ELECTRICAL AND ELECTRONICS ENGINEERING DEPARTMENT

EE464 POWER ELECTRONICS – II FINAL REPORT

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1. Introduction

In this project, we were expected to design an isolated DC-DC converter that meets the set requirements. For this purpose, we first look at the requirements. The designed circuit should satisfy the requirements below.

- Minimum Input Voltage: 220V
- Maximum Input Voltage: 400V
- Output Voltage: 12V
- Output Power: 100W
- Output Ripple: 4%
- Line Regulation: 3%
- Load Regulation: 3%

We assess the requirements carefully and then start the design process. First, a suitable topology is selected by discussing the advantages or disadvantages of each topology. After determining the topology, the controller IC options are discussed and the most suitable one is selected. Circuit design process is started with controller IC and other circuit parameters are selected by considering the limitations and requirements of the controller, which are stated in the datasheet.

After determining the circuit parameters, we conduct some simulations. According to the simulation results, some changes are made and commercial products are selected for some important power components like switches, capacitors and inductances. Also, magnetic design for the transformer is done after assessing the requirements of the circuit.

After that, we do some thermal calculations to see whether the components need any additional cooling. We then design the PCB to meet the set requirements. Finally we use software to thermally analyze the completed product.

2. Topology Selection

To design this isolated DC/DC converter, we start by choosing a topology that satisfies the needs of our project. The first and most critical requirement is that this converter had to have the input and output isolated from each other. This means that we are limited to isolated topologies. The topologies that we have considered for the design are the flyback, forward and push-pull converter topologies. There are certain advantages and disadvantages for using each of these converters.

The advantage of the flyback converter is that it uses the transformer core as an energy storage element, making an extra inductor unnecessary. This situation makes flyback a cheaper and simpler option. The disadvantage of the flyback is that the transformer needs to be larger compared to other topologies in order to avoid high core losses. The flyback is not suited for high power conversion and the required 100W is certainly not low but falls just in range of the recommended power of a flyback. We must also keep in mind that the output current will exceed an average of 8A and have a RMS value even higher. Due to this the component in charge of rectification at the output will dissipate a lot of heat if it happens to be a diode. Due to this synchronous switching is preferred.

Also it is worth noting that the group members have experience with flyback converter design from their summer practices and part-time working activities, which makes the flyback converter design process much easier. Therefore, a flyback converter is selected to be the primary design option for this project.

While the topologies themselves are important, we must also consider the available IC controllers on the market while making our decision since if we do not have a proper IC for our topology the design is pretty much impossible. There are certain large IC controller manufacturers that have simulation models for their ICs like Analog Devices and Texas Instruments which would make our design process much easier. Therefore, controller IC selection holds a crucial role in converter design.

3. Controller IC Selection

After researching the conventional controllers, some ICs are determined to be possible solutions and their advantages and disadvantages are further discussed. The first option available was the InnoSwitch3-EP.

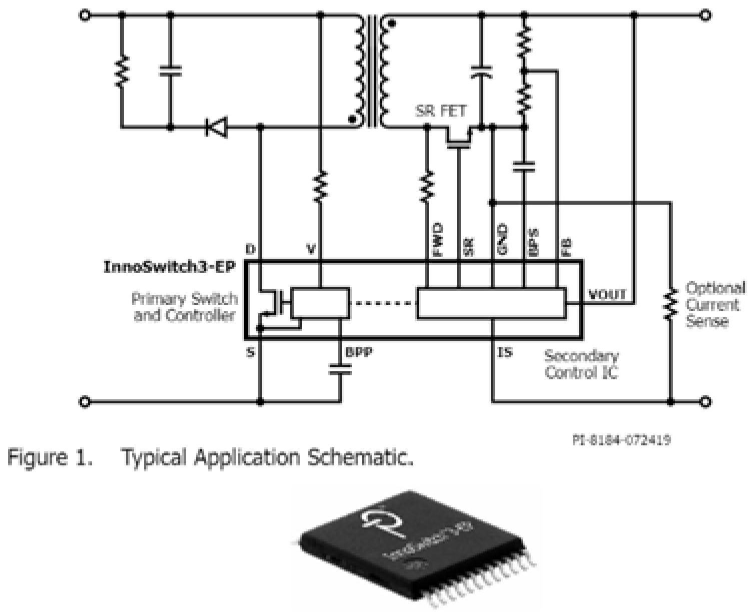


Figure 1 Alternative controller IC - InnoSwitch3-EP

This controller has an internal GaN FET, which increases the switching efficiency while also decreasing required parts. Also, it provides a synchronous drive for the secondary in a single chip. According to the datasheet, no heatsink is required up to 100W. However, there is no simulation model provided for this controller and some critical details might be missed if it was manually modelled. Therefore, this controller is discarded as an option.

Other two options are LT8316 and LTC3805 manufactured by Analog Devices. Since these devices have built-in models in LTSpice, it is logical to use ICs manufactured by AD. Both of these ICs are used for conventional Flyback Converters with the difference between them being the control and feedback mechanisms. The LTC3805 is a current mode controller with an adjustable frequency between 70kHz and 700kHz. For isolated feedback, this controller must be used in tandem with an opto-coupler driver and an opto-coupler. The advantage of this is that the load and line regulations are essentially perfect, and the frequency of the controller is constant. The main disadvantage is that it requires too many components for too few benefits.

The LT8316 on the other hand is a Quasi-resonant boundary operation mode controller and takes its feedback from the third winding of the transformer. The operation frequency of this controller changes depending on the load and while it has very good load and line regulation compared to other no-opto controllers, it does not give the same performance as a controller with an opto-coupler. The main advantage of the LT8316 is that it uses less parts since there is no need for an opto and also that it can be naturally started and then fed from the input and third windings. Both ICs are simulated, and it is seen that both of them satisfy the given conditions and are both suitable designs but in the end, we must decide on one of them to continue with our project. Note that by using a Secondary-Side Synchronous Rectifier Driver or more specifically the LT8309, we can get rid of the output diode and replace it with a MOSFET with low Rds to dramatically increase efficiency. Also, this IC works without any feedback from the primary side. Keeping this in mind we decide to use the LT8316 in conjunction with the LT8309 to achieve an efficient and simple system.

4. Transformer Design

Flyback converter isolates a primary side and a secondary side with a transformer. Transformer can be used to step up or down an input voltage according to the turns ratio. Besides, the input-output relation is the same as a buck boost converter. Input-output relation is shown in Equation (1).

	$V_o = \frac{N_s}{N_p}\frac{D}{1-D}V_{in}$	(1)

Transformer is mainly used for the applications transmitting energy without storage. However, the transformer of the flyback converter stores energy, and then passes it to the secondary side, which is the reason for why the flyback transformer is also called a coupled inductor. Hence, considerations for energy stored, saturation current, inductance are important for transformer design. The design of the transformer is started with application information given in the datasheet.

Table 1 shows some system level information and controller requirements having an impact on the transformer design.

Table 1 Parameters for the transformer design

Description	Symbol	Value
Input voltage range	$V_{in,min}$ - $V_{in,max}$	220V – 400V
Output voltage	V_o	12V
Max. output power	P_{out}	100W
Max. switching frequency	f_s	140 kHz
Min. Switch off time*	$t_{OFF(MIN)}$	800ns
Min. Switch on time*	$t_{ON(MIN)}$	300ns
Backup time*	t_{BU}	50μs

** These parameters will be explained in detail while calculating their limitations on primary inductance.*

Application information given in the datasheet determines the minimum and maximum limits for the primary, magnetizing, inductance due to parameters given in Table 1 such as $t_{OFF(MIN)}$, $t_{ON(MIN)}$ and t_{BU}. Before finding the limitations, the peak of the primary current should be determined according to the working principle of the controller and the delivered maximum output power. While the minimum current limit of the controller affects the limitations of primary inductance due to $t_{OFF(MIN)}$ and $t_{ON(MIN)}$, the maximum output power has to be satisfied according to magnetic energy storage shown in Equation (2).

$$E_{mag} = \frac{1}{2}Li^2 \tag{2}$$

The maximum peak current in the primary side can be found from the desired output power. Since the peak current is observed at maximum power and the controller has boundary mode control, the peak current can be calculated as seen in Equation (3). Efficiency is taken as %90 as a safety margin.

$$I_{peak} = \frac{2P_{out}}{\eta V_{in} D} = 3.37A \quad (3)$$

$$V_{in} = 220V,\ \eta = 0.9,\ D = 0.3,\ P_{out} = 100W$$

4.1. Limitations of Primary Inductance

I. $t_{OFF(MIN)}$ – Minimum Switch Off Time

The feedback of the output voltage is given through the tertiary winding. The nonzero voltage between the tertiary winding is created when the secondary current flows. The required minimum time to sample voltage by the sample-and-hold error amplifier is 800ns. "*In order to ensure proper sampling, the secondary winding needs to conduct current for at least 800ns*." [1] The limitation due to $t_{OFF(MIN)}$ is given in Equation (4).

$$L_m > \frac{t_{OFF(MIN)} N(V_o + R_{ds} I_{o(RMS)})}{I_{SW(MIN)}} = 0.12\text{mH} \quad (4)$$

$$N = 8,\ R_{ds} = 5m\Omega,\ I_{SW(MIN)} = 0.70A,\ I_{o(RMS)} = 13A$$

II. $t_{ON(MIN)}$ – Minimum Switch Off Time

The controller has a minimum switch on time for the sake of blanking initial switch turn-on current spike. The current, $I_{SW(MIN)}$, should not be reached for $t_{ON(MIN)}$, which creates restrictions on the selection of the flyback transformer in terms of magnetizing inductance. The limitation due to switch on time is determined in Equation (5).

$$L_m > \frac{t_{ON(MIN)} Vin(MAX)}{I_{SW(MIN)}} = 0.17\text{mH} \quad (5)$$

III. Power Storage Capability of the Transformer

Equation (2) shows the magnetic energy stored in an inductor for a given current and inductance. Since the flyback transformer stores magnetic energy due to the working principle of the flyback converter, the inductance of the transformer should be able to store sufficient power. Equation (6) shows the limitation of the inductance due to magnetic energy storage.

$$L_m > \frac{2(V_o + R_{ds} I_{o(RMS)})I_{o(RMS)}}{\eta I_{SW(MAX)} f_{s(MAX)}} = 0.15\text{mH} \quad (6)$$

IV. Backup Time

Backup timer is stimulated to avoid low output voltage levels. Backup timer turns the primary switch on unless the secondary side diode/switch turns off. This feature restricts the inductance value with maximum limit. Limitation of inductance of the flyback transformer due to backup time can be seen in Equation (7).

$$L_m < \frac{0.8N(V_o + R_{ds} I_{o(RMS)})t_{BU}}{I_{SW(MAX)}} = 1.1\text{mH} \quad (7)$$

4.2. Determination of Turns Ratio

The output voltage is lower than the input voltage level. Transformer should step down the input voltage for reasonable values so that sensitivity of duty cycle should not be high, i.e., duty cycle below %10 is a problem for controllers. Therefore, the turns ratio should be as high as possible to keep the duty cycle in a range between %50 and %20.

Turns ratio also has an impact on breakdown voltage of semiconductors. Table 2 shows maximum terminal voltages of the primary switch and the secondary switch. The turns ratio has a reverse effect on the semiconductors. While the turns ratio, N, increases reverse voltage of the primary switch, it reduces reverse voltage of the secondary switch.

Table 2 Relation between turns ratio and reverse voltage of the semiconductors

Semiconductor	**General Formulation for Reverse Voltage**
Primary switch	$V_{DS} = V_{in} + N(V_o + R_{ds}I_{o(RMS)}^2) + V_{Leakage}$
Secondary switch	$V_R = (V_{in}/N) + (V_o + R_{ds}I_{o(RMS)}^2) + V_{Leakage}$

The turns ratio is selected as 8, which keeps the duty cycle between %20 and %30. Also, this selection can be optimized in terms of reverse voltage of semiconductors and duty cycle. The increase in reverse voltages results in increase in switching losses and decrease in safety margin of devices. Duty cycle of the flyback converter in a range of input voltage is shown in Figure 2.

