2 Voltage regulation at 20%(300mA) load. Fig. 90 shows the output voltage change at 20% (300mA) load when 77VDC-137.5VDC maximum input voltage range is applied. In Fig. 90, symbol A indicates the change in input voltage value and symbol B indicates the output voltage. Fig. 90 analysed, a constant 24.4VDC output voltage is generated between 77VDC-137.5VDC input voltages at 20% load to the flyback converter.

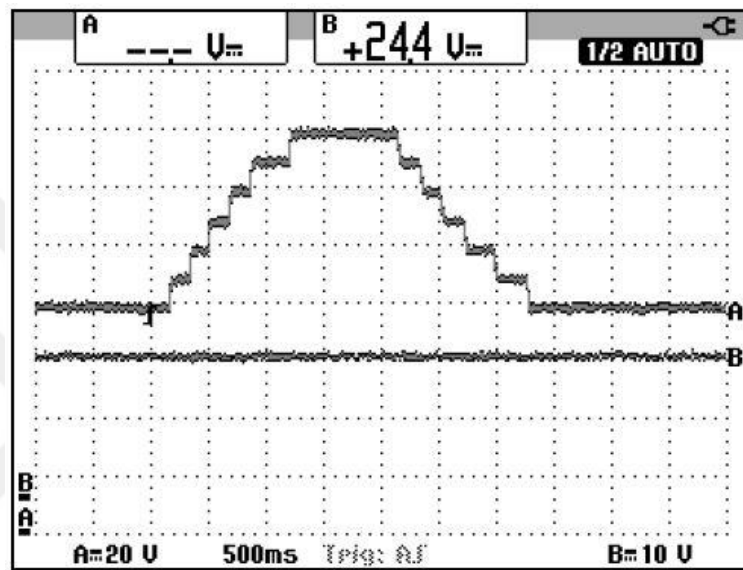


Figure 90. Prototype Flyback Converter Output Voltage Regulation at 20% Load (300mA) between 77VDC-137.5VDC Input Voltage Range

4.14.3 Voltage regulation in DIM mode at 100%(1.5A) load. Fig. 91 shows the output voltage variation at 100% (1.5A) load when 77VDC-137.5VDC maximum input voltage range is applied while the converter is operating in DIM mode. In Fig. 91, symbol A indicates the change in input voltage value and symbol B indicates the output voltage. Fig. 91 is analysed, a constant 20.4VDC output voltage is generated between 77VDC-137.5VDC input voltages at 100% load to the flyback converter.

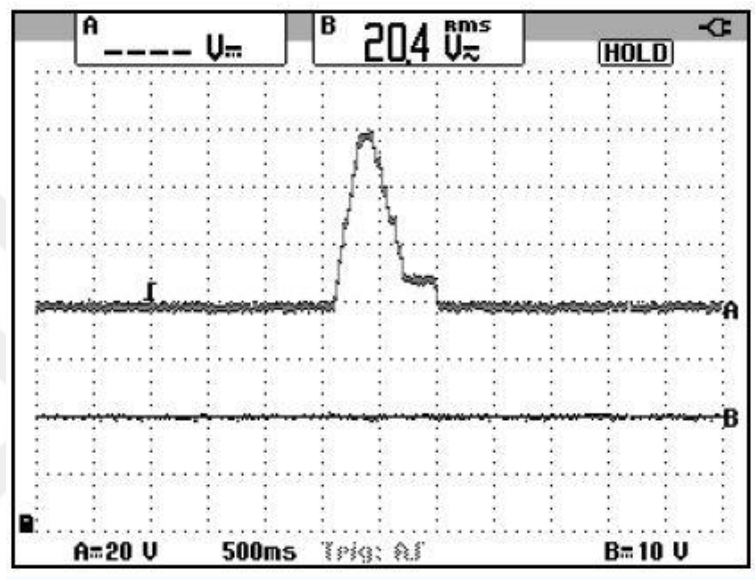


Figure 91. Output Voltage Regulation at 100% Load (1.5A) between 77VDC-137.5VDC Input Voltage Range in Prototype Flyback Converter DIM Mode

4.14.4 Voltage regulation in DIM mode at 20% (300mA) load. Fig. 92 shows the output voltage variation at 20% (300mA) load when the converter is operating in DIM mode with a maximum input voltage range of 77VDC-137.5VDC. In Fig. 92, symbol A indicates the change in input voltage value and symbol B indicates the output voltage. Fig. 92 analysed, a constant 20.3VDC output voltage is generated between 77VDC-137.5VDC input voltages at 20% load to the flyback converter.

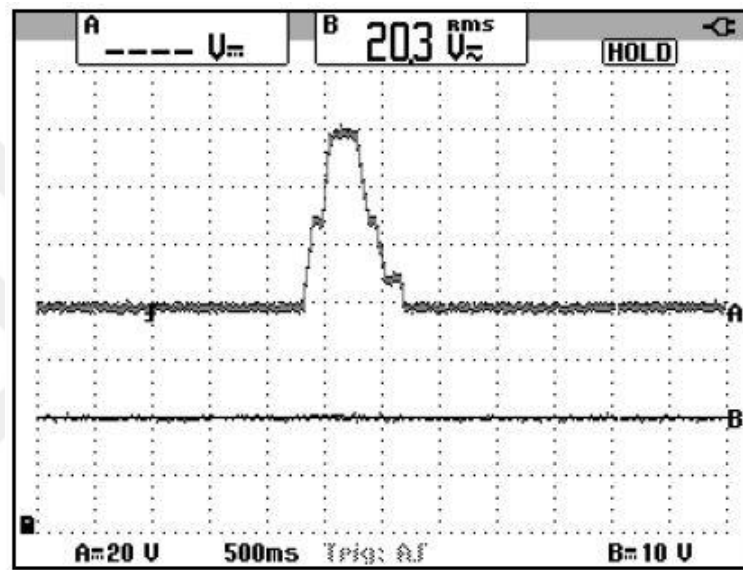


Figure 92. Output Voltage Regulation at 20% Load (300mA) between 77VDC-137.5VDC Input Voltage Range in DIM Mode of Prototype Flyback Converter

4.15 No-Load Condition

The input and output voltages of the designed converter at 77VDC, 110VDC and 137.5VDC are shown with oscilloscope. In addition, output voltage regulation between 77VDC-137.5VDC voltages without load connected to the converter is shown with oscilloscope.

4.15.1 Minimum input voltage (77VDC) output voltage in no load condition. Fig. 93 shows the output voltage at no-load condition when 77VDC minimum input voltage is applied. In Fig. 93, symbol A indicates the 77VDC input voltage value and symbol B indicates the output voltage. Fig. 93 analysed, 24.2VDC output voltage is generated when 77VDC minimum input voltage is applied to the flyback converter at no-load.

