MIDDLE EAST TECHNICAL UNIVERSITY

EE464- STATIC POWER CONVERSION-II

HARDWARE PROJECT FINAL REPORT

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Introduction

In this project, we aimed to design a regulated and isolated DC-DC power supply using Forward Converter topology. This report contains detailed simulation results of our Forward Converter design, PCB design and 3D view of this design, component selection, bill of materials of component selection and thermal equivalent circuit of our design is provided. Basic circuitry and specifications of the Forward Converter is given below:

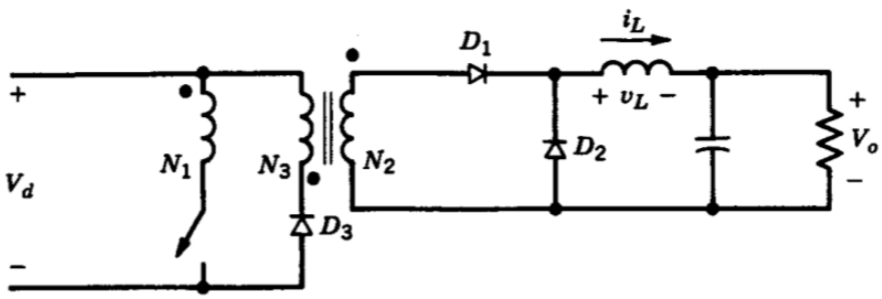

Figure 1: Forward converter circuit schematic

Table 1: Project Design Specifications of Forward Converter

Minimum Input Voltage	12V
Maximum Input Voltage	24V
Output Voltage	10V
Output Power	48 W
Output Voltage Peak to Peak Ripple	2 %
Line Regulation	2 %
Load Regulation	2 %

Considering that using Forward Converter topology instead of Flyback Converter increases cost, we aimed to come up with a more efficient converter instead of cheaper converter since using this topology already increases cost and it may not be possible to offer cheapest solution to the customers. Minimizing volume also was not our primary concern since the topology has transformer which makes its volume larger than non-isolated converter topologies such as Buck converter, Cuk converter.

Design Decisions

Since we aim to make our circuit as efficient as possible, some of the design decisions that we took in the Simulation Report are revised. We tried to make core loss and copper loss of magnetizing elements as close as possible since we assume it is the ideal and most efficient case. Semiconductor devices are also revised to minimize losses on them.

1) Magnetic Design

a) Transformer Design

Firstly, we selected our switching frequency. We selected it as 10 kHz. Although it is relatively small value for that kind of application, it decreases losses which is our main goal. Then we calculated our desired area product for the transformer considering our design specifications as follows, where P_{out} is output power, J is current density, K is topology constant, B_{max} is maximum flux density and f is frequency:

$$W_a A_c = \frac{P_{out} J}{K B_{max} f} = \frac{48 * 500}{0.0005 * 1300 * 10 * 10^3} = 3.69 cm^4$$

Since we desire to have a high permeability core for Forward Converter, ferrite cores fit better for our application. Considering these, 0P44022EC is selected. It is an E shaped, P material type ferrite core, whose area product is $4.59 cm^4$. Its geometric shape is given below:

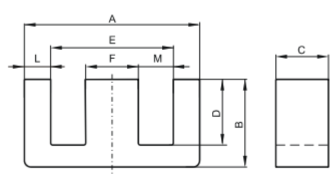

Figure 2: Geometric Shape of Transformer Core

After that, we calculated our required primary number of turns for the transformer as follows:

$$n_1 > \frac{V_{imax} * D_{max}}{f_s * B_{sat} * A_e} = \frac{24 * 0.5}{10^4 * 0.3 * 0.000233} = 17.167$$

n_1 is selected as 18. Considering we always need to stay below 0.5 duty cycle so that we do not saturate the core and considering additional voltage losses in realistic simulation, we calculated n_3 $\frac{n_3}{n_1} = 1.2$. As a result, n_3 is initially selected as 22. On the other hand, later simulation results with non-idealities showed that this turn ratio causes duty cycle to go over 0.5 in minimum input voltage, 24 V. As a result, it is adjusted as 40, according to simulation results.

n_2 is selected as 18 as well, equal to n_1. After that, appropriate cable sizes are selected for each winding for current density of $4A/mm^2$. Maximum input current is assumed to be 2.08 A and output current is assumed to be 5 A considering power and voltage ratings. As a result, AWG18 for primary side, AWG14 for secondary side and AWG25 for reset winding is selected. Fill factor is calculated to whether we can fit these cables inside the core or not as follows:

$$k_f = \frac{A_{cable}}{A_e} = \frac{((18 * 0.82 + 39 * 2.08 + 18 * 0.16)mm^2}{233mm^2} = 0.424$$

Fill factor is relatively small, on the other hand it shows that cables can fit into the core and having a large core can help us reduce core losses.

Copper losses are calculated with selected cable sizes. To do that, mean length per turn value was necessary, which is calculated as below for E shaped transformer:

$$MLT = \frac{\pi * (E - F)}{4} = 13.59mm$$

After that, primary and secondary winding resistances are found as follows:

$$R_1 = 20.9428 * 18 * 13.59 * 10^{-6} = 5.123mOhms$$

$$R_3 = 8.282 * 40 * 13.59 * 10^{-6} = 4.502mOhms$$

Using I^2R formula, copper loss is calculated as follows:

$$P_{copper} = I_1^2R_1 + I_3^2R_3 = 2.08^2 * 5.123 * 10^{-3} + 5^2 * 4.502 * 10^{-3} = 134mW$$

To calculate core loss, Steinmetz's Equation is used. It is specified as follows:

$$P_{CL} = \frac{a * f^x * B^y * L(T)}{1000} mW/cm^3$$

To use this equation, a, x and y constants, which are material specific, are needed. For P type material Magnetics provided these values as $a = 3.2$, $x = 1.46$ and $y = 2.75$. L(T) is also a function of ambient temperature and it is given as 1 for $100Celcius$ temperature. It is assumed to be 1 in the calculation:

$$P_{CL} = \frac{3.2 * (10 * 10^3)^{1.46} * 0.3^2.75}{1000} = 137.91mW/cm^3$$

Multiplying it with V_e of the core which is $5590mm^3$, we can find P_{core} as follows:

$$P_{core} = P_{CL} * V_e = 137.91 * 5.59 = 770.92mW$$

Lastly, to find magnetizing inductance of the transformer, following equation is used:

$$L_m = A_L N^2 = 5000nH * (40 + 18 * 2)^2 = 28.88mH$$

b) Inductor Design

At inductor design, we want to limit output current ripple which would also ensure continuous conduction mode. We want to keep the converter in CCM since forward converter's gain changes a lot in DCM. On the other hand, choosing a too large inductor would increase DCR value, which would reduce efficiency. We considered following equations for calculation of our output inductor considering 2 % ripple rating:

$$I_{L_O,min} = I_{O,max} * 0.02 = 4.8 * 0.02 = 0.096A$$

Ripple value must be less than twice of this value which is given below:

$$\Delta I_{L_o} = (\frac{n_3}{n_1} * V_i - V_o) * \frac{1}{L_o} * t_{on} < 0.192A$$

As a result, we need an inductor with $3.8mH$ inductance in order to meet criteria. This value is relatively high since our switching frequency is small. As a result, we needed to select a core with high inductance per turn square value. Using this catalog of Magnetics, 0W44715TC is selected as our inductor core. Its geometric shape is given below:

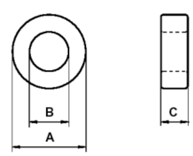

Figure 3: Geometric Shape of Inductor Core

Its A_L value is defined as $16100nH/T^2$ at 10 kHz, which is our switching frequency. Number of turns is calculated as follows:

$$N = 10^3 * \sqrt{\frac{L}{A_L}} = 10^3 * \sqrt{\frac{3.8mH}{16100nH}} = 15.36 \approx 16turns$$

Assuming it will carry more or less same amount of current with the secondary winding of the transformer, we can choose AWG14 cable for inductor as well.Fill factor according to these selection is calculated below:

$$K_f = \frac{A_{cond}}{W_a} = \frac{2.08 * 16}{145} = 0.230$$

After that, length of cable that is going to used for inductor is necessary to determine copper loss on the inductor.

$$MLT = (A - B) + 2C = 49.9mm$$

$$P_{copper} = I^2R = 4.8^2 * 8.282 * 16 * 49.9 * 10^{-6} = 317mW$$

To find core losses, Steinmetz's Equation is used once more. According to that, core loss is calculated below:

$$P_{core} = V_e * a * B^b * f^c = 16.063 * 10^{-3} * 31.320 * 0.3^{1.585} * 40^{1.371} = 44.26mW$$

2) Closed Loop Control

Since we desire to have a constant output voltage with variable input voltage, we need to have a closed loop controller. To do that, we used TL494 integrated circuit. Its pin output is given below:

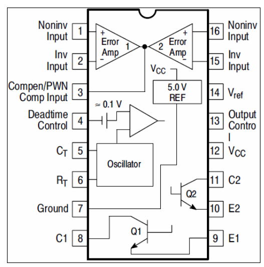

Figure 4: TL494 Pin Output Diagram

First thing to do while deciding our parameters was setting oscillation frequency equal to our selected switching frequency, which is 10 kHz. R_T and C_T values are calculated according to the following formula:

$$f_s = \frac{1}{R_T C_T}$$

C_T is selected as 1 nF and R_T is selected as 100 kOhms.
After that, we moved to Error Amplifier circuit in order to keep output voltage constant, whose schematic is given below:

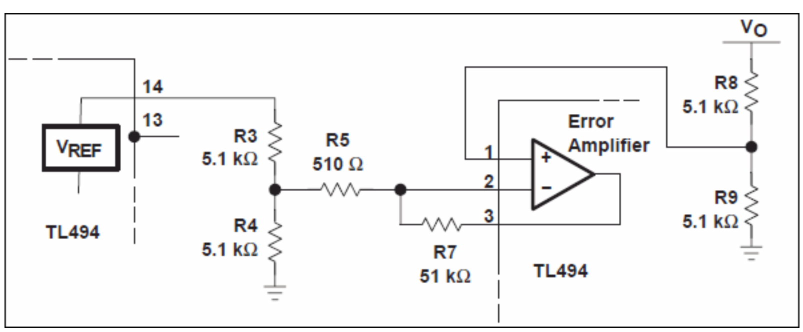

Figure 5: TL494 Error Amplifier Circuit

In this circuitry, R_3 and R_4 resistances create a reference voltage of 2.5 V. R_8 and R_9 enables us to sample our output voltage and R_5 and R_7 determines the gain of the amplifier. We decided to stick to resistance values that were given in the example circuit given above, since they meet our requirements. From the TL494 schematic, we can see that there are two error amplifiers in this circuit. However, the other amplifier is usually used for current limiting, which we do not want to have in this case. As a result, we decided to connect its positive input to supply voltage and negative input to ground, so that we never limit the current.

After that, we moved on to soft start future of TL494, which can be helpful in limiting stresses on the devices by avoiding high starting currents. Soft start circuit of TL494 is given below:

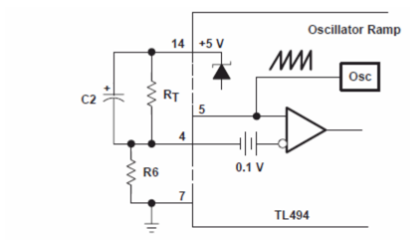

Figure 6: TL494 Soft Start Circuit

Soft start time is selected as 15 ms, which is 150 cycles period for 10 kHz frequency. According to that, C_2 is selected as 3.3 uF AND R_6 is selected as 4.7 kOhms.

TL494 is good at creating the PWM signal that we desire. However, it cannot drive a MOSFET directly with its output signal. Considering this, we decided to use an TLP250 digital optocoupler. It also creates isolation between high voltage part and low voltage part of the circuit, which is another advantage. TLP250 circuit is given below:

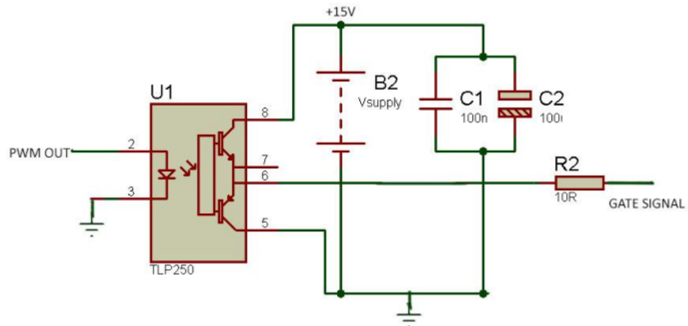

Figure 7: TLP250 Digital Optocoupler Circuit

We needed an another optocoupler in order to isolate output from the controller circuit. We decided to use TLP521 for this purpose. Its pin diagram is given below:

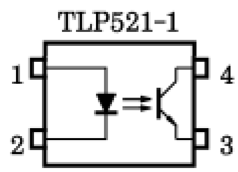

Figure 8: TLP521 Analog Optocoupler Pin Diagram

Simulation Results and Component Selections

Component selection that was done in the Simulation Report is revised since it was based on ideal Forward Converter Simulink model and hence voltage and current ratings of the components were too strict. As a result, new components are selected with higher ratings and their characteristics are put in the non-ideal Simulink model. Calculated magnetizing inductance and leakage inductance of the transformer, resistances of the windings, voltage drops on the semiconductors, resistance of the inductor and capacitor are added to the model and stress on the components are checked once again in order to verify that components can operate properly in the circuit. Simulink model of the converter is given below:

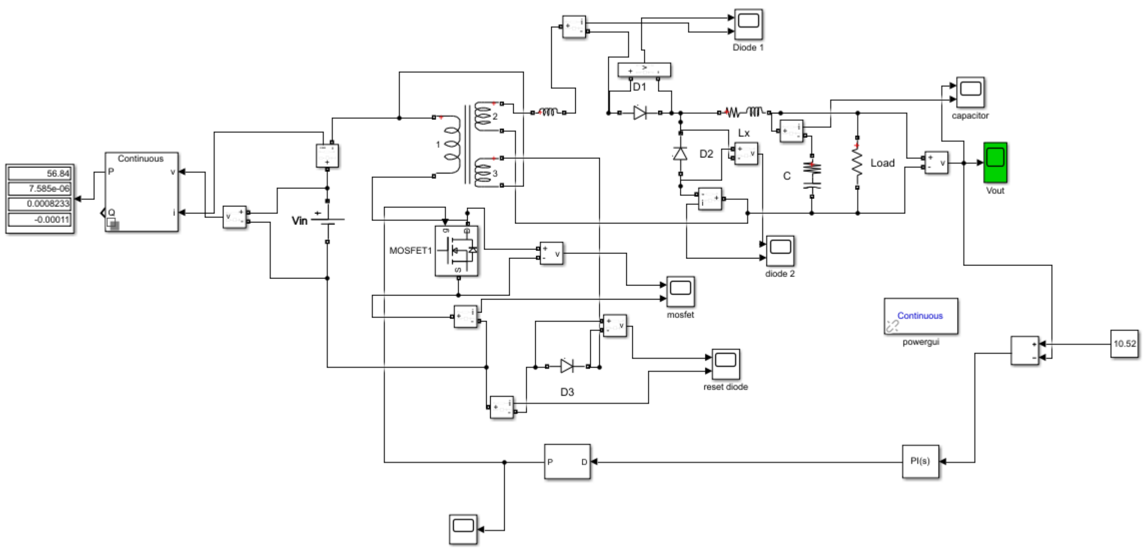

Figure 9: Simulink Model of the Forward Converter with Non-Idealities

The circuit is simulated for two extreme cases, having 24 V input and 48 V input. Output voltage waveform for these cases are given below:

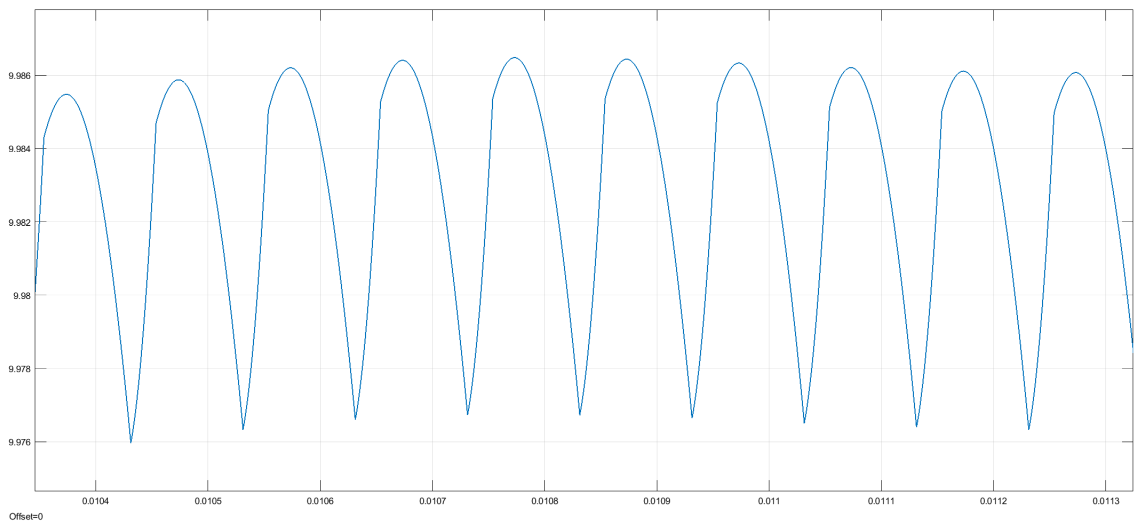

Figure 10: Output Voltage Waveform when Input Voltage is 24 V

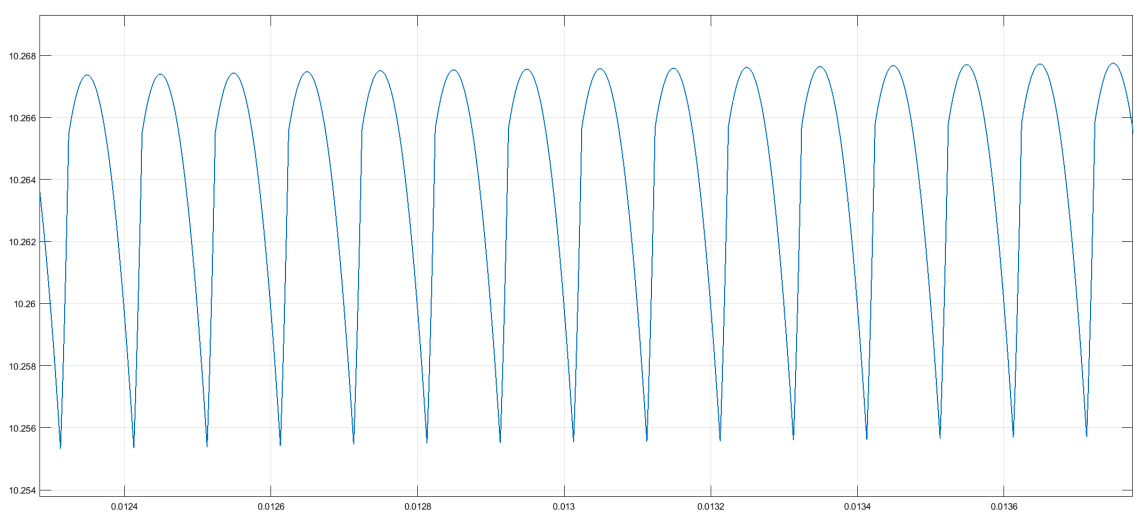

Figure 11: Output Voltage Waveform when Input Voltage is 48 V

For 24 V input, peak to peak voltage ripple is 0.0101 V, which corresponds to 0.1% and it is 0.01242 V for 48 V input which is also very close to 0.1% ripple percentage, easily meeting the criteria.

Input current waveforms are also given for both input voltage cases below:

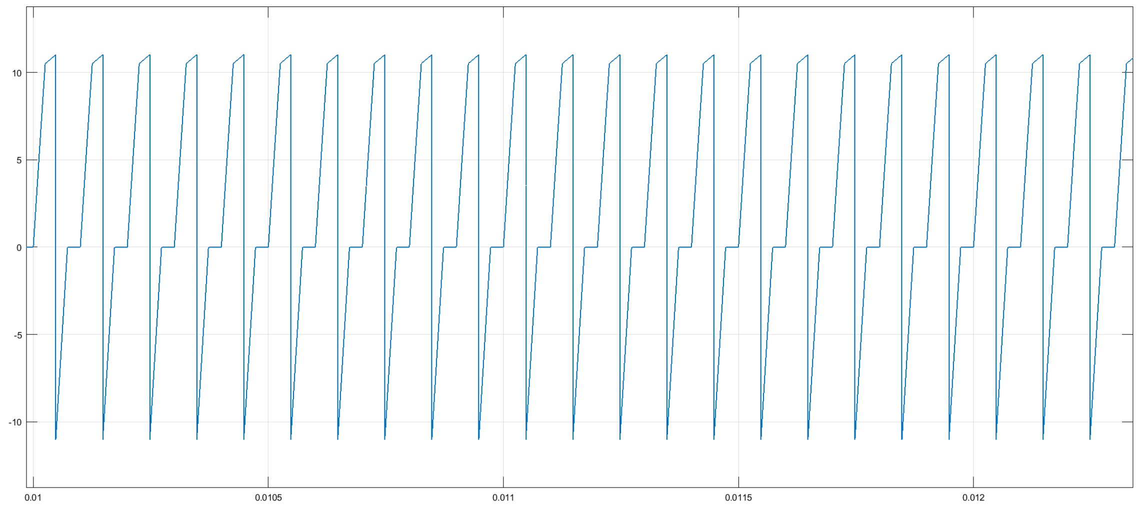

Figure 12: Input Current Waveform when Input Voltage is 24 V

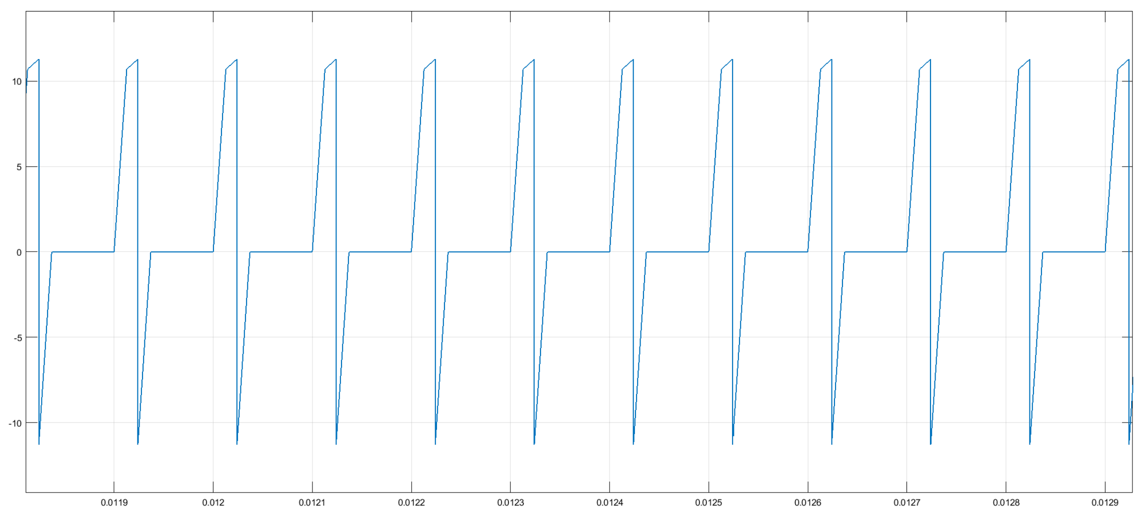

Figure 13: Input Current Waveform when Input Voltage is 48 V

After that, load resistance is changed from 2.08 Ohms to 20.8 Ohms to check voltage deviation with respect to previous case with 24 V input. It could not be done dynamically, changing resistance during the simulation, since variable resistor is not included in Power Systems library in Simulink. Output voltage waveform for this case is given below:

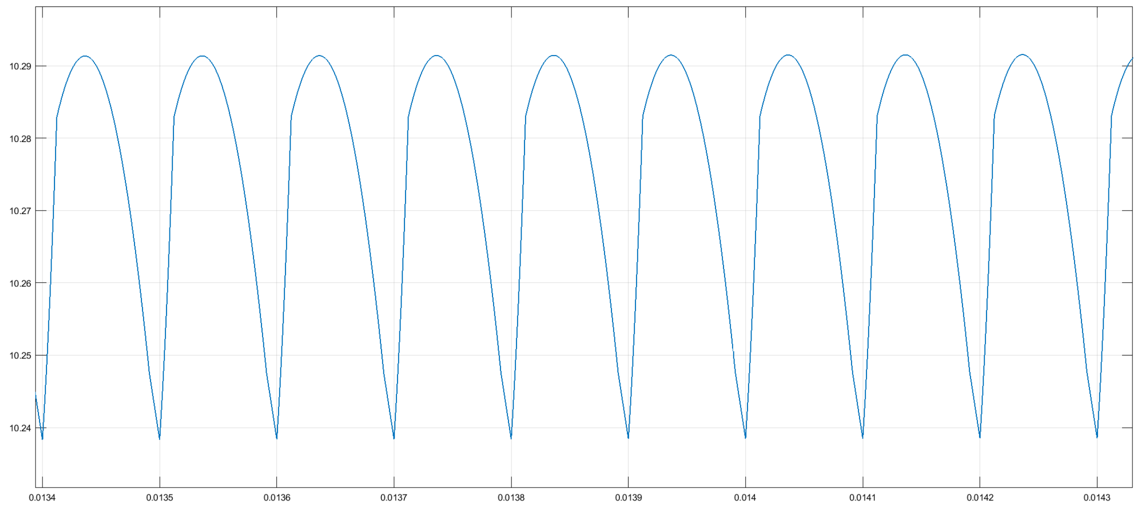

Figure 14: Output Voltage Waveform when Load Resistance is Increased

The RMS value of output voltage is decreased from 10.05 V to 10.02 V, meeting the load regulation criterion.

In order not to saturate the transformer core, we need to have maximum 50 % duty cycle at most. When the input is 24 V, we had our maximum duty cycle, which was 47.801 %. For 48 V input voltage, duty cycle was at 25.217 %. Graphs of duty cycle for both cases are provided below:

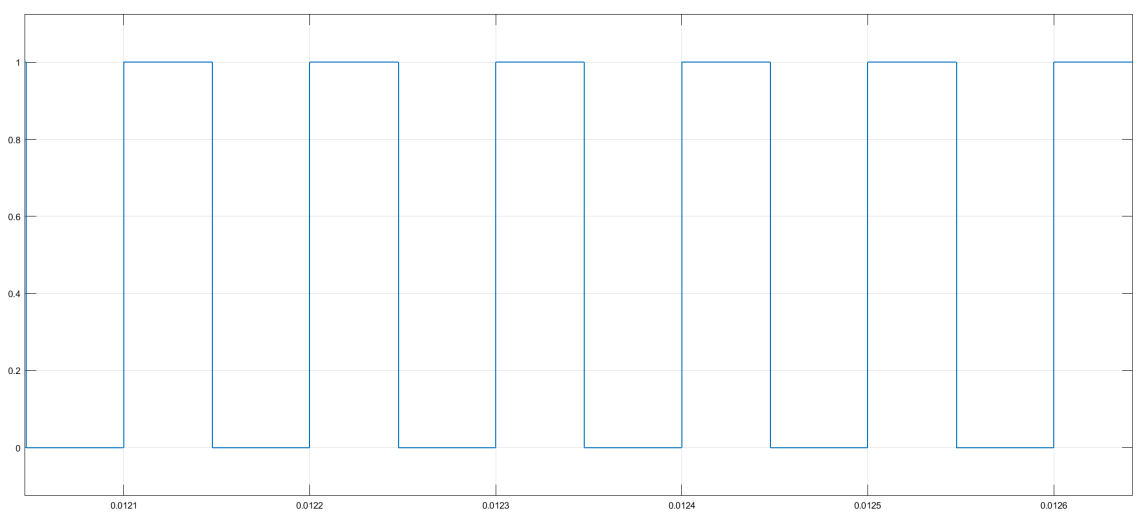

Figure 15: Duty Cycle Waveform when Input is 24 V

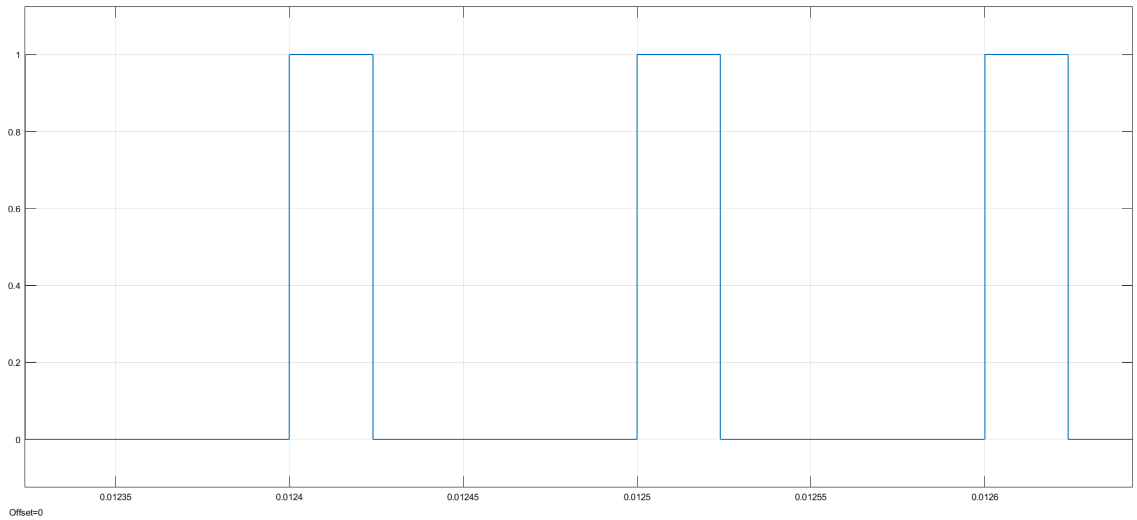

Figure 16: Duty Cycle Waveform when Input is 48 V

For MOSFET, IRF5Y31N20 is selected. It has 200 V drain to soure breakdown voltage, 18 A drain current, 1.3 V internal diode forward voltage and 0.092 Ohms R_{DS} resistance. It has 148 ns rise time and 27 ns fall time, which is fine for our switching frequency. It can handle the stresses and it also has low drain to source resistance which is good for the efficiency. Graphs of MOSFET voltage and current for both cases are given below:

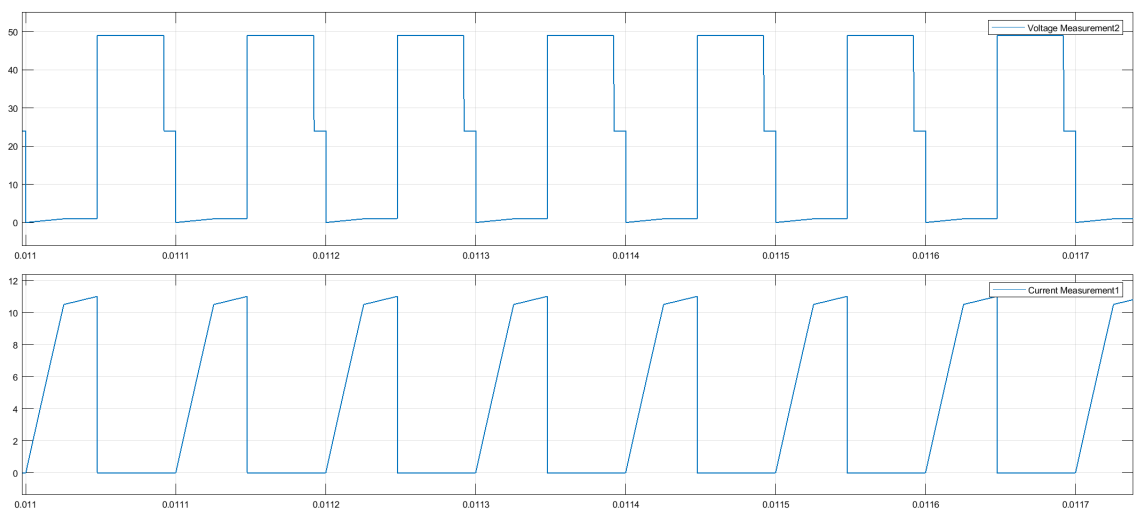

Figure 17: MOSFET Voltage and Current when Input is 24 V

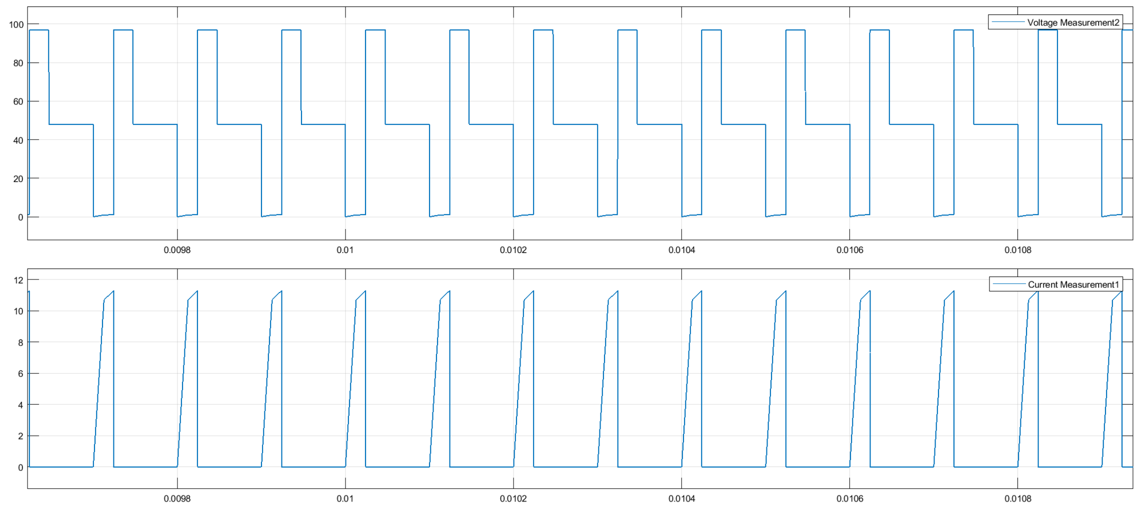

Figure 18: MOSFET Voltage and Current when Input is 48 V

It carries 11 A peak current and its maximum voltage is less than 100 V. As a result, it can handle these stresses. Its picture is also given below:

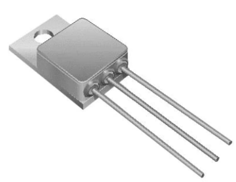

Figure 19: Picture of the MOSFET

For reset diode, SFAF503GHC0G is selected. It has 150 V blocking voltage and 0.975 V voltage drop on it. It can carry up to 5 A average current and 125 A peak current with 35 ns maximum reverse recovery time. Graphs of diode voltage and current for both cases are given below with the picture of diode:

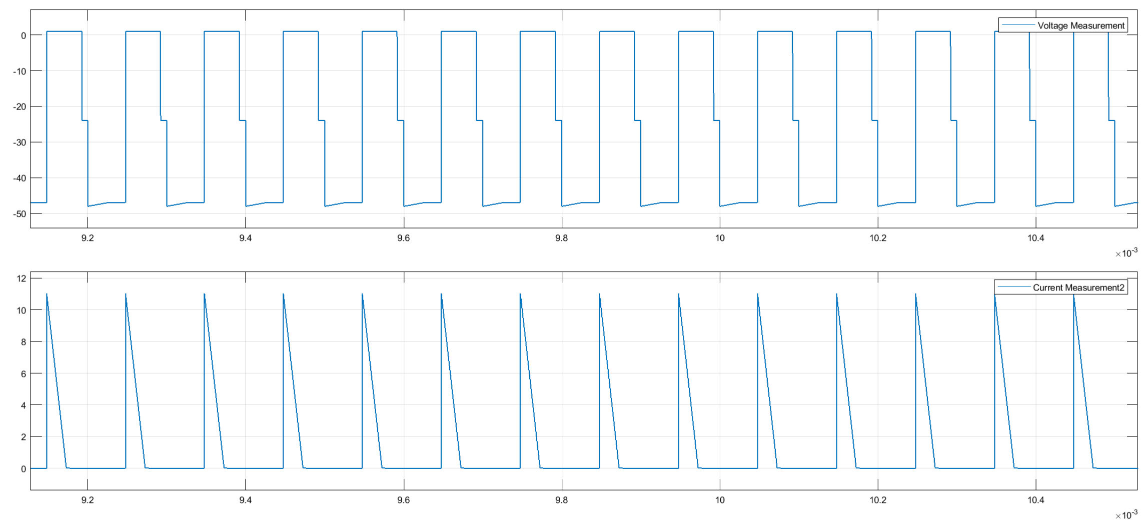

Figure 20: Reset Diode Voltage and Current when Input is 24 V

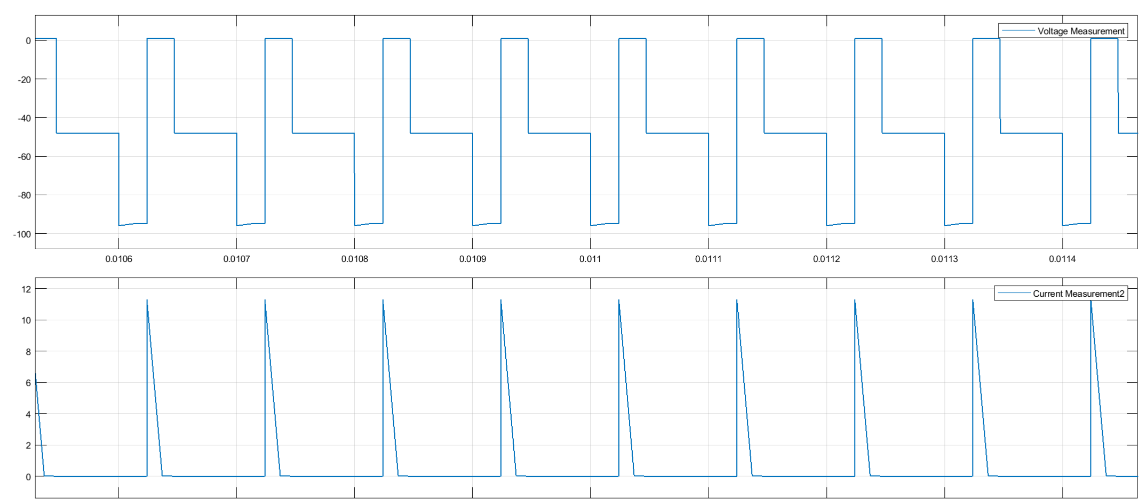

Figure 21: Reset Diode Voltage and Current when Input is 48 V

Figure 22: Picture of the Reset Diode

For secondary side diodes, SRA10150 C0G is selected. Since it is Schottky diode, it has no recovery problem. It has 0.7 V voltage drop on it. Graphs for secondary side diode voltage and current for both cases and picture of the diode are given below:

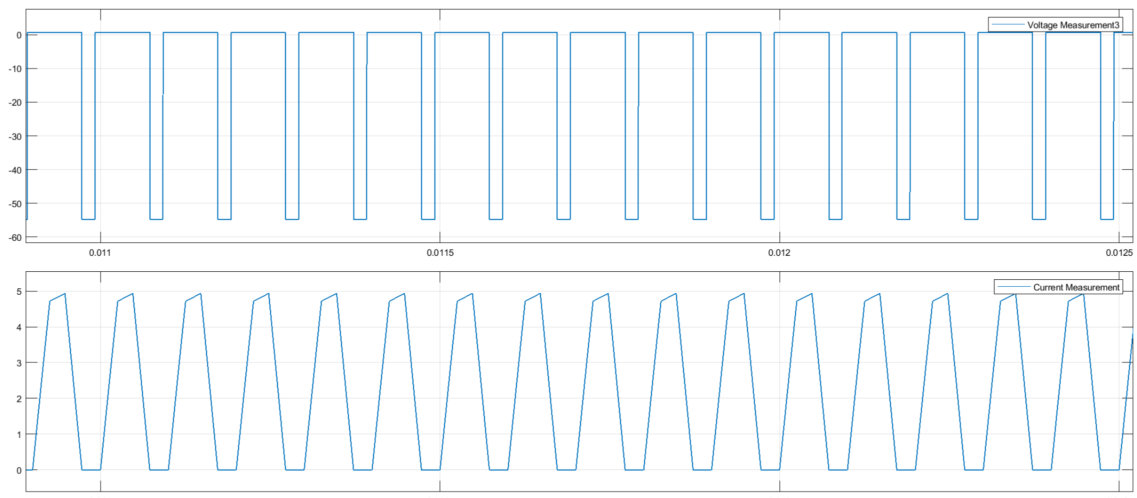

Figure 23: Secondary Diode Voltage and Current when Input is 24 V

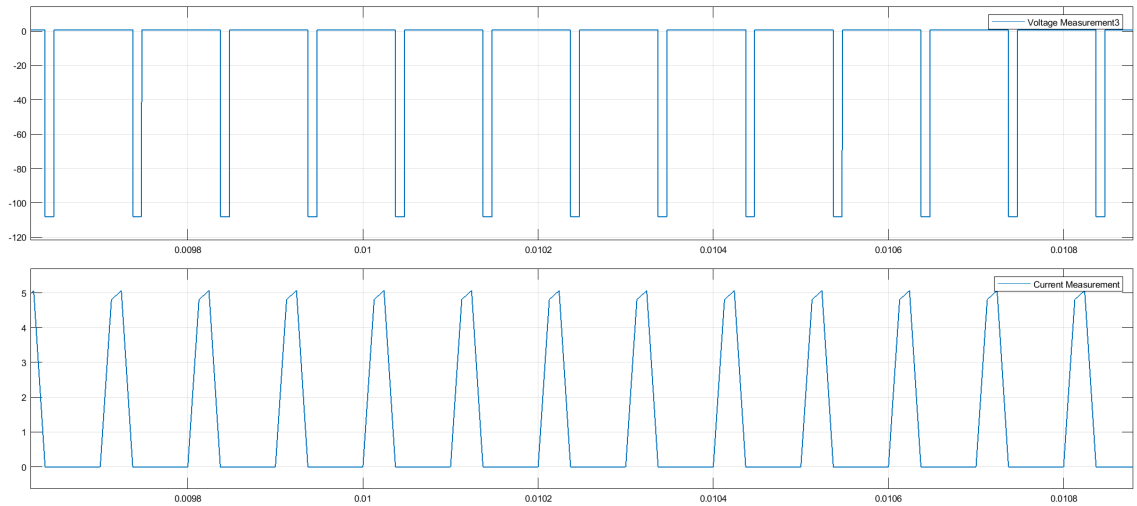

Figure 24: Secondary Diode Voltage and Current when Input is 48 V

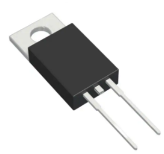

Figure 25: Picture of the Secondary Side Diode

For output capacitor, ESY477M035AH4AA is selected. It is a 470 uF, 35 V capacitor. Although its capacitance is more than we need to achieve our output voltage ripple aim, we chose it due to its small ESR resistance, which is defined as 26 mOhms for 100 kHz. Assuming electrolytic capacitor will have higher ESR value at lower frequency, it is taken as 40mOhms for simulations. Its picture is given below:

Figure 26: Picture of the Output Capacitor

Efficiency

To calculate efficiency from the simulation model, Power Measurement block is used by measuring input and output current and voltages. However, core losses are not included in the simulation model. Hence, their calculated values in Design Decisions section using Steinmetz's Equation are used for transformer and inductor.