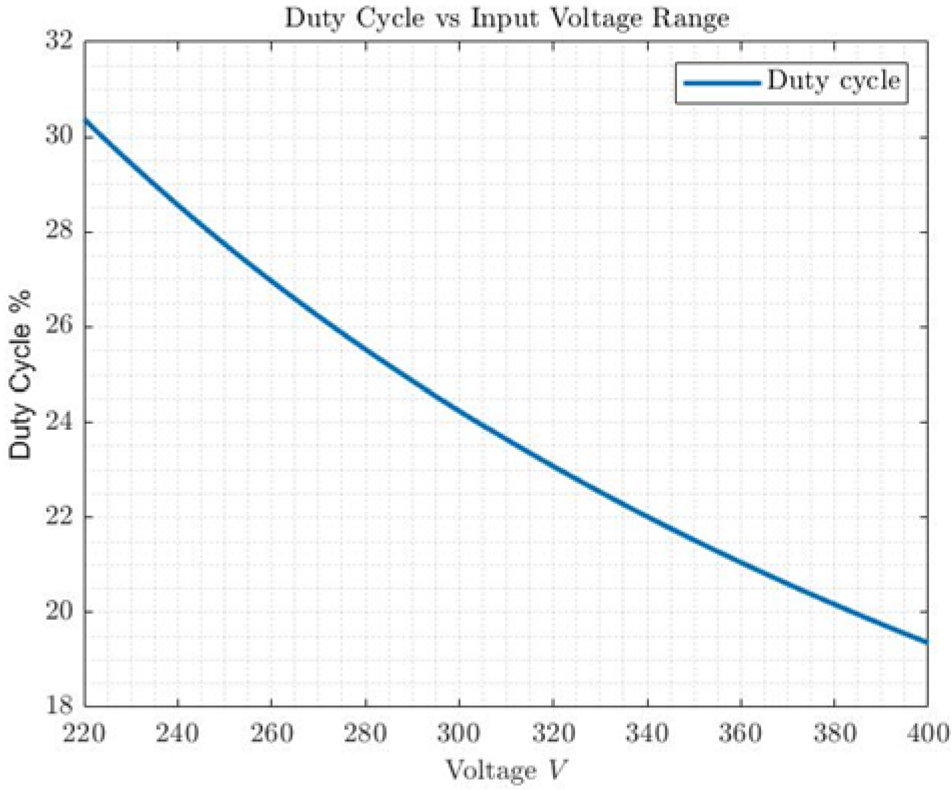


Figure 2 Duty cycle vs input voltage range for N=2

4.3. Determination of the Primary Inductance

In the part 4.1, 0.17mH and 1.1mH are found as minimum and maximum limitations, respectively. The value between the limits may probably work. However, the safety margin for each limit should be determined to eliminate unexpected practical issues. The safety region for each limit can be taken as %25 of limiting value above and below for minimum and maximum limits, respectively. Hence, the primary inductance value should be between 0.2mH and 0.81mH.

The boundary mode control requires variable switching frequency according to the load. When the primary switch is on, the inductor current starts from the zero. Therefore, the frequency relation between frequency and inductance for the same input voltage and output voltage is obvious, which is shown in Equation (8). Since duty cycle and turns ratio is constant, switching frequency and primary inductance are inversely proportional for the same amount of inductor current ripple. The inductor current ripple is determined according to output power.

$$\Delta I_{L,ON} = \frac{DT_s V_{in}}{L_m} \qquad (8)$$

$$\Delta I_{L,OFF} = \frac{(1-D)T_s(V_{in}+N(V_o+R_{ds}I_{o(RMS)}))}{L_m}$$

Transition to DCM occurs as the load decreases. Duty cycle calculation in DCM can be seen in Equation (9). Both calculations are the same except the P_{out}, V_o, and R relation.

$$D = \frac{Vo}{V_{in}}\sqrt{\frac{2L_m f_s}{R}} \qquad (9)$$

$$D = \sqrt{\frac{2P_{out}L_m f_s}{V_{in}}}$$

Supply of the input power occurs during the on time of primary switch. Therefore, the current ripple of the inductor is directly related to the delivered power. Equation (10) shows the inductor current ripple calculation in DCM, which shows that L_m and f_s are also inversely proportional in DCM.

$$\Delta I_L = \frac{V_o\sqrt{2}}{\sqrt{RL_m f_s}} \qquad (10)$$

According to analysis of the relation between L_m and f_s, increase in L_m results in low switching frequency, hence, low switching loss. On the other hand, boundary mode and DCM eliminate turn-on loss of the primary switch and reverse recovery of the secondary diode. Even if the increase in L_m can be beneficial for switching losses, the tradeoff due to increase in L_m is increase in size of flyback transformer and increase in windings, which causes more copper and core losses. The problem is that there is a relation between copper, core losses and frequency. Actually, this type of relations are the signs of need in optimization.

After considering the iterative relation between L_m and f_s, the primary inductance is chosen to be between 0.4-0.5mH.

4.4. Core Selection

Core selection is based on the shape of a core, size of a core and magnetic characteristics of a core. Here we will determine the type, shape, size and air gap of the core.

I. Magnetic Characteristics of Core

General approach for the selection of core type is divided into two for flyback transformers; distributed air-gap powder cores and gapped ferrite cores. Both types of cores have their own advantages. For example, ferrite cores provide more permeability and higher frequency of operation with less core losses. On the other hand, distributed air-gap powder cores may have less air gap loss compared to gapped ferrite cores and they also have higher saturation points.

Ferrite core with an air-gap is selected based on the fact that they are more commonly used in flyback transformers and also provide much more options compared to their distributed gap counterparts.

II. Shape of Core

Common preference for flyback transformers is some kind of core similar to E type core where the core is made up of two pieces. The reason for this is while toroidal cores are more efficient due to their geometry, they are very difficult to use while mass manufacturing transformers since the winding process is hard to automate. On the other hand other core shapes allow us to use bobbins to wind the windings before we put them in the core. In the end we chose the ETD type cores since they seem to be recommended for higher power applications.

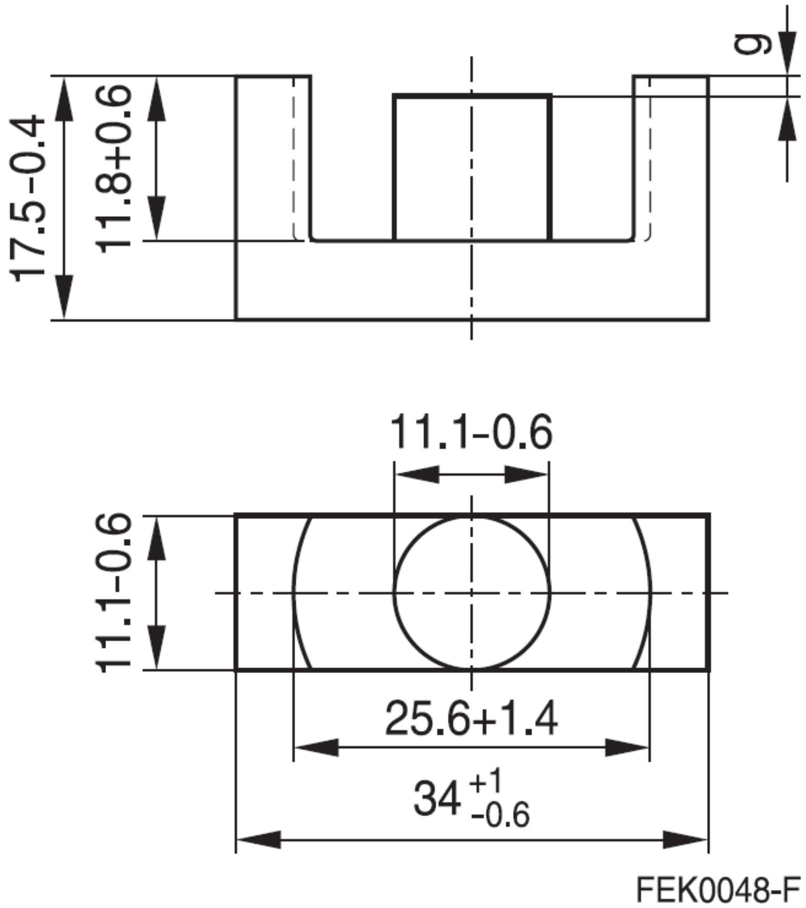


Figure 3 The core chosen for the flyback transformer

III. Size of Core

Size of core will be determined from available manufactured cores. Cores are researched on the TDK website and are readily available at Mouser.

From our calculations in MATLAB, we decide to use B66361G1000X187 which is an ETD34 core and has a gap of 1 mm on each piece which gives us an A_L of 153 nH/N^2.

Equation (11) shows the turns needed to achieve around 500μH.

$$N_p = \sqrt{\frac{L_m}{A_L}} = 57.16\, Turns \approx 56\, Turns \quad (11)$$

$$N_s = \frac{N_p}{8} = 7 \text{ Turns}$$

$$L_m = A_L N_p^2 = 479.81 \mu H$$

We use the equation below to calculate the maximum flux density B_m using the effective cross sectional area A_e (cm^2) for the core selected. B_m is recommended to be in the range of 0.2 tesla to 0.3 tesla due to the fact that below 0.2 the core would be underused, and above 0.3 there may be a possibility of saturation occurring depending on the ferrite material used.

Equation (12) shows the calculation of maximum flux density.

$$B_m = \frac{L_m I_{peak}}{N_p A_e} = 0.297T \quad (12)$$

We conclude that this core satisfies our requirements.

The AC flux density (B_{AC}) can be used with the core loss curves from manufacturers. This gives the AC component of the magnetic flux instead of the peak to peak. This is simple in the case of a discontinuous transformer design and is simply calculated from Equation (13).

$$B_{AC} = \frac{B_m}{2} = 149\, mT \quad (13)$$

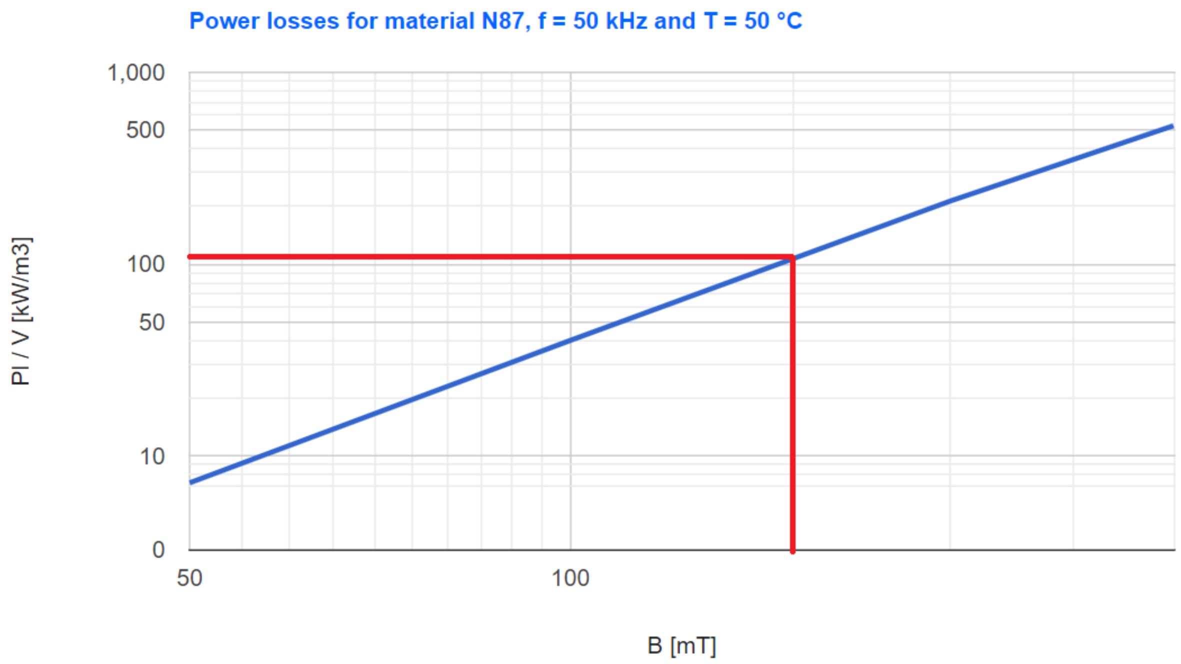


Figure 4 Power loss for the material used in the core at 50kHz

From Figure 4 we see that we have a core loss of about 110mW per cm^3 when the converter is operating at full load. The core has a volume of 7.63 cm^3 per set and this amounts to about 839mW of power loss for the whole core. Also when we use the resistance and I_{RMS} values of the primary and secondary, we find the copper loss to be about 500mW. This means that the total loss seen on the transformer at full load is 1339mW.

4.5. Winding and Fill Factor

A common rule of thumb for cable selection in a transformer is to assume the current density to be 4 A/mm^2. Using an AWG chart, we find the recommended wire types which allows us to have our desired current density. We decided to use AWG22 for the primary and AWG12 for the secondary windings. AWG28 is more than enough for the tertiary winding. The window area with the bobbin is 122 mm^2.