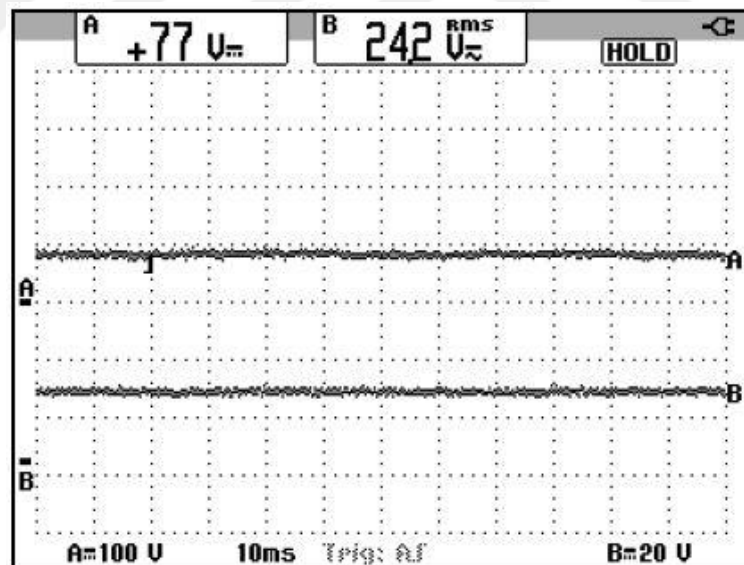


Figure 93. Prototype Flyback Converter Output Voltage at Minimum Input Voltage (77VDC) in No-Load Condition

4.15.2 Nominal input voltage (110VDC) output voltage at no load. Fig. 94 shows the output voltage at no load condition when 110VDC nominal input voltage is applied. In Fig. 94, symbol A indicates the 110VDC input voltage value and symbol B indicates the output voltage. Fig. 94 analysed, 24.2VDC output voltage is generated when 110VDC nominal input voltage is applied to the flyback converter at no-load.

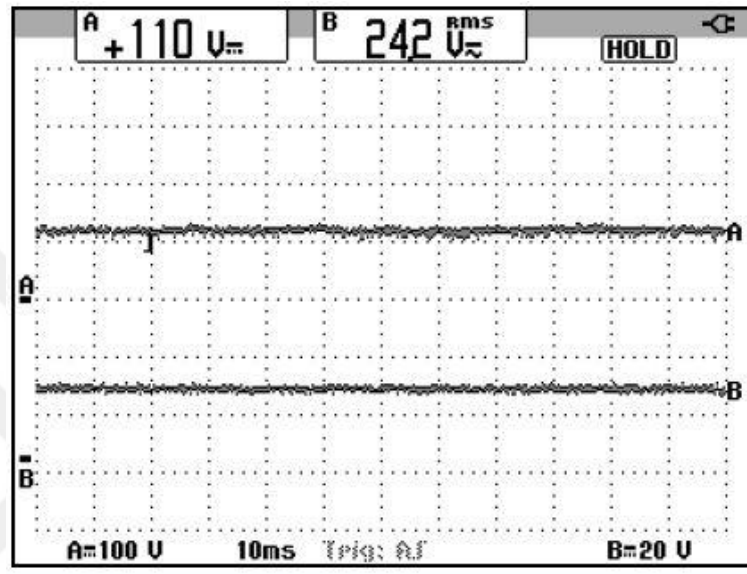


Figure 94. Prototype Flyback Converter Output Voltage at Nominal Input Voltage (110VDC) in No-Load Condition

4.15.3 Maximum input voltage (137.5VDC) output voltage at no load. Fig. 95 shows the output voltage at no load condition when the maximum input voltage of 137.5VDC is applied. In Fig. 95, symbol A indicates the 137.5VDC input voltage value and symbol B indicates the output voltage. Fig. 95 analysed, 24.2VDC output voltage is generated when 137.5VDC nominal input voltage is applied to the flyback converter at no-load.

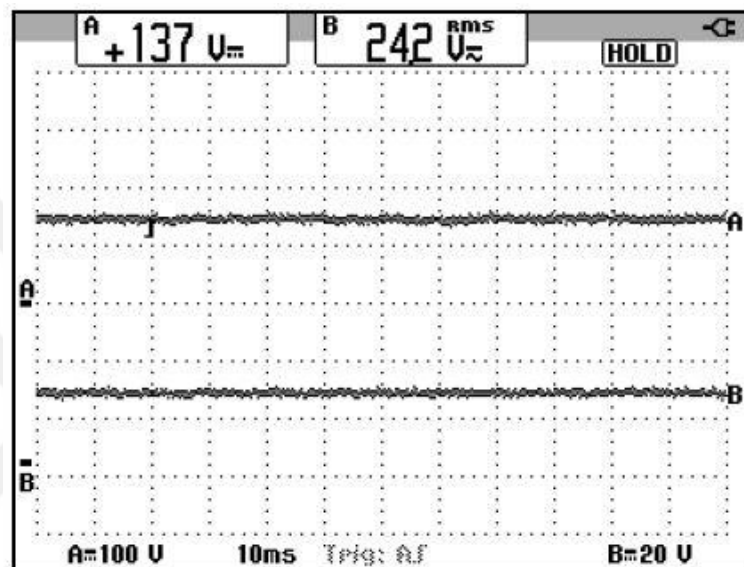


Figure 95. Prototype Flyback Converter Output Voltage at Maximum Input Voltage (137.5VDC) in No-Load Condition

4.15.4 Voltage regulation in no-load condition. Fig. 96 shows the output voltage variation in the no-load condition when the maximum input voltage range of 77VDC-137.5VDC is applied. In Fig. 96, symbol A indicates the change in input voltage value and symbol B indicates the output voltage. Fig. 96 is analysed, a constant 24.3VDC output voltage occurs between 77VDC-137.5VDC input voltages in the flyback converter in the no-load condition.

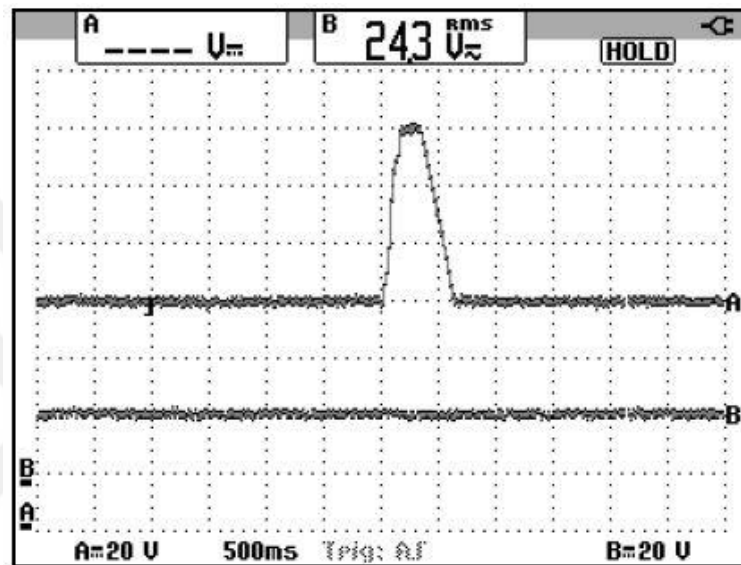


Figure 96. Output Voltage Regulation of the Prototype Flyback Converter between 77VDC-137.5VDC Input Voltage Range in No-Load Condition

4.16 Thermal Measurements

The prototype converter is tested and verified under maximum operating conditions. The prototype converter is under normal conditions at 137.5VDC maximum input voltage at 1.5A 100% load, +70°C and -40°C thermal camera measurements are displayed.

4.16.1 Thermal measurement under normal conditions. Fig. 97 shows the thermal image of the prototype flyback converter under normal conditions at 137.5VDC maximum input voltage and 1.5A 100% load. Fig. 97 analysed, it is observed that the ambient temperature is 25.8°C, LT8316 integration 38.1°C, MOSFET 48.7°C, transformer 52.5°C and output diode 68.2°C.