Firstly, efficiency is tested for 24 V input and 48 V input cases. After that, different load resistance values are used. Results are provided below:

$$\eta_{24V} = \frac{P_{out}}{P_{in} + P_{core}} = \frac{48.62}{57.05 + 0.81} = 84.03\%$$

$$\eta_{48V} = \frac{50.72}{56.91 + 0.81} = 87.87\%$$

$$\eta_{1.04Ohms} = \frac{94.45}{185.3 + 0.81} = 50.74\%$$

$$\eta_{4.16Ohms} = \frac{24.04}{27.96 + 0.81} = 87.05\%$$

$$\eta_{6.24Ohms} = \frac{16.69}{18.37 + 0.81} = 87.05\%$$

$$\eta_{8.32Ohms} = \frac{12.55}{13.72 + 0.81} = 86.37\%$$

As a result, we can say that efficiency increases in higher load resistances since we draw less current from the input side, meaning less losses on the resistive elements and smaller switching losses.

Thermal Design

Thermal design is the one of the important part of design. For thermal design we have to model the system as electrical circuit. Electrical model of thermal heat flow is shown below. We assume that ambient temperature is 50C.

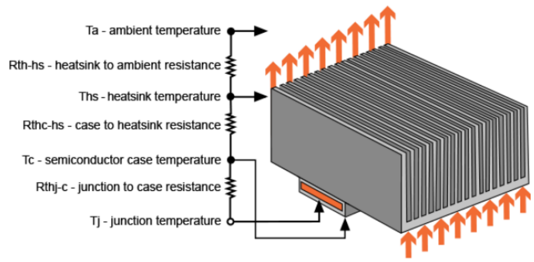

Figure 27: Thermal modelling

0.1 Mosfet Thermal Design

Mosfet heating rapidly due to the switching frequency. Mosfet heat transfer equivalent modelling shown below.

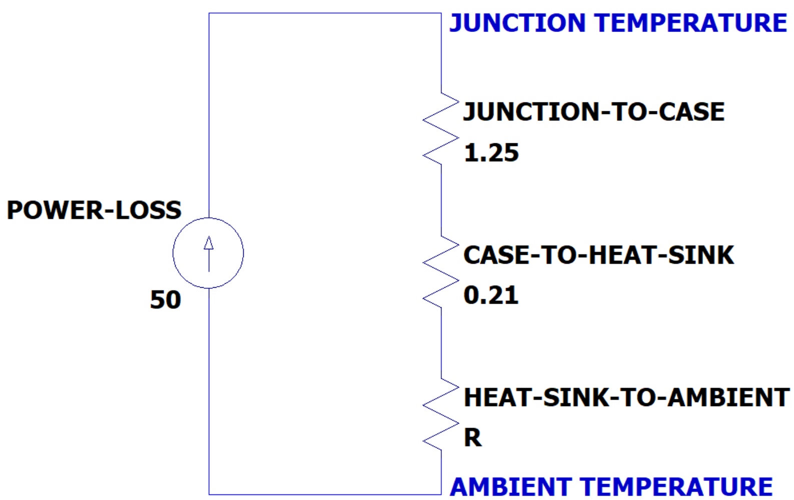

Figure 28: Mosfet Thermal Modelling

$$R_{eq} = \frac{JunctionTemperature - AmbientTemperature}{PowerLoss} = \frac{130-50}{50} \tag{1}$$

$$R_{eq} = 1.6 \tag{2}$$

Heat sink to ambient is 1.6-1.25-0.21=0.14 C/W. Considering this, 577102B00000G is selected as heatsink for MOSFET, whose thermal resistance is 0.20 C/W. Its package is TO220.Our MOSFET's package is TO-257AA, which is sutiable for TO220 thermal sinks to use. Its picture is given below:

0.2 Diode Thermal Design

Diode's heat do not increase as much as as MOSFET. Diode thermal equivalent model is shown below.

Figure 29: Picture of MOSFET Heat Sink

$$R_{eq} = \frac{JunctionTemperature - AmbientTemperature}{PowerLoss} = \frac{130 - 50}{25} \tag{3}$$

$$R_{eq} = 3.2 \tag{4}$$

Heat sink to ambient is 3.2-2.25=0.95 C/W. According to this, 530002B02500G is selected for diodes. Package in this case is also TO220. Its picture is shown below:

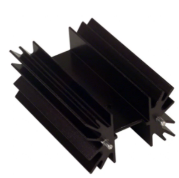

Figure 30: Picture of the Diode Heat Sink

PCB Design, Bill of Materials and 3D View

In PCB design we used Proteus design software. We chose Proteus because it has huge library and we worked with Proteus before.The package of chosen component are available on Proteus library. But transformer of the forward converter is not available on the library so there are holes for transformer block. We design inductor and we will make our inductor with our hands so there is no package. Our design shown below.

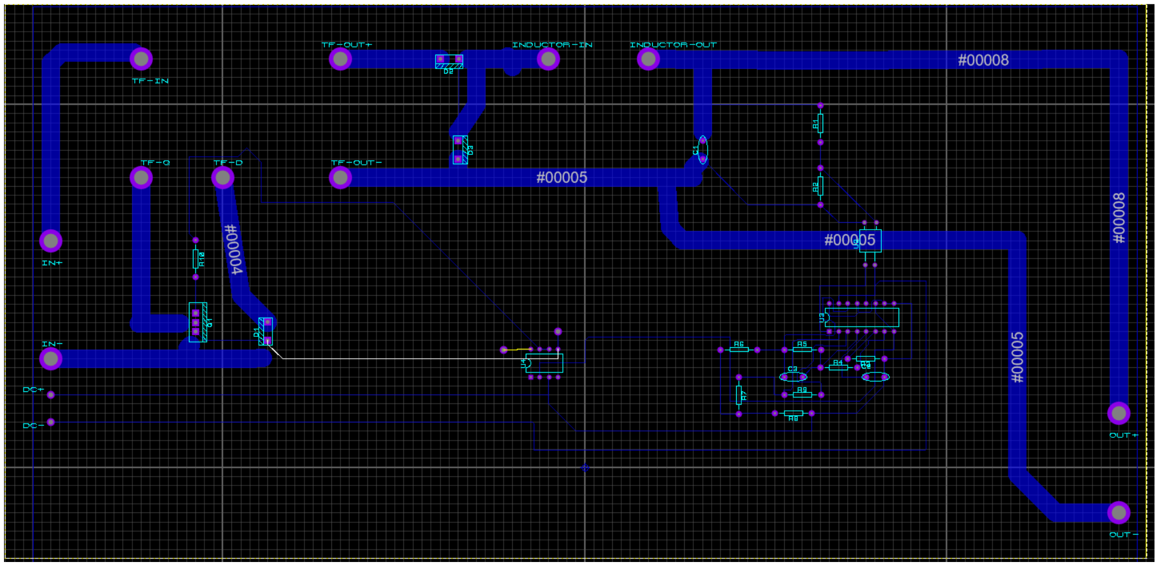

Figure 31: PCB design of forward converter

For calculating we use Digikey PCB with calculator. For high power application I choose $4oz^2$. 1 oz is 23.35g and $1ft^2$ is $0.093m^2$ so our max average current is 10 A so 5mm trace with ideal for design.Also Gerber files are uploaded to GitHub. 3D pc designs are shown below in figures.