Table 3 Transformer cable sizes and fill factor

	AWG	**# of Turns**	**Cross Section of Chosen Cable**	**Total Cross Section Area**

Primary	22	56	0.327 mm^2	18.31 mm^2
Secondary	12	7	3.31 mm^2	23.17 mm^2
Tertiary	28	7	0.08 mm^2	0.56 mm^2
Total	-	70	-	42.04 mm^2
Fill Factor	34.46%			

We find the fill factor to be satisfactory since with the inclusion of insulation tape and such the available window area will be filled. Also since we must take in account the skin effect of the windings, the chosen windings will either have to use litz wire equivalent to the selected AWG rating or choose multiple thinner wires to have the same area as the selected wires and combine them. In the end we opted to use multiple strands of AWG28 wire due to the fact that its skin effect can be ignored for the entirety of the converter's frequency range.

5. Simulations

After finalizing the analytical calculations, the simulation model is constructed and simulated. For the simulations we are using LTSpice since we have opted to use controllers manufactured by Analog Devices. LTSpice is a SPICE-based analog electronic circuit simulation software that is produced by the semiconductor manufacturer Analog Devices. It is the most widely used distribution of the SPICE software. Here the converter design with the LTC8316 and LT8309 controllers are implemented and have been analyzed. The resulting waveforms have been captured and displayed below.