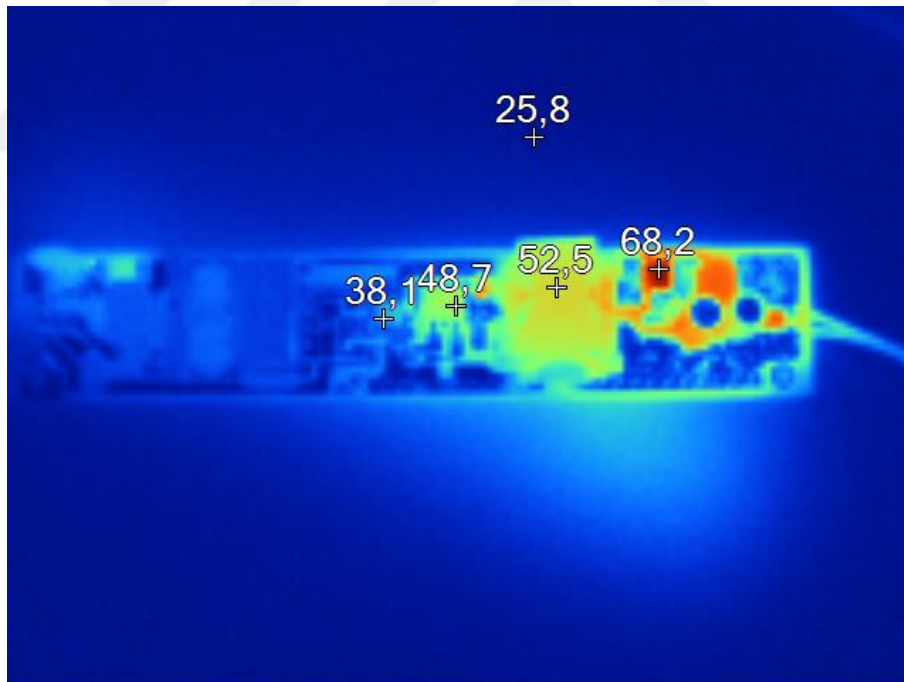


Figure 97. Prototype Flyback Converter Thermal Measurement at 100% Load (1.5A) at Maximum Input Voltage (137.5VDC) under Normal Conditions

4.16.2 Thermal measurement under high temperature conditions. Fig. 98 shows the thermal image of the prototype flyback converter at 137.5VDC maximum input voltage at 1.5A 100% load at 70°C high temperature conditions. Fig. 98 analysed, it is observed that the ambient temperature is 58°C, LT8316 integration is 67.7°C, MOSFET is 77.7°C, transformer is 76.9°C and output diode is 85.6°C.

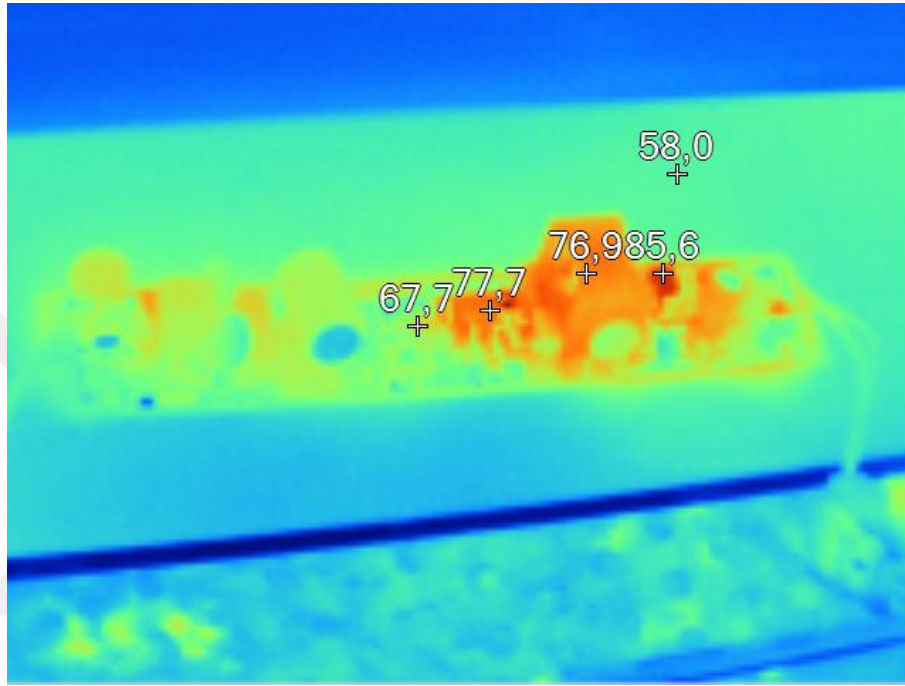


Figure 98. Prototype Flyback Converter Thermal Measurement at 100% Load (1.5A) at Maximum Input Voltage (137.5VDC) under High Temperature Conditions

4.16.3 Thermal measurement under low temperature conditions. Fig. 99 shows the thermal image of the prototype flyback converter at 137.5VDC maximum input voltage at 1.5A 100% load at -40°C° low temperature conditions. Fig. 99 shows that the ambient temperature is -40°C°, LT8316 integration is -18.6°C, MOSFET is -3.8°C, transformer is 8.7°C and output diode is 29.2°C.

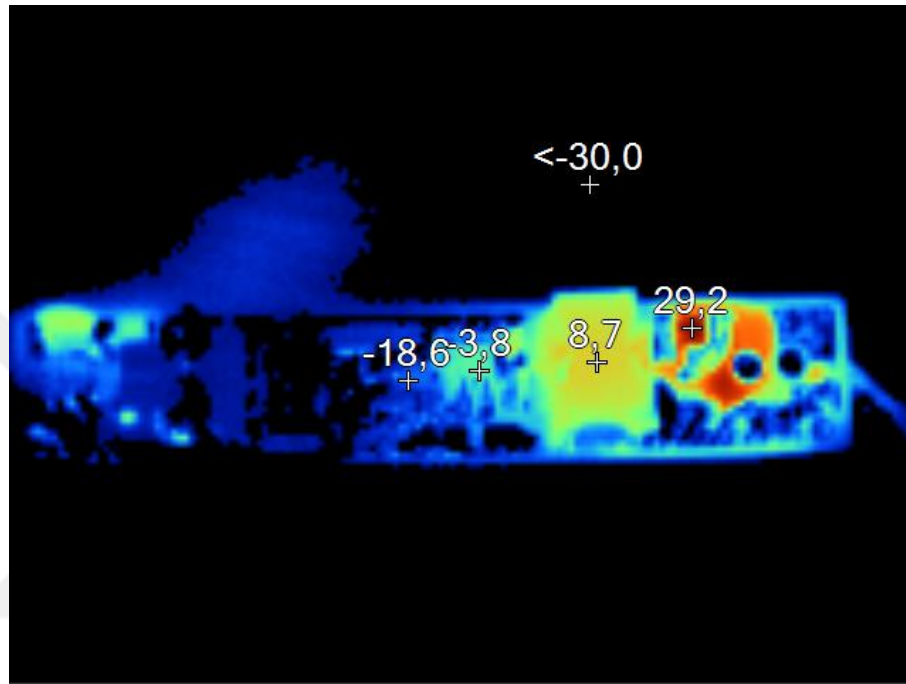


Figure 99. Prototype Flyback Converter Thermal Measurement at 100% Load (1.5A) at Maximum Input Voltage (137.5VDC) under Low Temperature Conditions

4.17 EMC Tests

The prototype flyback circuit was EMC conducted test using ROHDE&SCHWARZ brand HMS-X model Spectrum Analyser and ROHDE&SCHWARZ brand V-Network filter. EMC tests were performed at 77VDC minimum input voltage and 1.5A 100% load in DIM mode. Since the impedance of the circuit will be at the highest value at 77VDC minimum input voltage, it is the most decisive issue for the EMC filter and for this reason, tests were performed at 77VDC. Ambient noise is shown in light blue in the test graphs. The result in dark blue is the circuit noise.