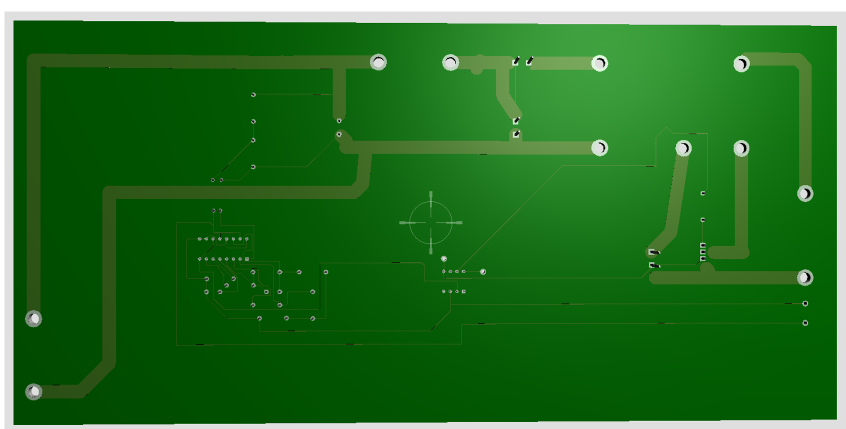

Figure 32: Back view of the design

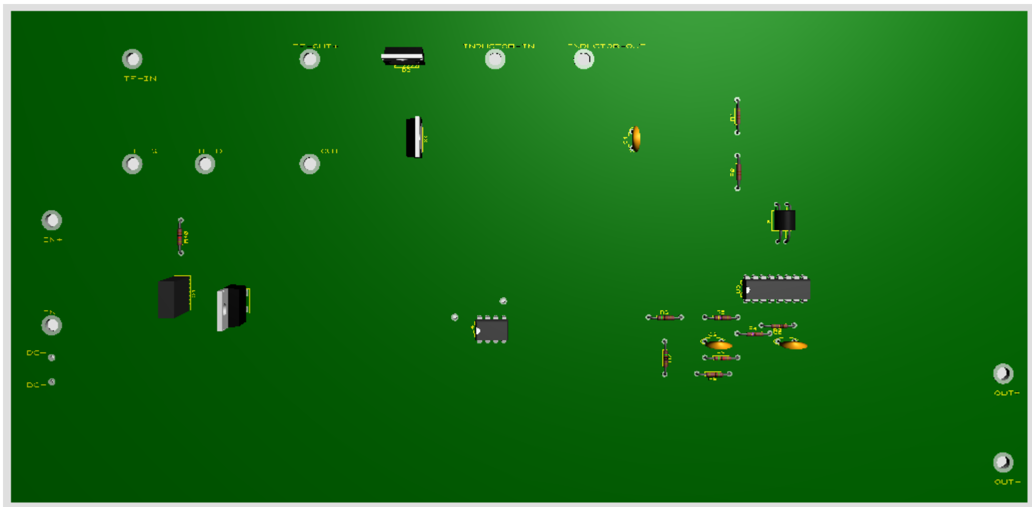

Figure 33: Top view of the design

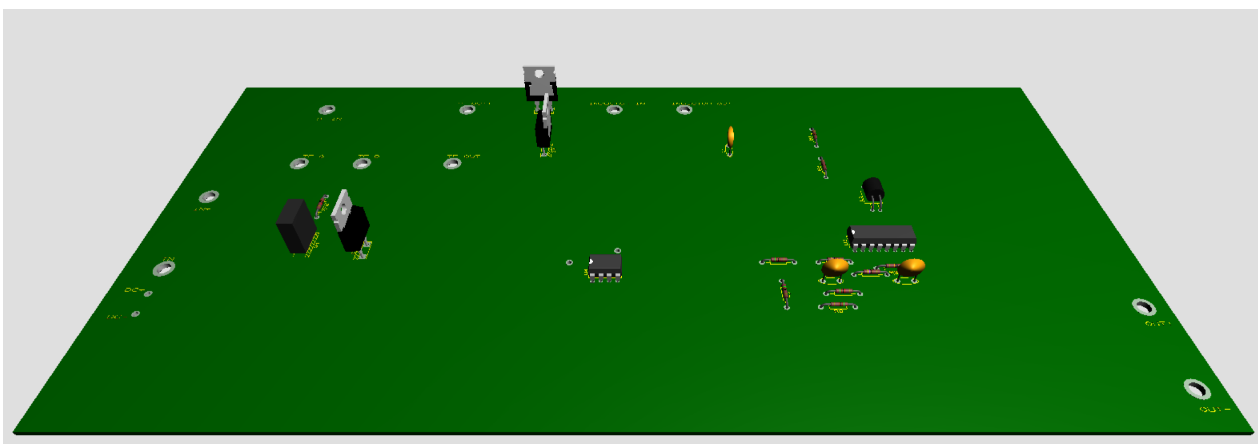

Figure 34: Front view of the design

Bill of materials are shown below figure

Bill Of Materials for DCP-464HARDWARE

Design Title	DCP-464HARDWARE
Author	
Document Number	
Revision	
Design Created	23 Haziran 2020 Salı
Design Last Modified	23 Haziran 2020 Salı
Total Parts In Design	22

0 Modules			
Quantity	References	Value	Unit Cost
Sub-totals:			₺0,00
3 Capacitors			
Quantity	References	Value	Unit Cost
1	C1	1p8	₺3,60
1	C2	1n	₺0,50
1	C3	3.3u	₺0,50
Sub-totals:			₺4,60
10 Resistors			
Quantity	References	Value	Unit Cost
2	R1-R2	5.1K	₺0,10
1	R3	100k	₺0,10
2	R4-R5	4.7K	₺0,10
2	R6,R8	5.1k	₺0,10
1	R7	510	₺0,10
1	R9	51k	₺0,10
1	R10	10k	₺0,10
Sub-totals:			₺1,00
3 Integrated Circuits			
Quantity	References	Value	Unit Cost
1	U2	NSL-32SR3	₺1,60
1	U3	TL494	₺3,20
1	U4	6N137	₺12,10
Sub-totals:			₺16,90
1 Transistors			
Quantity	References	Value	Unit Cost
1	Q1	IRF5Y31N20	₺5,25
Sub-totals:			₺5,25
3 Diodes			
Quantity	References	Value	Unit Cost
1	D1	18TQ045	₺3,11
2	D2-D3	18TQ045	₺4,20
Sub-totals:			₺11,51
2 Miscellaneous			
Quantity	References	Value	Unit Cost
1	L1	3.8mH	₺6,40
1	TR1	TRAN-2P3S	₺7,80
Sub-totals:			₺14,20
Totals:			₺53,46

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Figure 35: Front view of the design

Conclusion

In this report, our hardware design specifications for Forward Converter are described. Although our main target was implementing this design and having an actual product on our hands at the end of the process, this was not possible due to recent pandemic.

Nevertheless, our alternative approach for this project was really helpful since the project was covering a lot of steps that need to be done before creating an actual product in professional engineering. Using a more detailed simulation was teaching since as simulation gets more detailed, it gives us a better picture about what would actually happen if we tried to implement the project. We gained experience in PCB design which can be very helpful in professional engineering life since no customer wants to buy something built on stripboard. Using Bill of Material and 3D Viewer tools of the PCB design software gave us a better understanding about what would product actually look like how much it would cost more or less. Since these are critical in engineering product, they were also very beneficial. It was a useful experience covering lots of aspects of professional power electronics engineering.

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