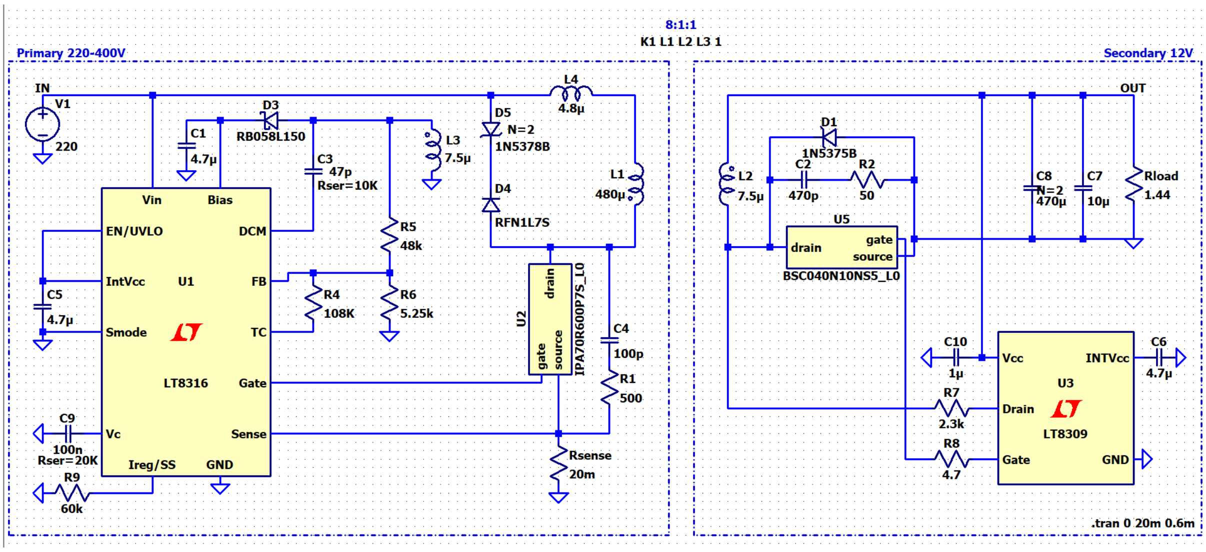


Figure 5 LTSpice Schematic of the Circuit

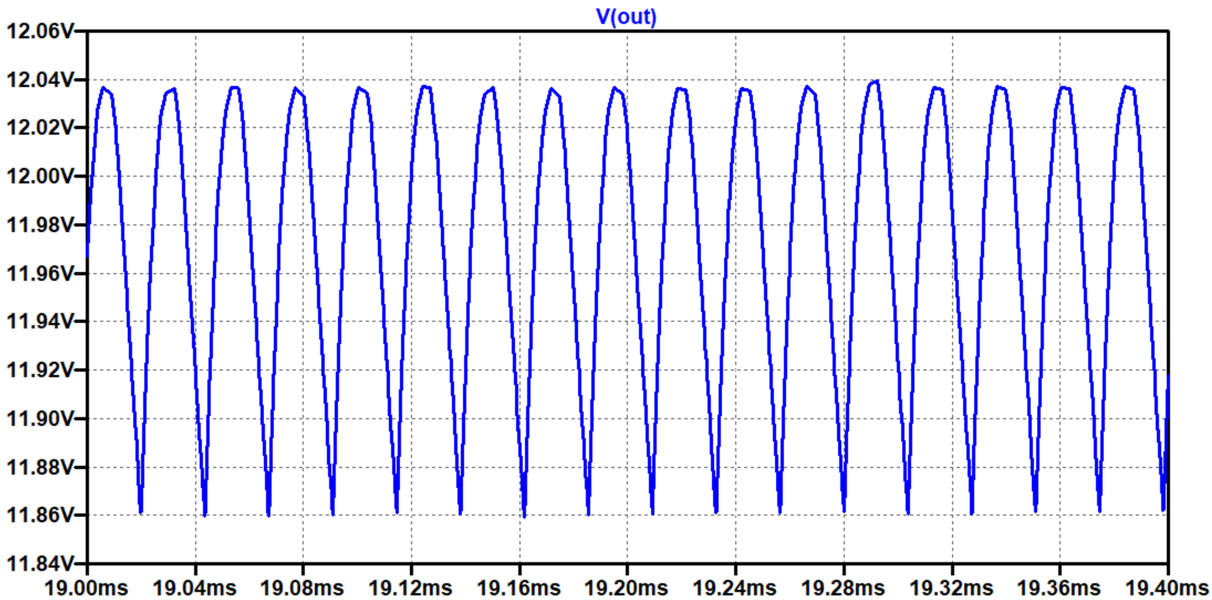


Figure 6 Output Voltage at Full Load 220V Input

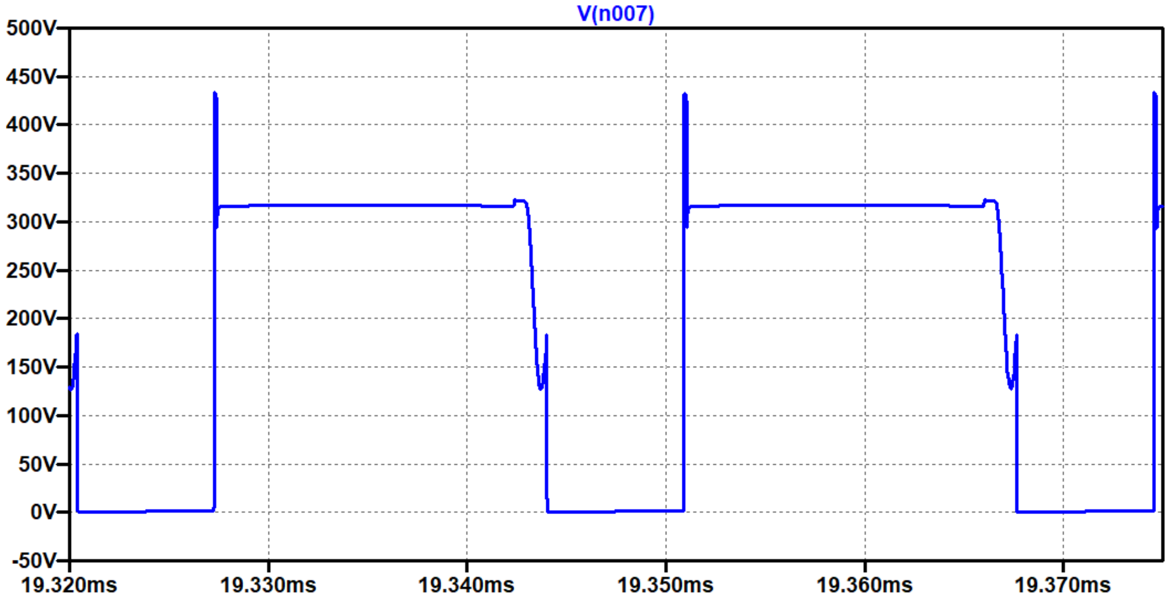


Figure 7 Primary MOSFET Voltage at Full Load 220V Input

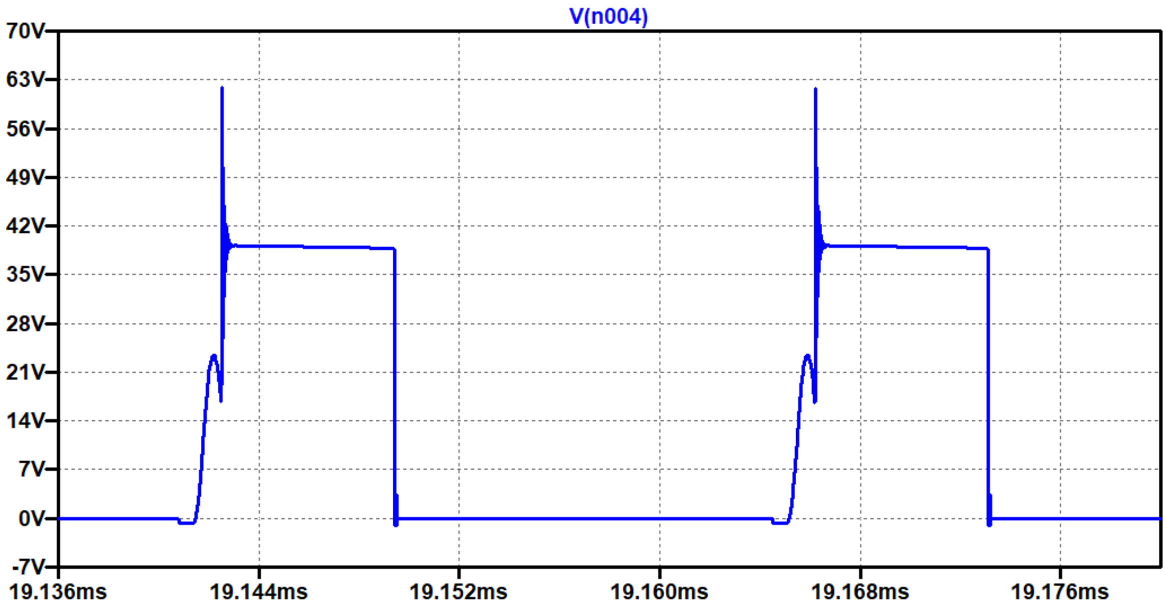


Figure 8 Secondary MOSFET Voltage at Full Load 220V Input

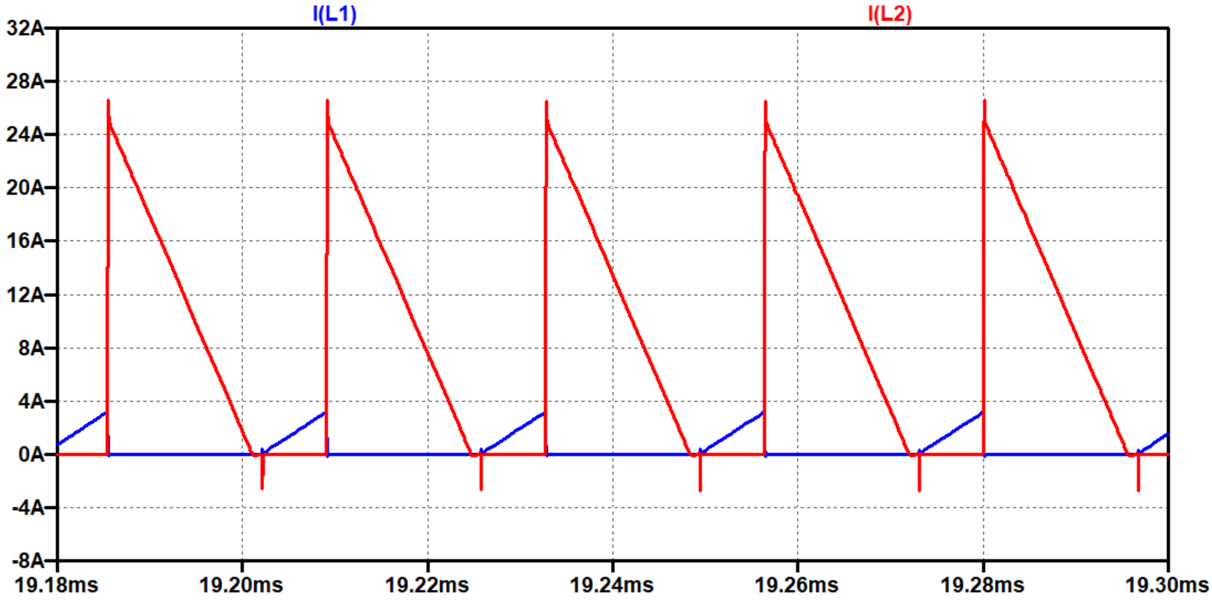


Figure 9 Transformer Current at Full Load 220V

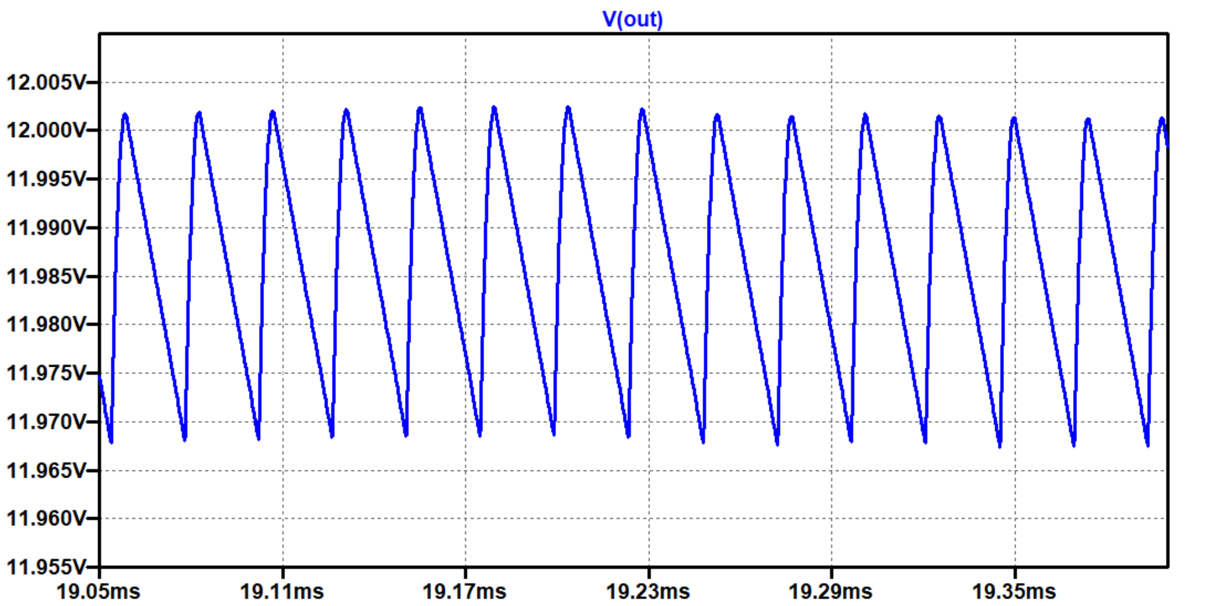


Figure 10 Output Voltage at 10% Load 220V Input

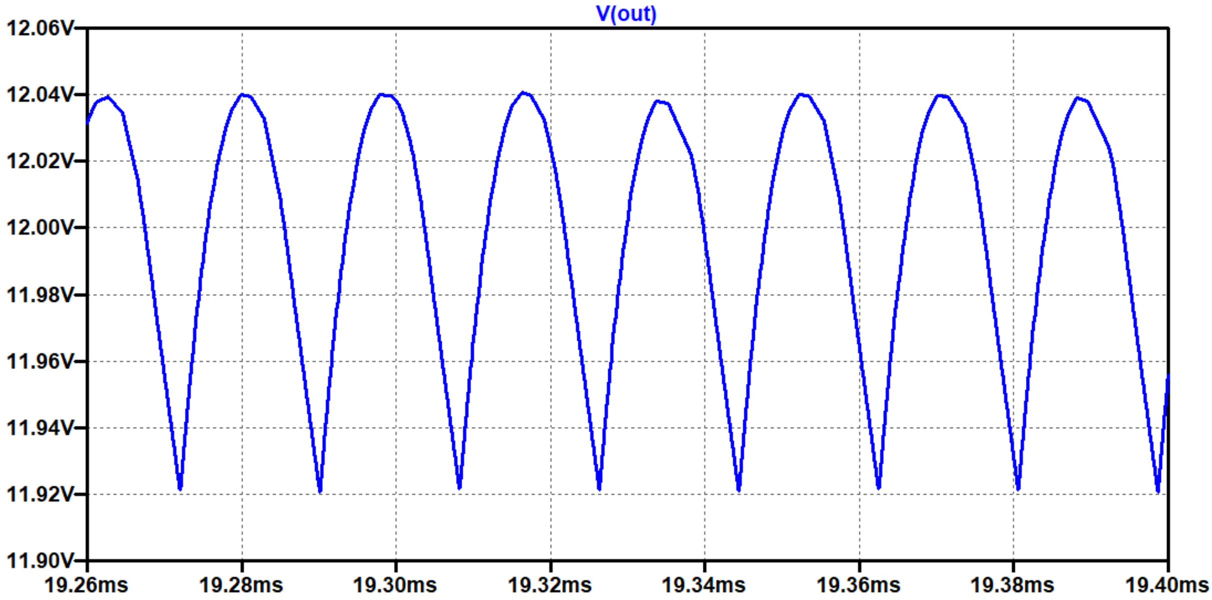


Figure 11 Output Voltage at Full Load 400V Input

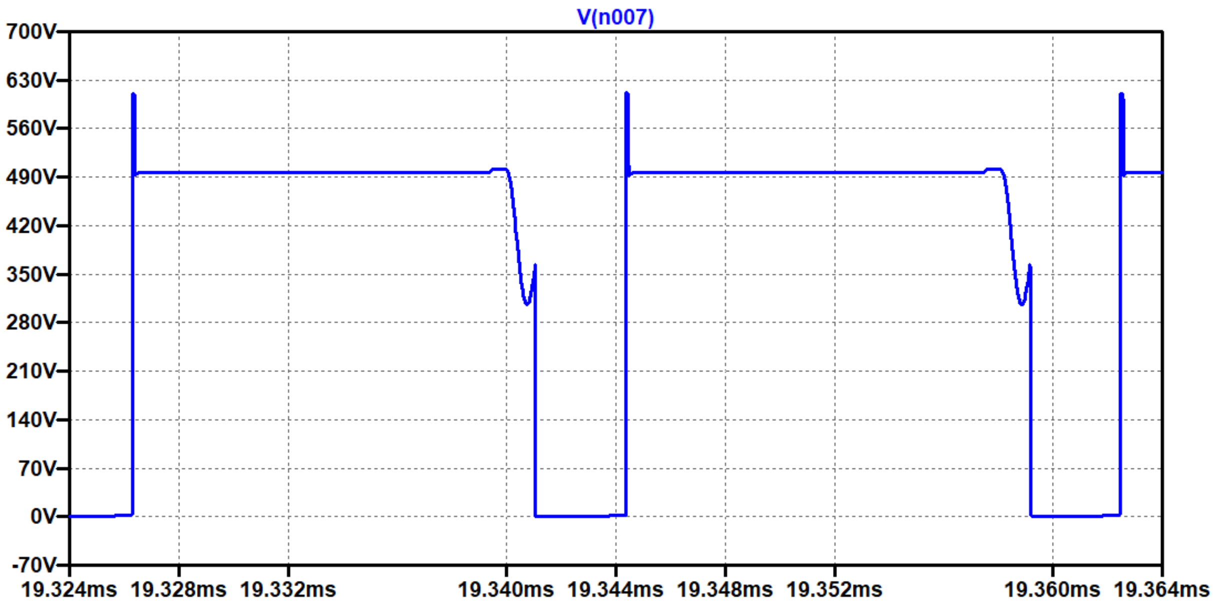


Figure 12 Primary MOSFET Voltage at Full Load 400V Input

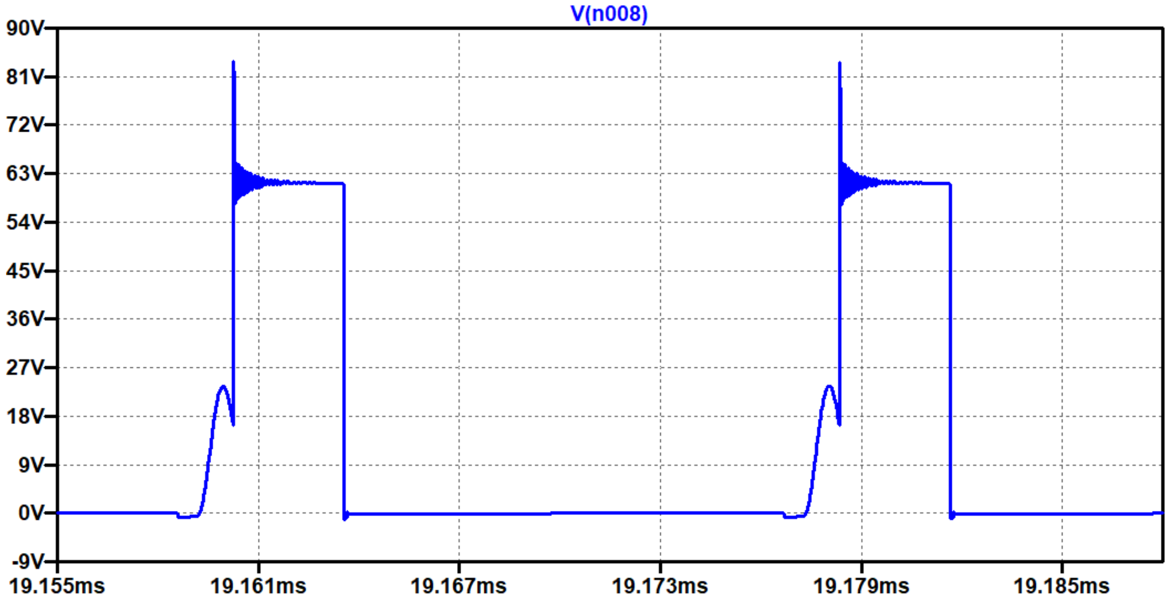


Figure 13 Secondary MOSFET Voltage at Full Load 400V Input

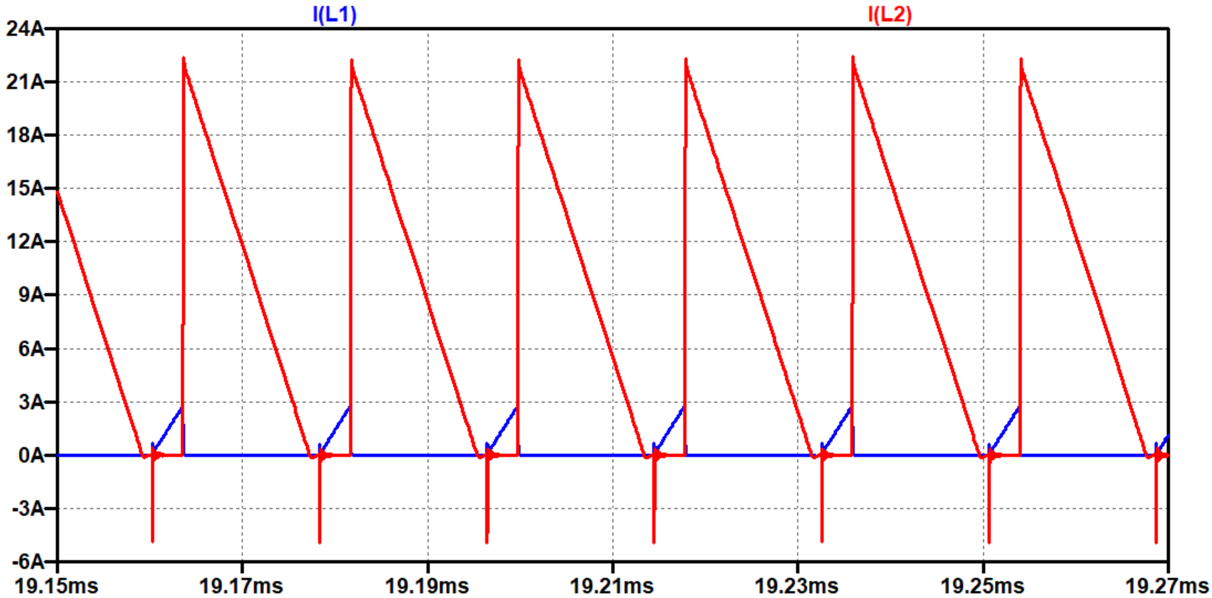


Figure 14 Transformer Current at Full Load 400V

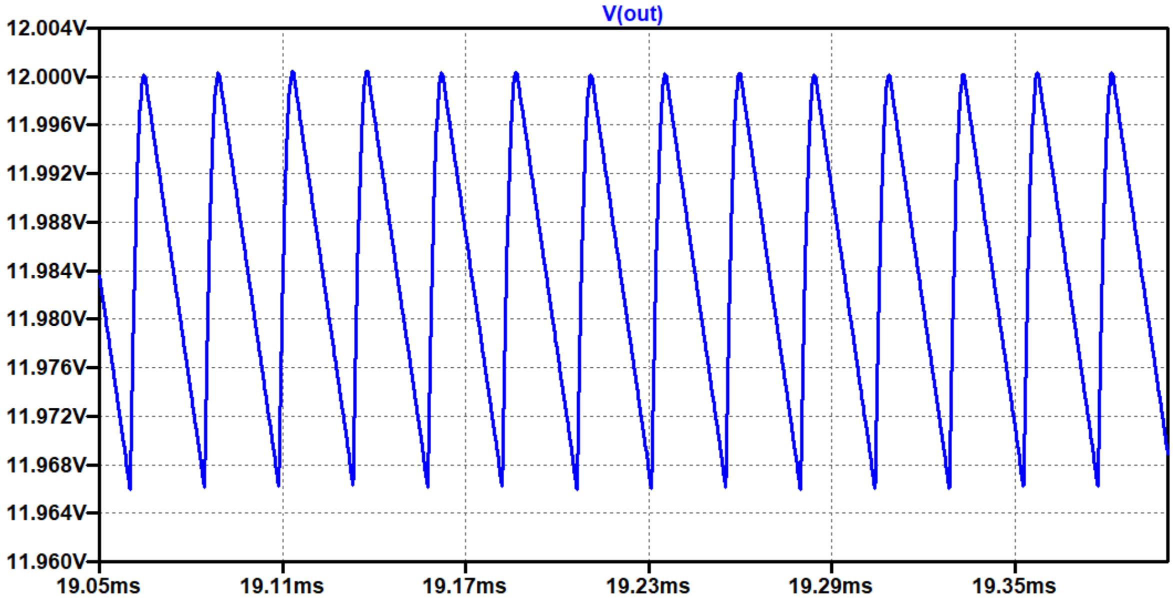


Figure 15 Output Voltage at 10% Load 400V Input

The converter is simulated in LTSpice and the resulting waveforms are given above. All the voltage and load regulation requirements are satisfied. From the MOSFET waveforms we will choose our switches accordingly.

6. Component Selection

There are three main component groups in this circuit. Power components, peripheral components of the ICs and protective components.

Peripheral components of the controller and secondary MOSFET driver are selected by following the guidelines given in datasheets. Power components are selected by considering the simulation results and some safety margin. These components are discussed in the simulation report. However, there are some new components that were added for synchronous switching update. In this report, only primary and secondary MOSFETs are dis

Primary MOSFET is selected according to the simulation results. At full load, peak voltage between drain and source terminals are approximately 630 V and peak current through the drain terminal is 3A. These values are considered while determining the ratings of the MOSFET. Controller IC can provide switching frequency up to 140kHz, which can create a high switching loss and thermal dissipation. By considering all the limitations above and adding some safety margin, IPA70R600P7S CoolMOS Power MOSFET is selected. Table below shows the ratings of this MOSFET. Detailed datasheet can be found in the Github repository.

Table 4: Primary MOSFET parameters

V_{DS}@T=25$\circ C$	700 V
$R_{DS(ON)}$	0.6 Ω
$I_{D,max}$	20 A
Q_g	10.5 nC

Secondary MOSFET is selected by adopting the same approach. Although voltage stress on it is relatively smaller than the primary MOSFET, secondary MOSFET should be able to handle approximately 10 A. Therefore, BSC040N10NS5 OptiMOS 5 Power MOSFET of the Infineon is selected. Some of its ratings are given below.

Table 5: Secondary MOSFET parameters

V_{DS}@T=25$\circ C$	100 V
$R_{DS(ON)}$	4 mΩ
$I_{D,max}$	18 A
Q_g	58 nC

In final simulations including models of selected components, results were satisfying. Finally, total cost of 1000 boards are calculated, including PCB, case transformer cost. Final Bill of Materials report is given below. Total cost of the board for 1000 pieces is 18,8 $. Selection of protective devices is discussed in the protection section.

Table 6: BOM

Comment	Designator	Order Quantity	Unit Price ($)	Total Cost/Board ($)	Cost for 1000 Boards ($)
4.7uF	C1, C5, C13	3000	0.05951	0.17853	178.53
100nF	C2	1000	0.00696	0.00696	6.96
0.15uF	C3, C4, C15	3000	0.14268	0.42804	428.04
47pF	C6	1000	0.02178	0.02178	21.78
470u	C7, C8	2000	0.23132	0.69396	693.96
10u	C9	1000	0.06279	0.06279	62.79