Fig. 100 shows the conducted test result without EMC filter at 100% load at 77VDC minimum input voltage. In the test between 150kHz-30MHz, noise signals of 102 dB μ V at 250kHz, 92.45 dB μ V at 550kHz, 52.39 dB μ V at 21.75MHz were observed. The device does not comply with TS EN 50121-3-2 standard because the signals exceed 93 dB μ V and 99 dB μ V peak values.

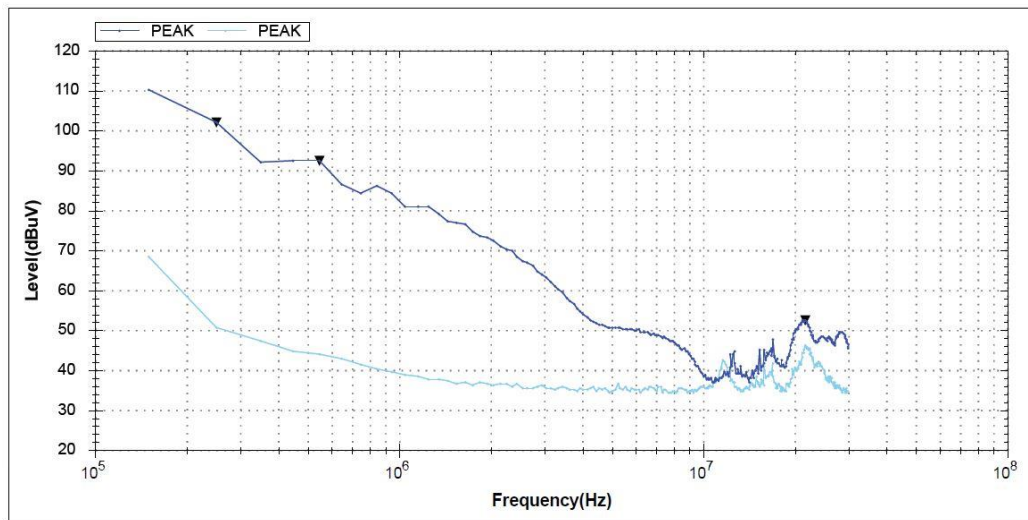


Figure 100. Prototype Flyback Circuit without EMC Filter

Fig. 101 shows the EMC filtered conducted test result at 100% load at 77VDC minimum input voltage. In the EMC test between 150kHz-30MHz, 52.16 dB μ V at 550kHz, 48.77 dB μ V at 13.05MHz, 48.26 dB μ V at 24.95MHz, 48.26 dB μ V at 24.95MHz and 48.88 dB μ V at 22.15MHz peak noise signals are observed. Flyback converter complies with TS EN 50121-3-2 standard when EMC filter is added.

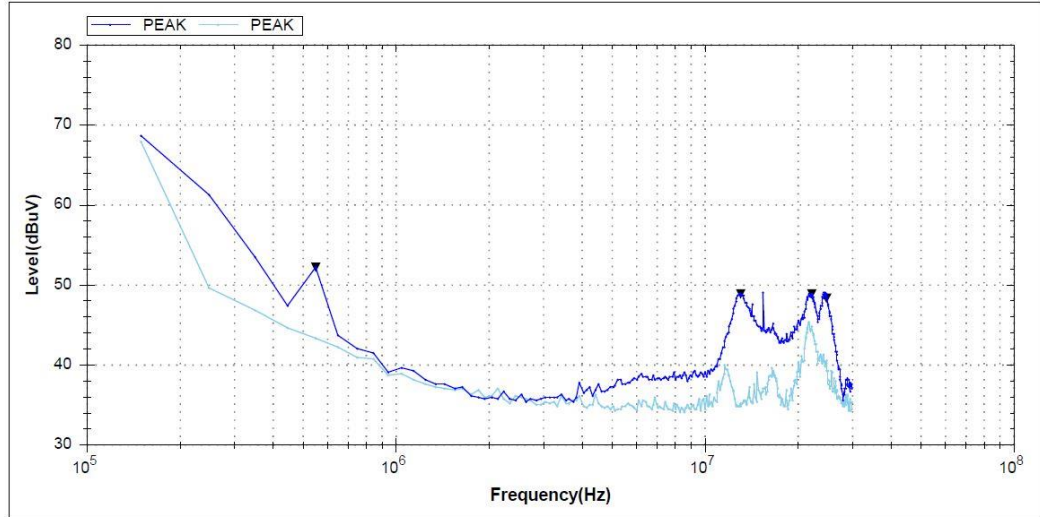


Figure 101. Prototype Flyback Circuit with EMC Filter

Fig. 102 shows the EMC filtered conducted test result at 100% load in DIM mode. In the EMC test between 150kHz-30MHz, noise signals with peak values of 50.84 dB μ V at 550kHz, 52.88 dB μ V at 13.05MHz, 50.03 dB μ V at 22.05MHz, 50.03 dB μ V at 24.75 MHz 50.80 dB μ V are observed. When EMC filter is added to the flyback converter, it complies with TS EN 50121-3-2 standard when operating in DIM mode.

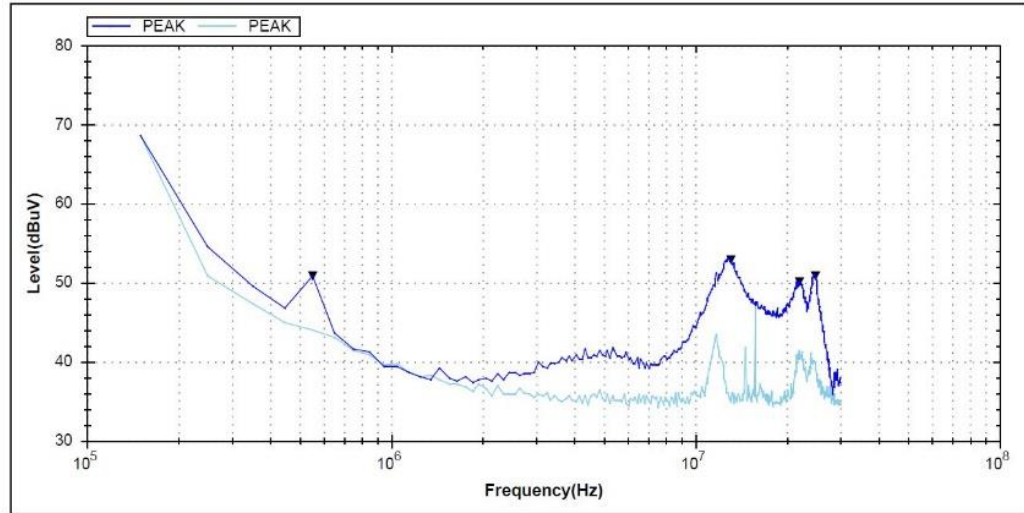


Figure 102. Prototype Flyback Circuit with EMC Filter in DIM Mode

4.18 Efficiency

The efficiency analysis of the prototype flyback converter at 100% (1.5A) load is given in Table 4. Measurements were made with FLUKE brand 115 model multimeter. The efficiency of the converter is calculated to be approximately 90%.

Table 4

Flyback Converter Efficiency

Efficiency				
V_{in}	V_{out}	I_{in}	I_{out}	$Eff(\%)$
77VDC	24.59VDC	0.543A	1.5A	88.22
80VDC	24.59VDC	0.521A	1.5A	88.50
90VDC	24.59VDC	0.462A	1.5A	88.71
100VDC	24.59VDC	0.411A	1.5A	89.74
110VDC	24.59VDC	0.380A	1.5A	88.24
120VDC	24.59VDC	0.345A	1.5A	89.09
137.5VDC	24.59VDC	0.299A	1.5A	89.72

4.19 Railway Systems Industry Product Comparison

In this section, the features and prices of industrial products that are close to the flyback converter designed within the scope of the thesis are researched and transferred to Table 4. The prices of industrial products are taken from Premier Farnell website.

Table 5

Flyback Converter Product Comparison

Manufacturer	MeanWell	Traco Power	This Thesis
Model	RSD-30H-24	TEQ40-7215WIR	
Input Voltage	40-160VDC	43-160VDC	60-160VDC
Output Voltage	24VDC	24VDC	24VDC
Output Current	1.25A	1.67A	1.5A
Output Power	30W	40W	36W
Efficiency	89%	91%	90%
Output V. Ripple	50mVp-p	150 mVp-pu	21mVp-p
Options	LED indicator for power on	Protection against overvoltage, Short circuit overtemperature	Protection against overvoltage Short circuit overtemperature Adjustable Dimmable IP5K1 protection
Standard	EN50155/EN 45545-2	EN 50155	EN 50155, EN 55015/EN 50121, EN 15016
FARNELL Price	43,59€	184,87€	20€

4.20 Flyback Converter Circuit Scheme

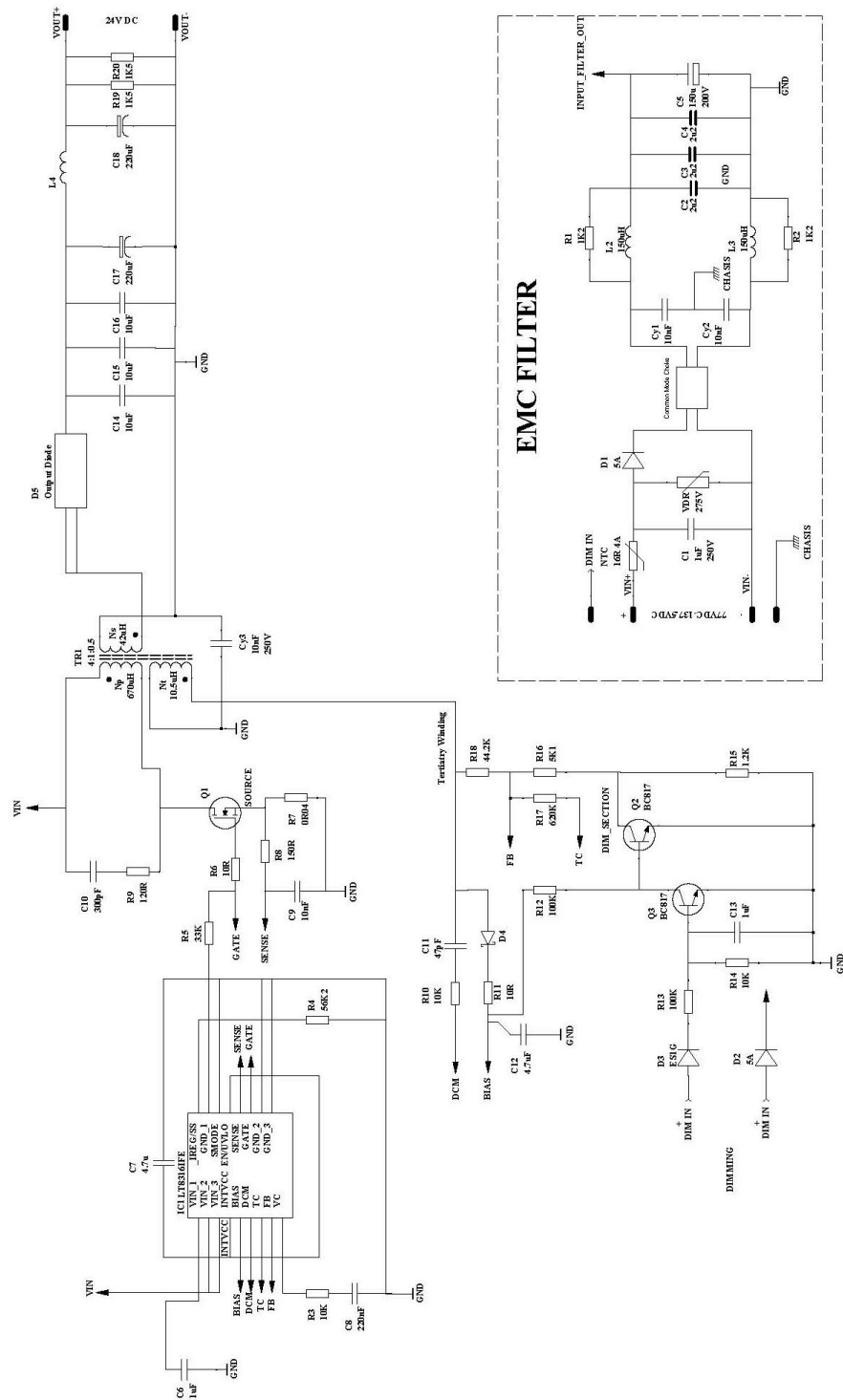


Figure 103. Flyback Converter Circuit Scheme

Chapter 5

Conclusions and Feature Work

In this thesis, flyback converter design in accordance with Railway standards is aimed and applied to practice. In the first step for the design, flyback topology is analysed and system parameters are determined. According to the system parameters, flyback converter calculations are made, switching element is determined and transformer design is made. RC snubber circuit is designed and implemented to reduce the switching losses of the flyback converter, increase the efficiency and operate in the safe region. EMC filter circuit has been designed and applied to the circuit to comply with railway standards. Considering the analyses and calculations, the flyback converter circuit was simulated with LTspice programme. The simulated converter was transferred to practice with PCB and tested in the laboratory environment. An additional feature DIM mode has been added to the converter and an alternative is offered to the user, when the led strip is connected, the user will be able to determine the light brightness by providing soft transition, and the converter has been made suitable for use in the industry with the aim of increasing the demand for the converter.

As a result of the tests, an output voltage of 24.4VDC was observed at 100% (1.5A) and 20% (300mA) load within the parameters specified for the converter on the oscilloscope, in other words, in the 77VDC-137.5VDC input voltage range. It is observed that the output ripple voltage is less than 100mV at 100% and 20% load in the 77VDC-137.5VDC input voltage range. MOSFET calculations were performed and a low-priced MOSFET with low R_{DS}^{on} resistance was selected for thermal performance in order to reduce the cost. An RC snubber circuit was designed and applied to the circuit in order to dampen the voltage spikes that occur during switching. The RC snubber circuit reduces the switching losses caused by the ringing on the MOSFET. The converter was tested at high and low temperatures and the device did not deteriorate. EMC filter was designed and applied to the circuit and made suitable for electromagnetic compatibility. EN 50155 Railway Applications Electronic Devices tests and TS EN 50121 Railway Applications Electromagnetic Compatibility tests

have been met. Due to the hard switching in the converter, power loss occurs and the leakage inductances of the transformer also cause power loss. This situation reduces the power loss, in other words, the efficiency of the converter.

Ouyang & Andersen achieved over 90% efficiency by applying active RCD clamp on flyback converter using asymmetric half bridge technology. Nagesha et al achieved a maximum efficiency of 85.14% on flyback converter despite using SiC MOSFET. Zhao et al achieved 63%-75% efficiency by applying lossless RCD snubber to the flyback circuit operating in DCM mode. Chandra et al achieved 70% efficiency by using soft start in flyback converter. The flyback converter designed in this thesis is more efficient than the converters in the literature by using hard switching and passive RC snubber, and its cost is lower than the flyback converters in the industry.