100p	C10	1000	0.12665	0.12665	126.65
1uF	C11	1000	0.14662	0.14662	146.62
470p	C12	1000	0.01328	0.01328	13.28
1500pF	C14	1000	0.6545	0.6545	654.5
TVS	D1	1000	0.25955	0.25955	259.55
RB058L150	D2	1000	0.11154	0.11154	111.54
3SMAJ5949B	D3, D4, D5	3000	0.1452	0.4356	435.6
RFN1L7S	D6	1000	0.1366	0.1366	136.6
Input/output	E1, E2, E3, E4	4000	0.07986	0.31944	319.44
Fuse	F1	1000	0.06384	0.06384	63.84
IPA70R600P7S	Q1	1000	0.5404	0.5404	540.4
Sec.	Q2	1000	1.75	1.75	1750
20k	R1	1000	0.00652	0.00652	6.52
60.4k	R2	1000	0.00701	0.00701	7.01
48.7k	R3	1000	0.00701	0.00701	7.01
9.76k	R4	1000	0.00701	0.00701	7.01
110k	R5	1000	0.0061	0.0061	6.1
5.36k	R6	1000	0.00701	0.00701	7.01
4.7ohm	R7	1000	0.00396	0.00396	3.96
2.32k	R8	1000	0.00305	0.00305	3.05
500ohm	R9	1000	0.12558	0.12558	125.58
50ohm	R10	1000	0.02422	0.02422	24.22
20m	R11	1000	0.04914	0.04914	49.14
ETD34	T1	1000	0.96782	0.96782	967.82
LT8316	U1	1000	3.42	3.42	3415.5
LT8309	U2	1000	3.43	3.43	3431.76
ETD34	T1	2000	0.72541	1.45081	1450.81
AWG28	T1	5	142	0.71	710
PCB		1000	0.976	0.976	976
Case		1000	1.69	1.69	1690
Total				18.84132	18838.58

7. Thermal Calculations

From LTSpice we take the worst case for all the components that dissipate more than 250mW of power.

- Primary MOSFET: 615 mW
- Secondary MOSFET: 620 mW
- Zener Clamp (per Zener): 505 mW
- Primary RC Snubber: 1325 mW
- Flyback Transformer: 1339 mW

If we assume ambient temperature to be 25°, then we can calculate the highest possible temperatures seen on the components using their thermal resistances from the data sheet.

Primary MOSFET:

$$T_{max} = T_{Ambient} + R_{thJA}P_{loss} = 25 + 80 \times 0.615 = 74.2\,°C$$

Secondary MOSFET:

The thermal resistance is increased to 75 W/K from 50 W/K, since the defined copper area in the datasheet is not equal to the copper area of the secondary MOSFET.

$$T_{max} = T_{Ambient} + R_{thJA}P_{loss} = 25 + 75 \times 0.620 = 71.5\,°C$$

Zener Clamp:

$$T_{max} = T_{Ambient} + R_{thJA}P_{loss} = 25 + 135 \times 0.505 = 93.17\,°C$$

Note that this is for the worst case and disregards the copper in the PCB.

Primary Snubber Resistor:

Although we could not find the thermal resistance of the resistor, in the datasheet we see that there is a chart that shows the temperature of the component with respect to the dissipated power. At full load, this resistor reaches about 70 C°. This is assuming

Flyback transformer:

From the equation below we can find the temperature of the transformer.

$$T_{max} = T_{Ambient} + \left(\frac{P_{loss}}{Surface\ Area}\right)^{0.833} = 25 + \left(\frac{1339}{30}\right)^{0.833} = 47.67\,C^{\circ}$$

From our calculations and observations we see that there is no need for a separate heatsink.

8. PCB Design

PCB is drawn according to prioritized loops and components. Loops, traces and components having impact on EMI are drawn at first. Figure xx1 shows the definition of the loops for flyback SMPS.

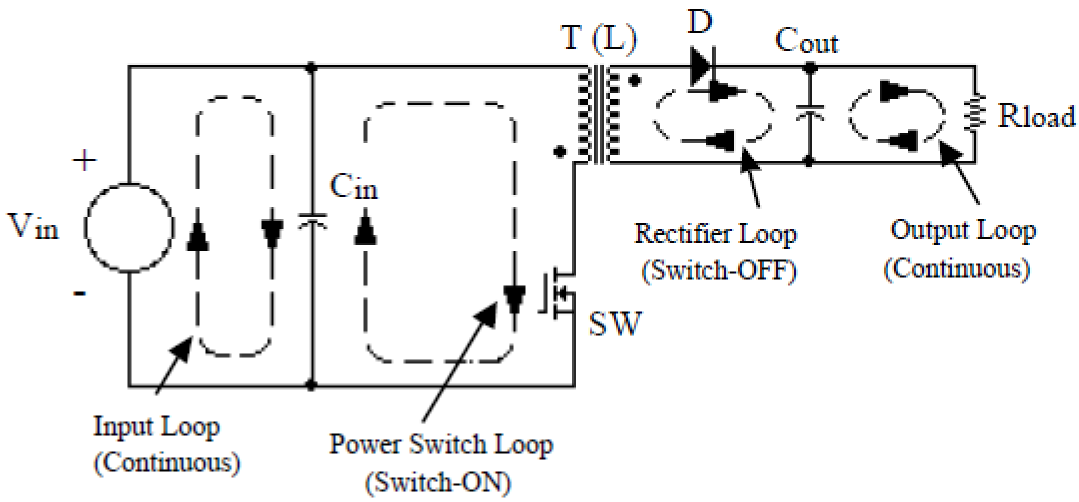


Figure 16: Current Loops in the Flyback Converter

As shown in Figure 16, the input loop and the power switch loop belong to the primary side of PCB. These loops are intentionally drawn short and wide for the sake of minimization of parasitic inductance and resistance. The width of traces for power current is defined as 2mm, which is higher than for the case temperature rise of traces equals 10°C. The importance of traces for power switch loop is significant since power switch loop current is pulsating DC. Hence, the di/dt of the power switch loop could be sensitive to parasitic inductances.

The 3D view, top and bottom layers of the designed PCB are given below. Detailed schematic and other drawings can be found in the appendix section.

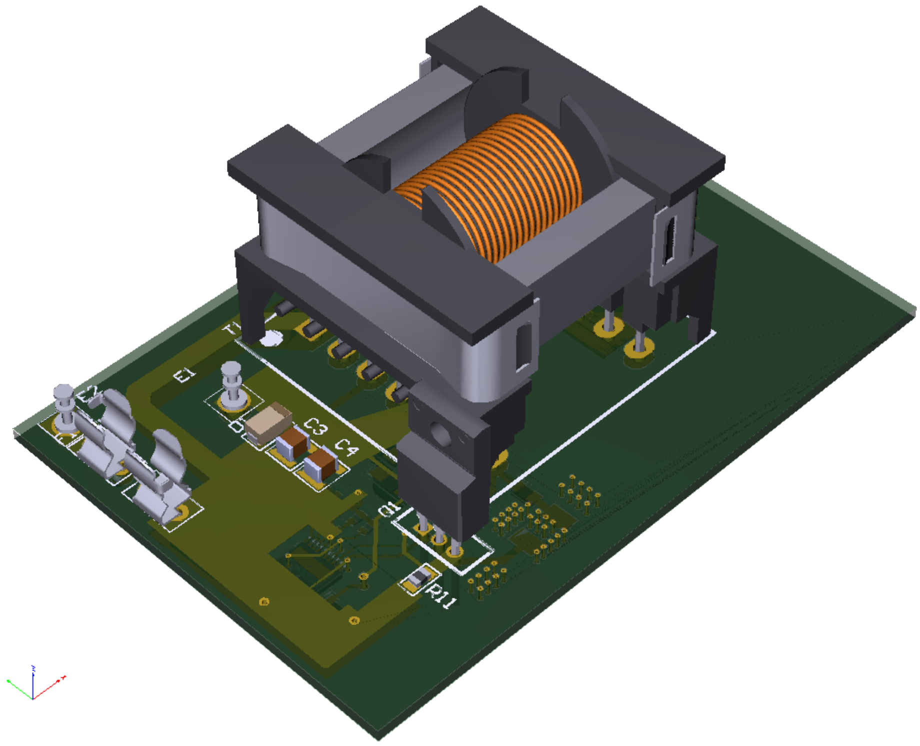


Figure 17: 3D View

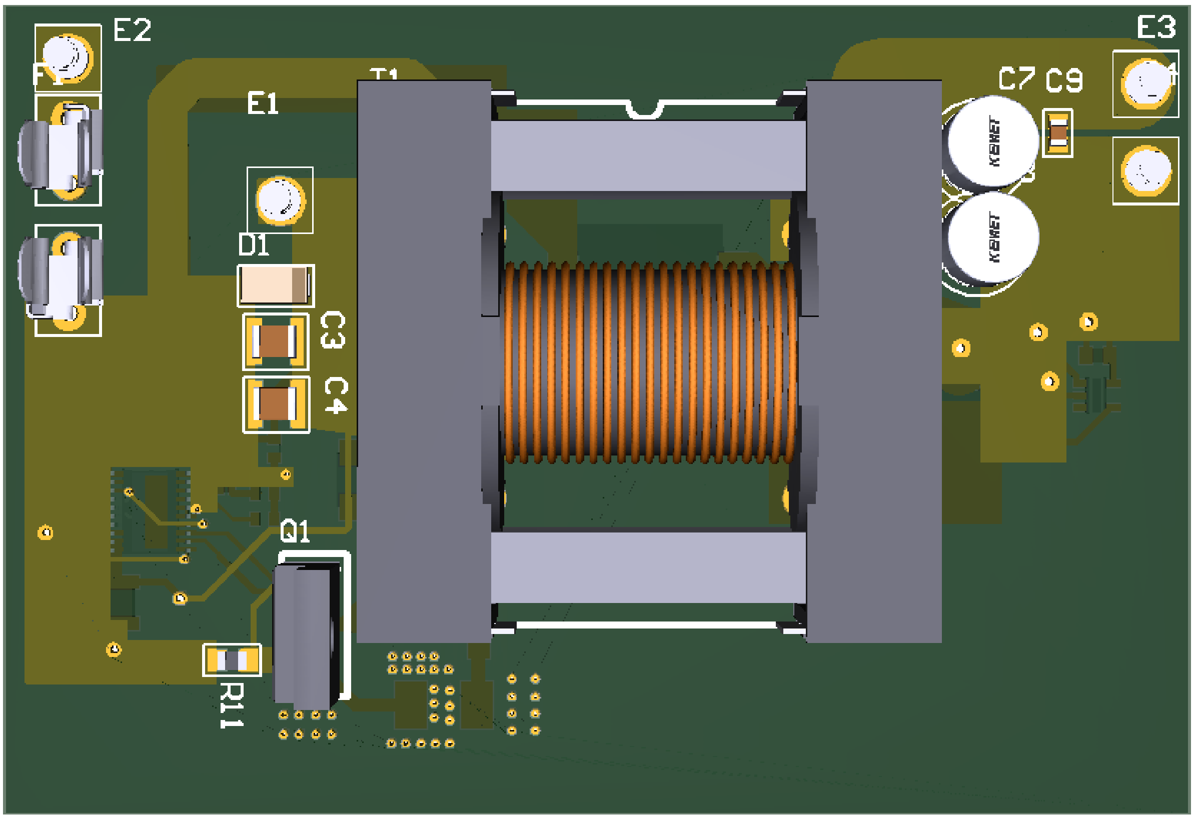


Figure 18: Top View

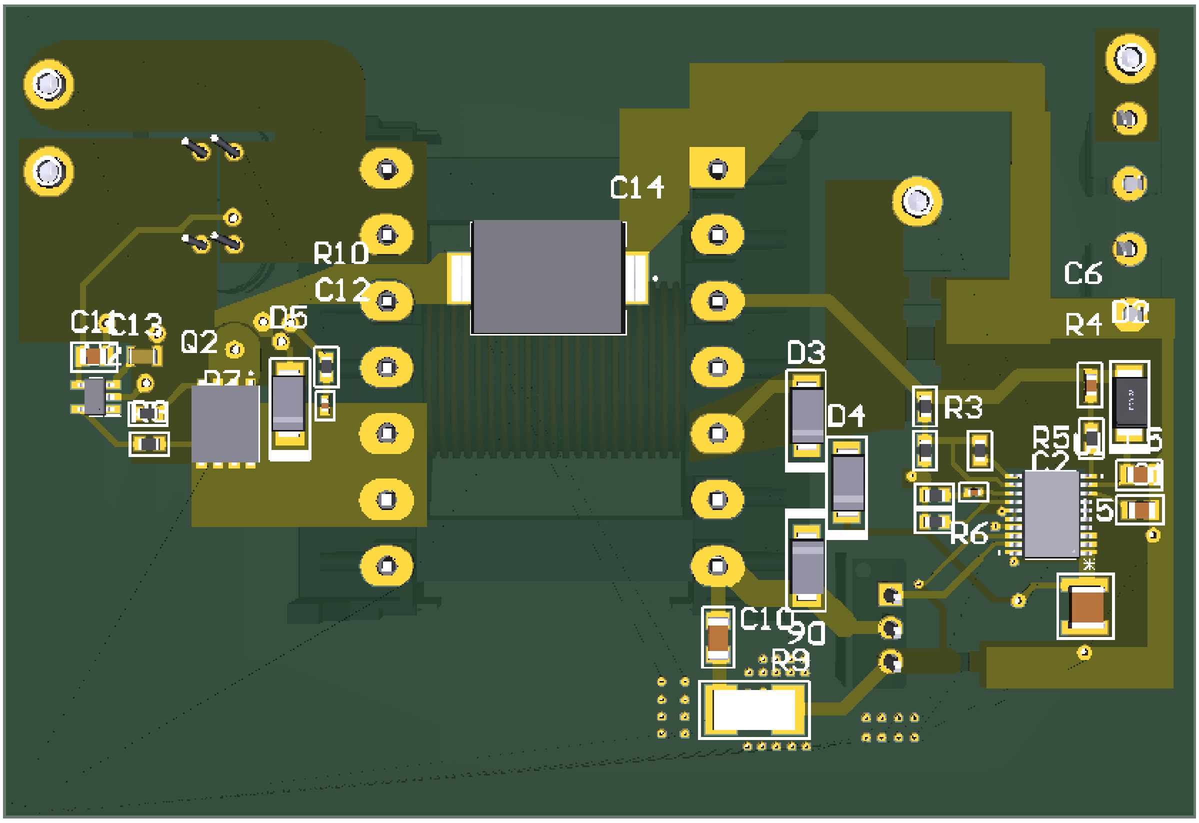


Figure 19: Bottom View

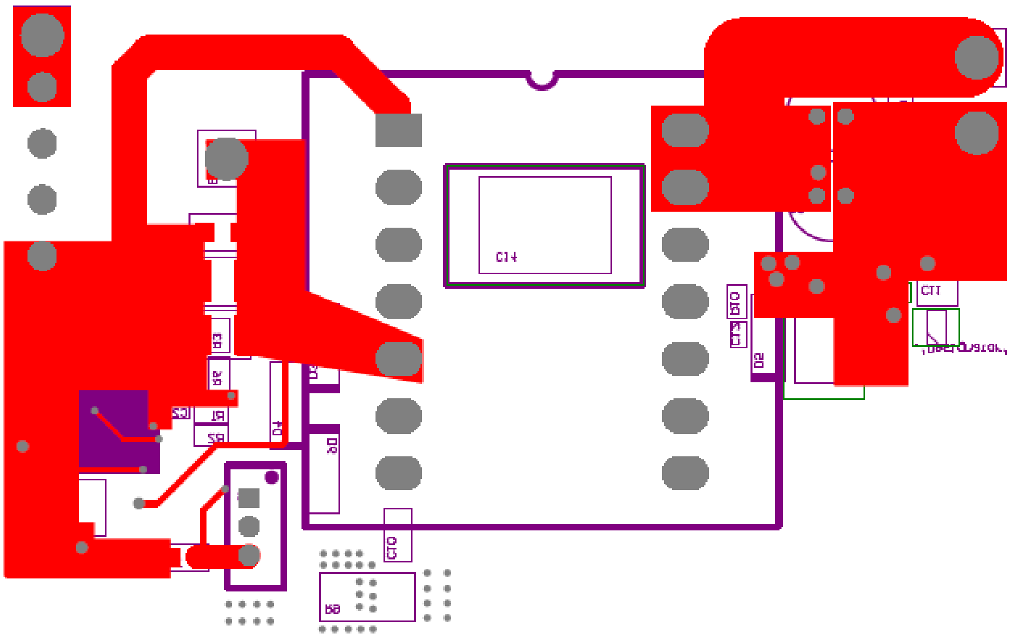


Figure 20: Top Layer

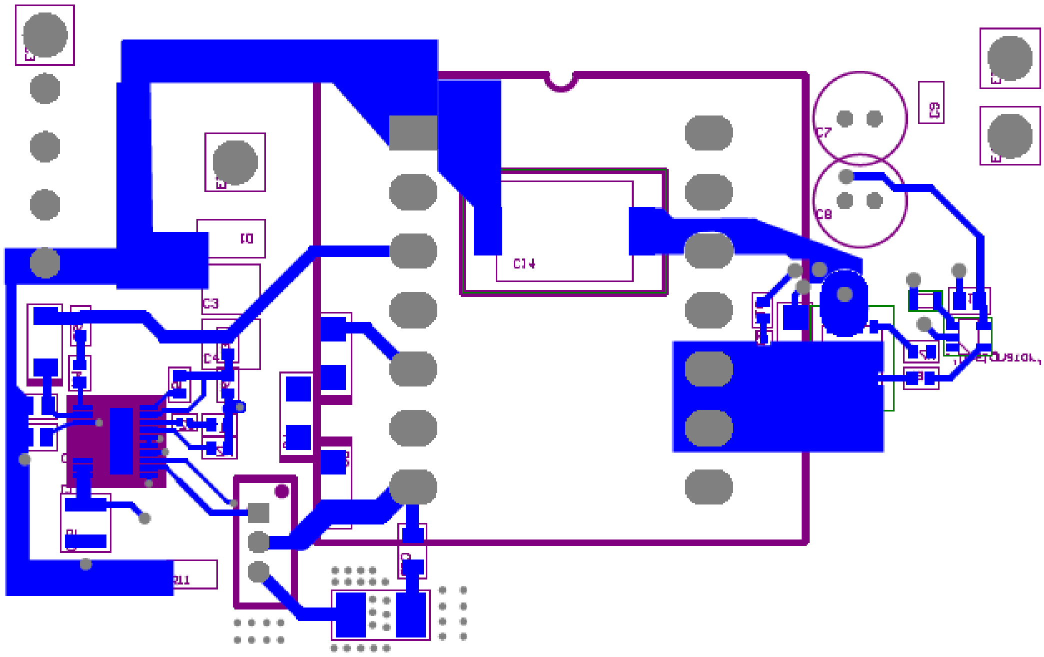


Figure 21: Bottom Layer

Primary side has importance due to the snubber circuits and gate loop having significantly high di/dt and dv/dt signal characteristics. Thus, these loops are also drawn short and wide, which can be seen in Figure 21.

Primary side polygons for power and ground are also placed without violation of clearance since the maximum input voltage is 400V. Polygons decrease trace inductance and resistance, minimize ground loops and decrease EMI effects. While diminish in inductance, resistance and length of ground loops is obvious, EMI effect can be explained with shielding. When the bottom and top layer is covered by copper, high frequency magnetic flux is absorbed on copper area as an eddy current loss.

Secondary side is designed with the same considerations. Traces having high di/dt and current rates are kept wide and short. Actually, a polygon plane is used to the region that carries 10 A current to decrease the resistance of the copper area. Controller signal traces are tried to be placed such that analog signals are affected from high current traces less. Top-bottom placement is determined by considering compactness and thermal management. Trace and plane clearances are considered during the placement. Clearance between primary and secondary sides is 1,5 cm at least.

9. Thermal Simulations

Thermal simulation is applied on PCB with SolidWorks Thermal Simulation tool. According to thermal calculations, there is no need to heatsink and natural convection of air is sufficient to keep components on PCB in operating conditions.