The flyback converter designed within the scope of this thesis complies with railway standards, operates at full power and low power between -40C +70C temperatures, adjusts the light brightness with the DIM feature, operates between 77VDC-137.5VDC input voltages, gives a maximum output current of 1.5A and provides 36W output power. In addition, the converter output has short circuit protection and over temperature protection features. The efficiency of the converter is approximately 90% as stated in Table 4. In order to increase the efficiency of the converter, the transformer should be carefully wound and if the shielding on the cables is applied to the transformer, the leakage inductances will be reduced. Soft switching instead of hard switching on the flyback converter will reduce power loss and increase efficiency. In addition, since the converter PCB has two layers, capacitive effects on the PCB are high. If the PCB is designed with four layers, the capacitive effects caused by the paths will decrease and the efficiency will increase.

REFERENCES

- A. A. Mohammed and S. M. Nafie, "Flyback converter design for low power application," 2015 International Conference on Computing, Control, Networking, Electronics and Embedded Systems Engineering (ICCNEEE), 2015, pp. 447-450, doi: 10.1109/ICCNEEE.2015.7381410.
- A. Gupta, K. Rana and D. Joshi, "Design and Analysis of Multioutput Flyback Converter under Continuous and Discontinuous Conduction mode using PID Controller," 2020 First IEEE International Conference on Measurement, Instrumentation, Control and Automation (ICMICA), 2020, pp. 1-5, doi: 10.1109/ICMICA48462.2020.9242853.
- A. V. Magar, S. G. Kanade and A. P. Kinge, "Transformerless Buck-Boost DC-DC Converter," 2018 IEEE Global Conference on Wireless Computing and Networking (GCWCN), 2018, pp. 200-203, doi: 10.1109/GCWCN.2018.8668646.
- Adams, J. (2001). "Application Note AN-1024 Flyback Transformer Design for the IRIS40xx Series". Retrieved on 15 January 2023 from <https://www.infineon.com/dgdl/an-1024.pdf?fileId=5546d462533600a401535591115e0f6d>.
- Analog Devices (2019). 560V_{IN} Micropower No-Opto Isolated Flyback Controller, Retrieved 6 November 2022 from <https://www.analog.com/media/en/technical-documentation/data-sheets/lt8316.pdf>.
- B. M. Hasaneen and A. A. Elbaset Mohammed, "Design and simulation of DC/DC boost converter," 2008 12th International Middle-East Power System Conference, 2008, pp. 335-340, doi: 10.1109/MEPCON.2008.4562340.
- B. M. Mrabet, A. Chammam and W. Nsibi, "Constant-Current LEDs driver based on DC-DC buck converter with dimming capability," 2020 6th IEEE International

Energy Conference (ENERGYCon), 2020, pp. 204-209, doi: 10.1109/ENERGYCon48941.2020.9236432.

B. R. Kiran and G. A. Ezhilarasi, "Design and analysis of soft-switched Buck-Boost Converter for PV applications," 2015 Annual IEEE India Conference (INDICON), 2015, pp. 1-5, doi: 10.1109/INDICON.2015.7443509.

Betten, J. (2016). "Power Tips: Calculate an R-C snubber in seven steps" Retrieved on 11 January 2023 from https://e2e.ti.com/blogs_/b/powerhouse/posts/calculate-an-r-c-snubber-in-seven-steps.

C. Deekshitha and K. L. Shenoy, "Design and simulation of synchronous buck converter for LED application," 2017 2nd IEEE International Conference on Recent Trends in Electronics, Information & Communication Technology (RTEICT), 2017, pp. 142-146, doi: 10.1109/RTEICT.2017.8256574.

C. Tao and A. Fayed, "Output spectrum analysis of buck converters in DCM with PFM control," 2012 IEEE International Symposium on Circuits and Systems (ISCAS), 2012, pp. 2267-2270, doi: 10.1109/ISCAS.2012.6271745.

Choi, H. S. (2012). "AN-4140 Transformer Design Consideration for Offline Flyback Converters Using Fairchild Power Switch (FPS™)". Retrieved on 16 January 2023 <https://www.onsemi.com/pub/Collateral/AN-4140.pdf>.

Choi, H.S. 2004. "Application Note AN4137 Design Guidelines for Off-line Flyback Converters Using Fairchild Power Switch (FPS)". Retrieved on 1 January 2023 from <https://www.all-electronics.de/wp-content/uploads/migrated/document/167525/46fbe8c82a6.pdf>

Coilcraft, Inc. (2022). Ferrite vs. *Pressed Powder-core Inductors*, Retrieved on 17 January 2023 https://www.coilcraft.com/getmedia/5196d074-40d5-4f53-ac0b-bf458214187d/Doc1733_Ferrite_Vs_Composite.pdf

D. Farrakhov, K. Barabanov, A. Podguzov, I. Yamalov and R. Urazbakhtin, "Quasi-resonant buck converter for high power application," 2020 International

Conference on Electrotechnical Complexes and Systems (ICOECS), 2020, pp. 1-4, doi: 10.1109/ICOECS50468.2020.9278426.

D. Joshi, J. Kumar and A. Pachori, "Design and development of dual output DC to DC converter based on flyback topology," 2020 International Conference on Power, Energy, Control and Transmission Systems (ICPECTS), 2020, pp. 1-4, doi: 10.1109/ICPECTS49113.2020.9337042.

D. Sevim and N. Bozkurt , "DA-DA Düşürücü Dönüştürücünün Sistem Tanımlaması ve Kontrolü", *Avrupa Bilim ve Teknoloji Dergisi*, no. 31, pp. 881-889, Dec. 2021, doi:10.31590/ejosat.1019650

Dinwoodie, L. (1999). "Design Review: Isolated 50 Watt Flyback Converter Using the UCC3809 Primary Side Controller and the UC3965 Precision Reference and Error Amplifier". Retrieved on 18 January 2023 https://www.ti.com/lit/an/slua086/slua086.pdf?ts=1678812512286&ref_url=https%253A%252F%252Fwww.google.com%252F.

E. Bektaş , K. Ç. Bayındır , M. Z. Erel , Y. Yalman and S. Kıvrak , "İki Çıkışlı Flyback Dönüştürücü Tasarımı ve Analizi", *Fırat Üniversitesi Mühendislik Bilimleri Dergisi*, vol. 30, no. 3, pp. 51-57, Sep. 201

F. Pouladi, H. Farzanehfard and E. Adib, "Battery Operated Soft Switching Resonant Buck–Boost LED Driver With Single Magnetic Element," in *IEEE Transactions on Power Electronics*, vol. 34, no. 3, pp. 2704-2711, March 2019, doi: 10.1109/TPEL.2018.2847406

G. D. Costa, C. H. I. Font and I. Barbi, "Comprehensive Analysis of a Flyback Converter for Voltage Equalization of Battery Strings," 2018 13th IEEE International Conference on Industry Applications (INDUSCON), 2018, pp. 1200-1207, doi: 10.1109/INDUSCON.2018.8627178.

G. VACHEVA, V. DIMITROV and N. HINOV, "Modelling and Control of Bidirectional Buck-Boost Converter for Electric Vehicles Applications," 2019

16th Conference on Electrical Machines, Drives and Power Systems (ELMA), 2019, pp. 1-4, doi: 10.1109/ELMA.2019.8771496.

G. Zamora-Mejía, J. Martínez-Castillo, A. Díaz-Sánchez, J. M. Rocha-Pérez and V. R. González-Díaz, "Lower limits of DC-DC power conversion in circuit integrated Buck Converters and 0.50 μ m CMOS 1.65V/250mA buck converter design," 2016 IEEE International Engineering Summit, II Cumbre Internacional de las Ingenierías (IE-Summit), 2016, pp. 1-5, doi: 10.1109/IESummit.2016.7459758.