Thermal simulation on PCB requires some information about material characteristics of components, power loss, thermal resistance of contacts and convection coefficient of air. After determination of requirements, components and PCB are meshed. The size of mesh is defined according to material size. For instance, the triangular mesh size for the transformer is 2mm and the mesh for snubber resistance is 0.6mm. On the other hand, even if the smaller mesh size results in more accurate results, the simulation time increases inversely with size of the mesh. Figure 22 shows the meshed PCB and components.

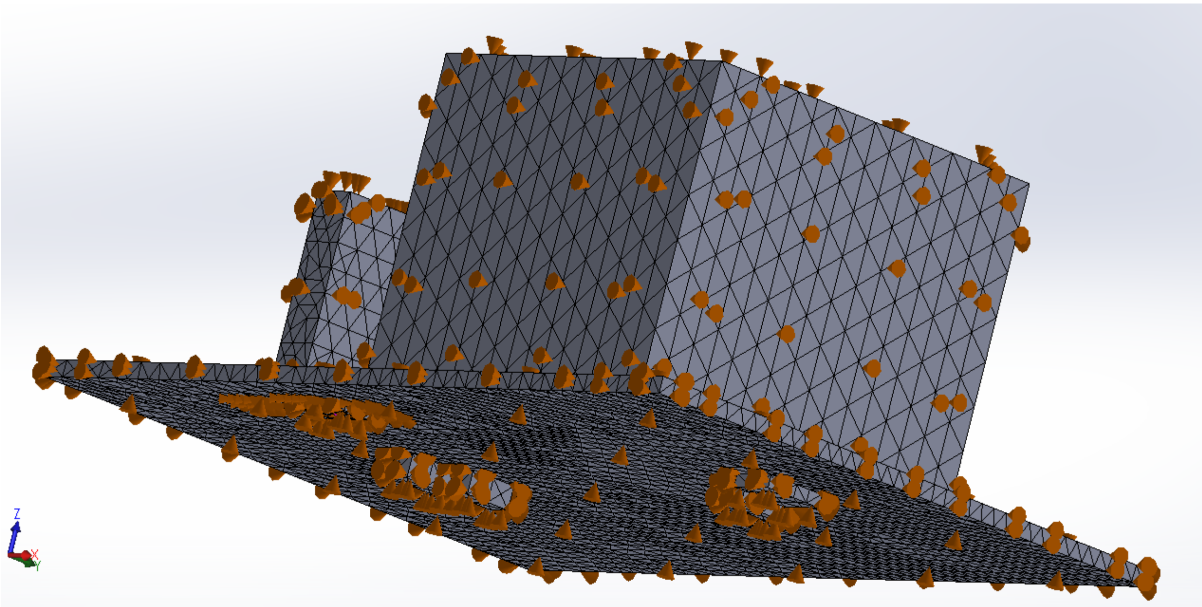


Figure 22

Thermal characteristics of semiconductors and the board, FR4, are determined according to the Engineering Database in Flow Simulation toolbox in SolidWorks. Thermal contact resistances between components and board is determined with consideration of soldering and junction-to-lead thermal resistances. For instance, solder has 50 W/m.K thermal conductivity. When the soldering area and length of soldering are considered, approximation for thermal contact resistance becomes 1.5 K/W based on the application not by ONSemi, AN9596. [2] Hence, as we can see that thermal contact resistances are relatively small.

Simplification of PCB in terms of thermal analysis is important and sometimes, it is a requirement. The simplification on PCB is done with filtering out many components which have no significant power loss. Moreover, the simplification of the 3D model of components is also very handy with CircuitWorks in order to mesh PCB properly. Simplification has a drawback that surface and contact area are changed slightly. However, according to the webinar by NineDot Connects, most of the simplifications in CircuitWorks have no significant effect on thermal analysis. [3] Figure 22, simplified model of PCB and components by CircuitWorks

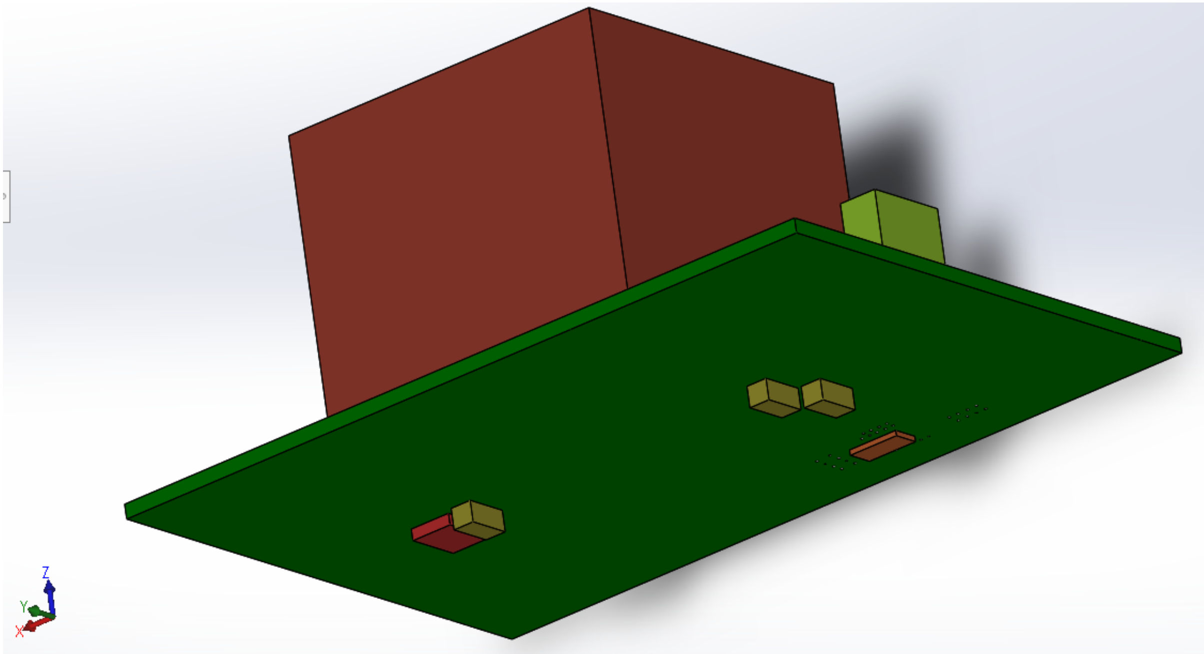


Figure 23

Power loss of components are calculated at 220V input case, since it is the case in which efficiency of the circuit is the worst. The power loss of each component shown in the Thermal Calculations part is assigned and the convection coefficient is taken as 20. The maximum temperature is observed on snubber resistance. The value can be taken as pretty much, but there is still approximately a 30 °C margin for the maximum point of optimum operating range. Thermal analysis of the PCB can be seen in top and bottom views in Figure 24 and 25.