H. A. Mohamed, H. A. Khattab, A. Mobarka and G. A. Morsy, "Design, control and performance analysis of DC-DC boost converter for stand-alone PV system," 2016 Eighteenth International Middle East Power Systems Conference (MEPCON), 2016, pp. 101-106, doi: 10.1109/MEPCON.2016.7836878.

Hart, W. D. (2010). *Power Electronics*. New York: McGraw-Hill Education.

J. S. A. Rahavi, T. Kanagapriya and R. Seyezhai, "Design and analysis of Interleaved Boost Converter for renewable energy source," 2012 International Conference on Computing, Electronics and Electrical Technologies (ICCEET), 2012, pp. 447-451, doi: 10.1109/ICCEET.2012.6203850.

Jean, P. 2010. "Under the Hood of Flyback SMPS Designs, Topic Categories: Design Reviews – Functional Circuit Blocks Specific Power Topologies". Retrieved on 1 January 2023 from <https://www.ti.com/seclit/ml/slup261/slup261.pdf>.

K. V. S. M. Chandra, R. Deekshit, Nageshwar Rao S, B. K. Singh, V. Chippalkatti and T. Kanthimathinathan, "Design and implementation of Current mode controlled multiple output flyback converter for military applications," 2015 IEEE International Conference on Electronics, Computing and Communication Technologies (CONECCT), 2015, pp. 1-6, doi: 10.1109/CONECCT.2015.7383910.

- Kamil M. 2007. "Switch Mode Power Supply (SMPS) Topologies (Part I) ". Retrieved on 20 December 2022 from <https://www.microchip.com/downloads/en/Appnotes/01114A.pdf>.
- L. Cheng, P. Acuna, R. P. Aguilera, M. Ciobotaru and J. Jiang, "Model predictive control for DC-DC boost converters with constant switching frequency," 2016 IEEE 2nd Annual Southern Power Electronics Conference (SPEC), 2016, pp. 1-6, doi: 10.1109/SPEC.2016.7846189.
- L. -S. Yang and T. -J. Liang, "Analysis and Implementation of a Novel Bidirectional DC–DC Converter," in IEEE Transactions on Industrial Electronics, vol. 59, no. 1, pp. 422-434, Jan. 2012, doi: 10.1109/TIE.2011.2134060.
- M. B. Martin, "A Design of 3.3W Closed Loop Flyback Converter with 3.3V, 1A Output for Low Voltage Applications," 2020 IEEE 8th Conference on Systems, Process and Control (ICSPC), 2020, pp. 86-90, doi: 10.1109/ICSPC50992.2020.9305790.
- M. D. Rahman, M. Nazaf Rabbi and G. Sarowar, "Development of DC-DC Converters – A Review," 2021 International Conference on Computational Performance Evaluation (ComPE), Shillong, India, 2021, pp. 341-347, doi: 10.1109/ComPE53109.2021.9752028.
- M. Forouzesh, Y. P. Siwakoti, S. A. Gorji, F. Blaabjerg and B. Lehman, "Step-Up DC–DC Converters: A Comprehensive Review of Voltage-Boosting Techniques, Topologies, and Applications," in IEEE Transactions on Power Electronics, vol. 32, no. 12, pp. 9143-9178, Dec. 2017, doi: 10.1109/TPEL.2017.2652318.
- M. Gerstner, M. Maerz and A. Dietz, "Review-based Selection Recommendation for Galvanically Isolated DC/DC Converters Designed For a Wide Input Voltage Range," 2021 IEEE 19th International Power Electronics and Motion Control Conference (PEMC), Gliwice, Poland, 2021, pp. 167-172, doi: 10.1109/PEMC48073.2021.9432537

- M. Li, Z. Ouyang and M. A. E. Andersen, "Analysis and Optimal Design of High-Frequency and High-Efficiency Asymmetrical Half-Bridge Flyback Converters," in IEEE Transactions on Industrial Electronics, vol. 67, no. 10, pp. 8312-8321, Oct. 2020, doi: 10.1109/TIE.2019.2950845.
- M. Nymand and M. A. E. Andersen, "High-Efficiency Isolated Boost DC–DC Converter for High-Power Low-Voltage Fuel-Cell Applications," in IEEE Transactions on Industrial Electronics, vol. 57, no. 2, pp. 505-514, Feb. 2010, doi: 10.1109/TIE.2009.2036024.
- M. Shadoul, R. A. Abri, H. Yousef and M. Masoud, "DC- DC Boost Converter Controller Design for PV Applications," 2017 9th IEEE-GCC Conference and Exhibition (GCCCE), 2017, pp. 1-6, doi: 10.1109/IEEEGCC.2017.8448164.
- METERS UK (2019). American Wire Gauge (AWG) Cable / Conductor Sizes. Retrieved 6 November 2022 <https://metersuk.co.uk/american-wire-gauge-awg-cable-conductor-sizes/>.
- N. Coruh, S. Urgun and T. Erfidan, "Design and implementation of flyback converters," 2010 5th IEEE Conference on Industrial Electronics and Applications, 2010, pp. 1189-1193, doi: 10.1109/ICIEA.2010.5515894.
- N. H. Baars, H. Huisman, J. L. Duarte and J. Verschoor, "A 80 kW isolated DC-DC converter for railway applications," 2014 16th European Conference on Power Electronics and Applications, 2014, pp. 1-10, doi: 10.1109/EPE.2014.6910741.
- N. -M. Sze, W. -H. Ki and C. -Y. Tsui, "Threshold Voltage Start-up Boost Converter for Sub-mA Applications," 4th IEEE International Symposium on Electronic Design, Test and Applications (delta 2008), 2008, pp. 338-341, doi: 10.1109/DELTA.2008.126.
- N. Nitesh Kumar, S. R. Patil, Prabhudeva, S. S and S. Ravichandran, "Feedback Controller Design For A DC-DC Buck Converter," 2021 Fourth International Conference on Electrical, Computer and Communication Technologies (ICECCT), 2021, pp. 1-5, doi: 10.1109/ICECCT52121.2021.9616954.

- N. Shinde, S. Sankad and S. L. Patil, "Design and Study Voltage Characteristics of Buck Converter by Matlab Simulink," 2018 2nd International Conference on Trends in Electronics and Informatics (ICOEI), 2018, pp. 680-683, doi: 10.1109/ICOEI.2018.8553695.
- Nagesha C, A. Sreedevi and M. Gopal, "Simulation and hardware implementation of 24 watt multiple output Flyback converter," 2015 International Conference on Power and Advanced Control Engineering (ICPACE), 2015, pp. 366-370, doi: 10.1109/ICPACE.2015.7274974.
- P. Suskis and D. Nikonorov, "Buck Converter Digital Control System Design and Performance Evaluation," 2018 IEEE 6th Workshop on Advances in Information, Electronic and Electrical Engineering (AIEEE), 2018, pp. 1-4, doi: 10.1109/AIEEE.2018.8592368.
- P. Zimoch, K. Kierepka and M. Kasprzak, "Feasibility of High Frequency Zero-Voltage Switching Boost Converters Achieving High Power Density Using Wide-Bandgap Devices," 2019 International Conference on Electrical Drives & Power Electronics (EDPE), 2019, pp. 24-29, doi: 10.1109/EDPE.2019.8883897.
- Pesonen J. 2014. "Improving the Performance of Traditional Flyback- Topology With Two-Switch –Approach". Retrieved on 22 December 2022 from <https://www.ti.com/lit/an/snva716/snva716.pdf>.
- Picard, J. (2010). "Under the Hood of Flyback SMPS Designs". Retrieved on 16 January 2023 from <https://www.ti.com/seclit/ml/slup261/slup261.pdf>.
- Pressman A. I., Billings K., Morey T. (2009). Switching Power Supply Design (3rd ed.). New York: McGraw-Hill.
- R. Ayop, C. W. Tan, S. N. Syed Nasir, K. Y. Lau and C. Ling Toh, "Buck Converter Design for Photovoltaic Emulator Application," 2020 IEEE International Conference on Power and Energy (PECon), 2020, pp. 293-298, doi: 10.1109/PECon48942.2020.9314582.

- R. Kanthimathi and J. Kamala, "Analysis of different flyback Converter topologies," 2015 International Conference on Industrial Instrumentation and Control (IIC), Pune, India, 2015, pp. 1248-1252, doi: 10.1109/IIC.2015.7150939.
- R. M. Sudhan Rao, M. Asad and A. K. Singha, "Analysis and Design of a Digital Average Current-mode Controlled Buck Converter," 2020 IEEE International Conference on Power Electronics, Drives and Energy Systems (PEDES), 2020, pp. 1-4, doi: 10.1109/PEDES49360.2020.9379636.
- R. R. Duarte, G. F. Ferreira, M. A. Dalla Costa and J. M. Alonso, "Performance investigation of silicon and gallium nitride transistors in an integrated double buck-boost LED driver," 2017 IEEE Industry Applications Society Annual Meeting, 2017, pp. 1-5, doi: 10.1109/IAS.2017.8101811.
- Ridley, R. (2005). Flyback Converter Snubber Design. Retrieved on 13 January 2023 https://ridleyengineering.com/images/pdf-design_center/Flyback_Snubber_Design.pdf.
- Rui Ling, Guoyan Zhao, Qin Huang, "High step-up interleaved boost converter with low switch voltage stress," Electric Power Systems Research, Volume 128, 2015, Pages 11-18, ISSN0378-7796, <https://doi.org/10.1016/j.epsr.2015.06.016>.
- S. Jagannath, N. Agarwal, S. Balasubramaniasarma, K. W. Ma and V. B, "Design and Analysis of Single SiC MOSFET Switch Flyback Converter based Control Power Supply for Renewable Applications," 2020 International Conference on Power, Instrumentation, Control and Computing (PICC), 2020, pp. 1-5, doi: 10.1109/PICC51425.2020.9362348.
- S. Masri, N. Mohamad and M. H. M. Hariri, "Design and development of DC-DC buck converter for photovoltaic application," 2012 International Conference on Power Engineering and Renewable Energy (ICPERE), 2012, pp. 1-5, doi: 10.1109/ICPERE.2012.6287236.
- S. N. Soheli, G. Sarowar, M. A. Hoque and M. S. Hasan, "Design and Analysis of a DC -DC Buck Boost Converter to Achieve High Efficiency and Low Voltage

- Gain by using Buck Boost Topology into Buck Topology," 2018 International Conference on Advancement in Electrical and Electronic Engineering (ICAESEE), 2018, pp. 1-4, doi: 10.1109/ICAESEE.2018.8643001.
- S. Oucheriah and L. Guo, "PWM-Based Adaptive Sliding-Mode Control for Boost DC–DC Converters," in IEEE Transactions on Industrial Electronics, vol. 60, no. 8, pp. 3291-3294, Aug. 2013, doi: 10.1109/TIE.2012.2203769.
- S. Singh, G. Bhuvaneswari and B. Singh, "Power quality improvement in multi-output forward boost converter," India International Conference on Power Electronics 2010 (IICPE2010), 2011, pp. 1-5, doi: 10.1109/IICPE.2011.5728105.
- S. Sunkara and N. Kondrath, "Design and PSpice simulation of synchronous bidirectional PWM DC-DC buck-boost converter operating in CCM," 2013 Annual IEEE India Conference (INDICON), 2013, pp. 1-6, doi: 10.1109/INDCON.2013.6725989.
- Siyang Zhao, Junming Zhang and Yang Shi, "A low cost low power Flyback converter with a simple transformer," Proceedings of The 7th International Power Electronics and Motion Control Conference, 2012, pp. 1336-1342, doi: 10.1109/IPEMC.2012.6258991.
- Sonst, M. (2021). "Application Note Verification Methods Of Snubber Circuits in Flyback Converters". Retrieved on 12 January 2023 from https://scdn.rohde-schwarz.com/ur/pws/dl_downloads/dl_application/application_notes/1sl363/1SL363_0e_SnubberVerificationMethods.pdf
- T. Brusev, "Buck-Boost DC-DC Converter Designed for PCB Applications," 2021 56th International Scientific Conference on Information, Communication and Energy Systems and Technologies (ICEST), 2021, pp. 103-106, doi: 10.1109/ICEST52640.2021.9483556.
- T. -F. Wu, Y. -C. Chen, J. -G. Yang and C. -L. Kuo, "Isolated Bidirectional Full-Bridge DC–DC Converter With a Flyback Snubber," in IEEE Transactions on Power

Electronics, vol. 25, no. 7, pp. 1915-1922, July 2010, doi: 10.1109/TPEL.2010.2043542.

T. Halder, "State Space Modeling and Stability Analysis of the Flyback Converter," 2020 IEEE 9th Power India International Conference (PIICON), 2020, pp. 1-6, doi: 10.1109/PIICON49524.2020.9113071.

T. Shamsi, M. Delshad, E. Adib and M. R. Yazdani, "A New Simple-Structure Passive Lossless Snubber for DC–DC Boost Converters," in IEEE Transactions on Industrial Electronics, vol. 68, no. 3, pp. 2207-2214, March 2021, doi: 10.1109/TIE.2020.2973906.

TDK Corporation (2013). Ferrites and Accessories, Retrieved on 18 January 2023 <https://www.tdk-electronics.tdk.com/download/519704/069c210d0363d7b4682d9ff22c2ba503/ferrites-and-accessories-db-130501.pdf>.

Todd, P. C. (1993, May). Snubber circuits: Theory, design and application. In *Unitrode-Power Supply Design Seminar* (p. 1993).

Trzynadlowski, A. M. (2015). *Introduction to Modern Power Electronics* (3rd ed.). New Jersey: Wiley.

Urgün Satılmış, Coruh Nasır, Erfidan Tarık, Oztürk Semra, "Flyback Dönüştürücü Tasarımı ve Analizi", 2009.

V. Ravi, C. Kuldeep and N. Lakshminarasamma, "High Voltage Flyback Converter for Pulsed Loaded Applications," 2020 IEEE International Conference on Power Electronics, Drives and Energy Systems (PEDES), 2020, pp. 1-6, doi: 10.1109/PEDES49360.2020.9379557.

Y. -F. Chen, T. D. Nguyen, J. -Y. Lin, Y. -C. Hsieh and H. -J. Chiu, "Hybrid-switching asymmetrical half-bridge flyback DC-DC converter," 2016 IEEE International Conference on Industrial Technology (ICIT), 2016, pp. 1313-1317, doi: 10.1109/ICIT.2016.7474945.

Yongsea Mo, Wonju Song, Ja-kyeong Lee, Yeji Han, S. Kumar and Hanjung Song,
"Design of a 5V-output boost DC-DC converter with improved protection
functions for battery-operated devices," 2017 International Conference on
System Science and Engineering (ICSSE), 2017, pp. 757-760, doi:
10.1109/ICSSE.2017.8030978.