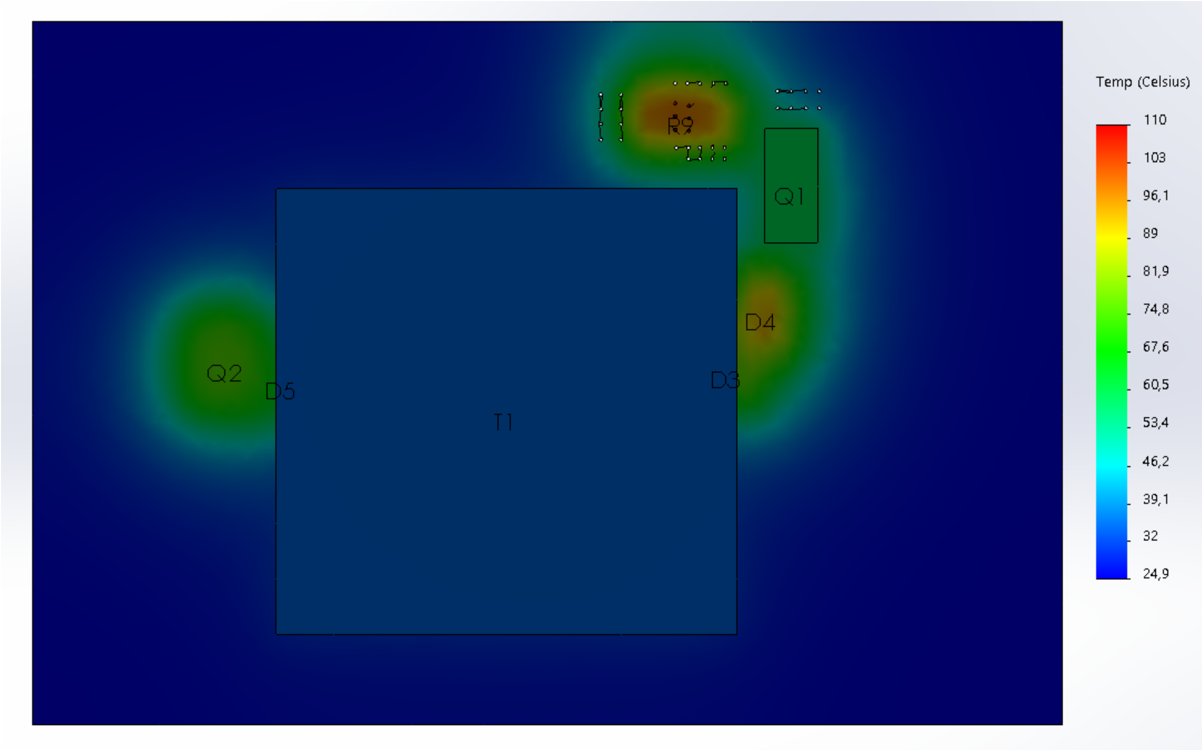


Figure 24: Thermal Simulation top view

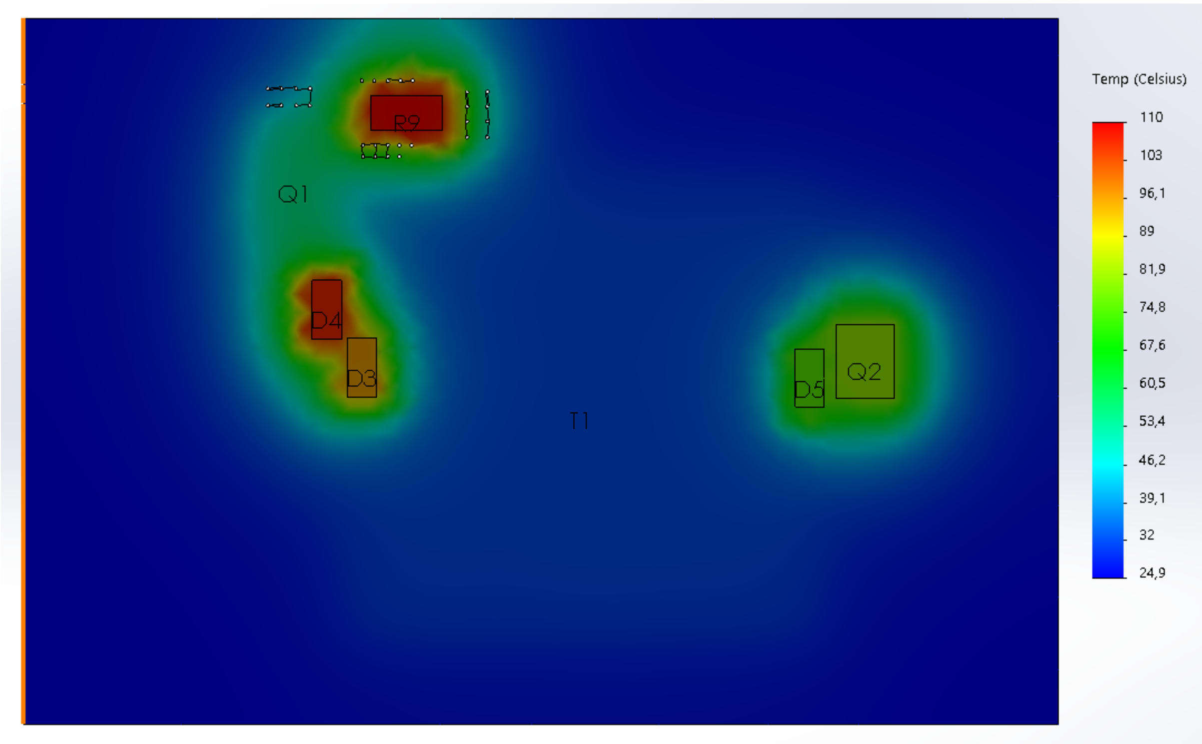


Figure 25: Thermal Simulation bottom view

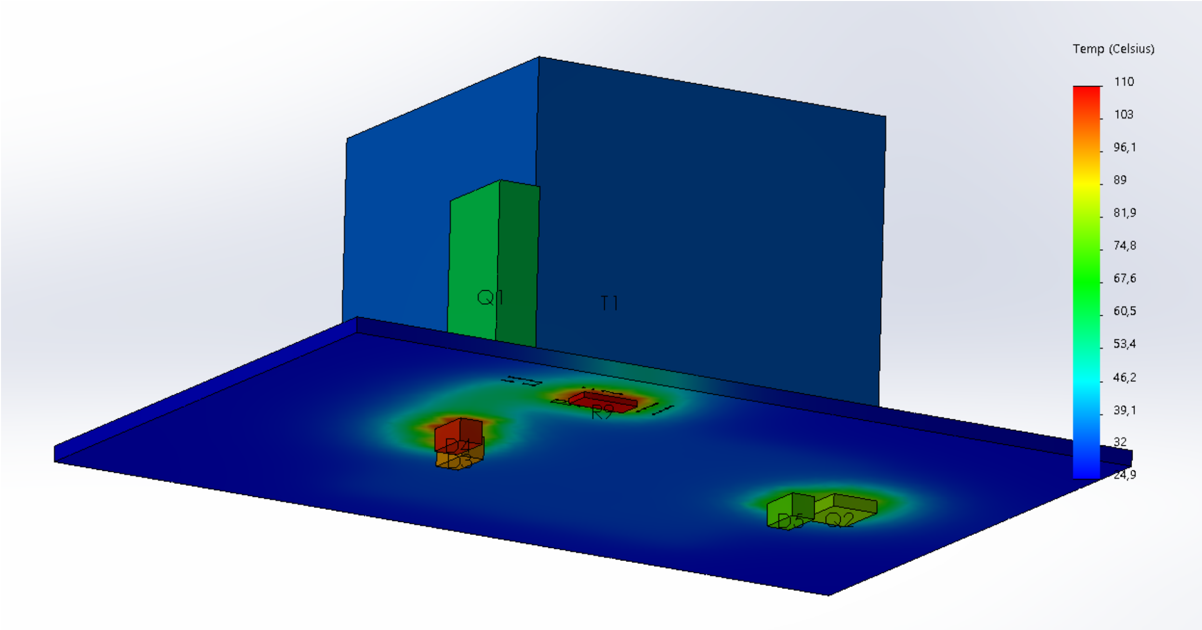


Figure 26: Thermal Simulation 3D view

The temperatures of components are measured in Figure 27 and Figure 28. The closeness of snubber resistance, R9, clamp zeners, D3 and D4, and primary mosfet, Q1, results in a high temperature in R9. However, the most important parameter making snubber hot is the small volume of the resistor package. Convection of air on snubber resistance is relatively low due to the small volume. Therefore, some thermal vias are added and the board is strengthened in the snubber resistance side.

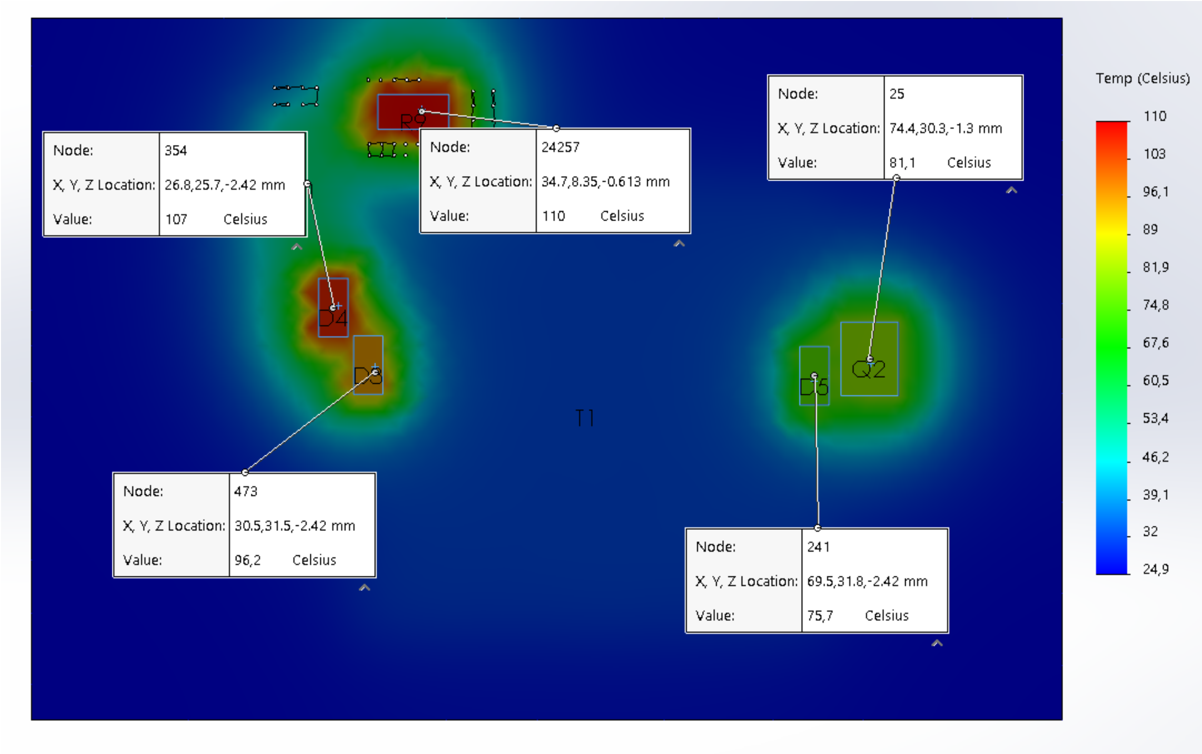


Figure 27: Thermal Simulation bottom view, component temperatures

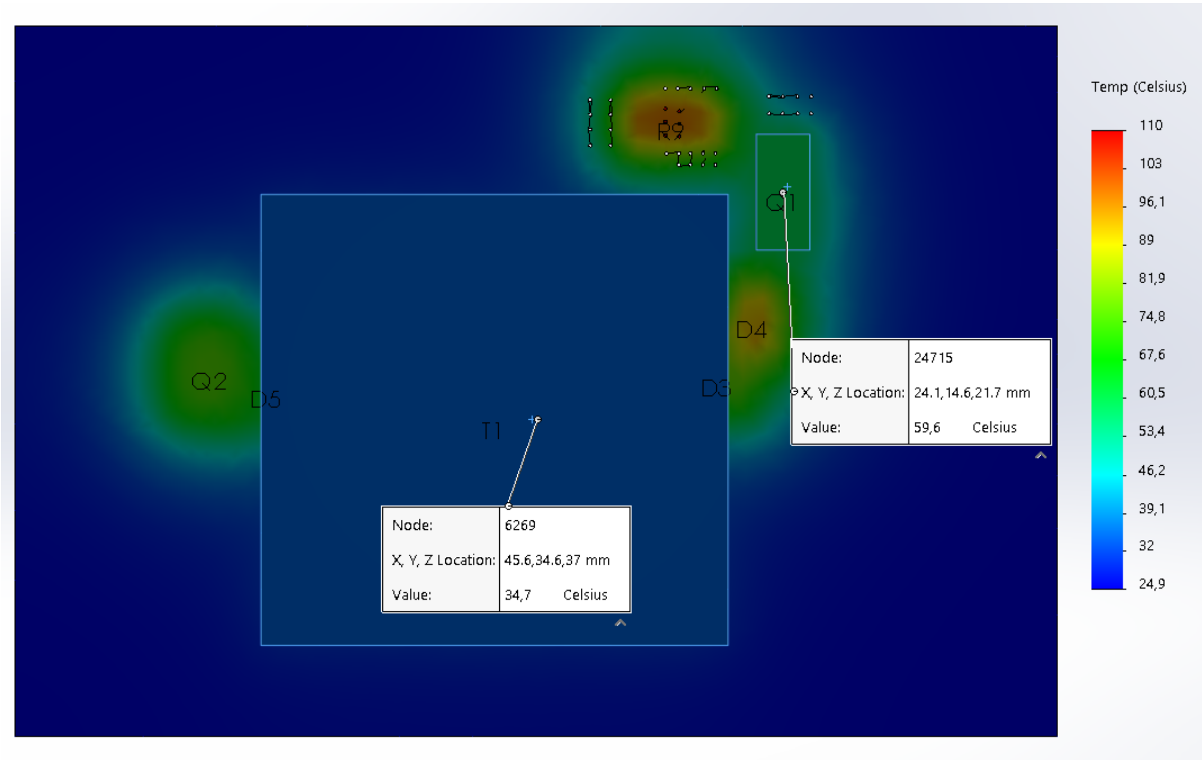


Figure 28: Thermal Simulation top view, component temperatures

10. Protection

Protection components, fuse and TVS, are added against high current and high voltage, respectively. Fuse interrupts the input when the maximum current rating of the fuse is exceeded. In order to provide a sustainable current protection, a fuse holder is placed on the PCB. Fuse is placed in this holder and can be changed in case of melting. Fuseholder is placed just after the neutral connector of the input. 3D view of the fuse holder is given below.

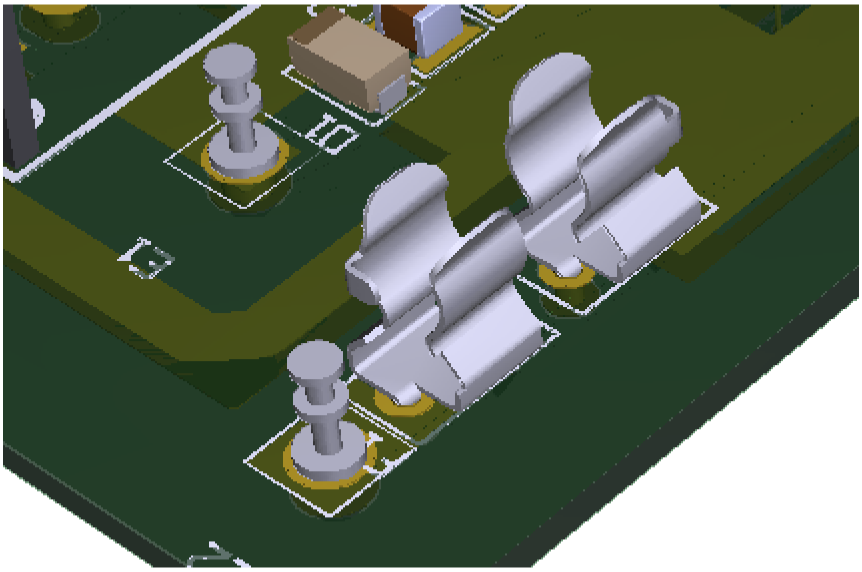


Figure 29: Input Connector and Fuseholder

TVS is used for suppress the input voltage transients. Controller IC receives its supply voltage directly from the input voltage. Therefore, it is important to limit the input voltage in case of fast transients like spikes. Maximum continuous supply voltage for the IC is 600V. Therefore, an TVS is selected such that input voltage is clamped to the 600-650V region during transients from 10 to 100 µs.

Finally, it can be concluded that our circuit is protected from input overcurrent and voltages with fuse and TVS usage.

11. Conclusion

Flyback topology is selected to implement the isolated DC/DC converter from variable input voltage range (220V-400V) to 12V output with 100W output power. The abundance of readily available ICs and the simplicity of the flyback converter were the main reasons for our decision.

LT8316 is selected among some other alternative controllers and is used in conjunction with the LT8309 for a more efficient design. The use of the LT8309 has provided the design a 6% increase in efficiency due to the absence of the forward voltage drop we would otherwise see in a diode. Since Analog Devices has LTSpice models of controllers, LT8316 and LT8309 are easily implemented on the simulation tool.

Determination of the topology and the controller have given us some parameters to follow while designing the transformer and clamp circuit and also selecting the semiconductors. Clamp circuits and semiconductor ratings depend on each other and they have been selected with that consideration. On the other hand, even if the transformer design has rule of thumbs and requirements, there is a huge gap for optimization in terms of core loss, copper loss and size. Thus, selection of turns ratio, primary inductance and winding have been explained in detail.

We have done the thermal calculations and decided that there is no need for any heatsinks. The PCB has also been designed and any protection circuit we have come up with has been added to the design. In the end we used Solidworks to run a thermal simulation on the finished product.

The end results are a product that meets the demands given to us and has a peak simulated efficiency of around 95%. We acknowledge the fact that this number will probably be less in real life.

12. References

[1] Analog Devices LT8316 Datasheet

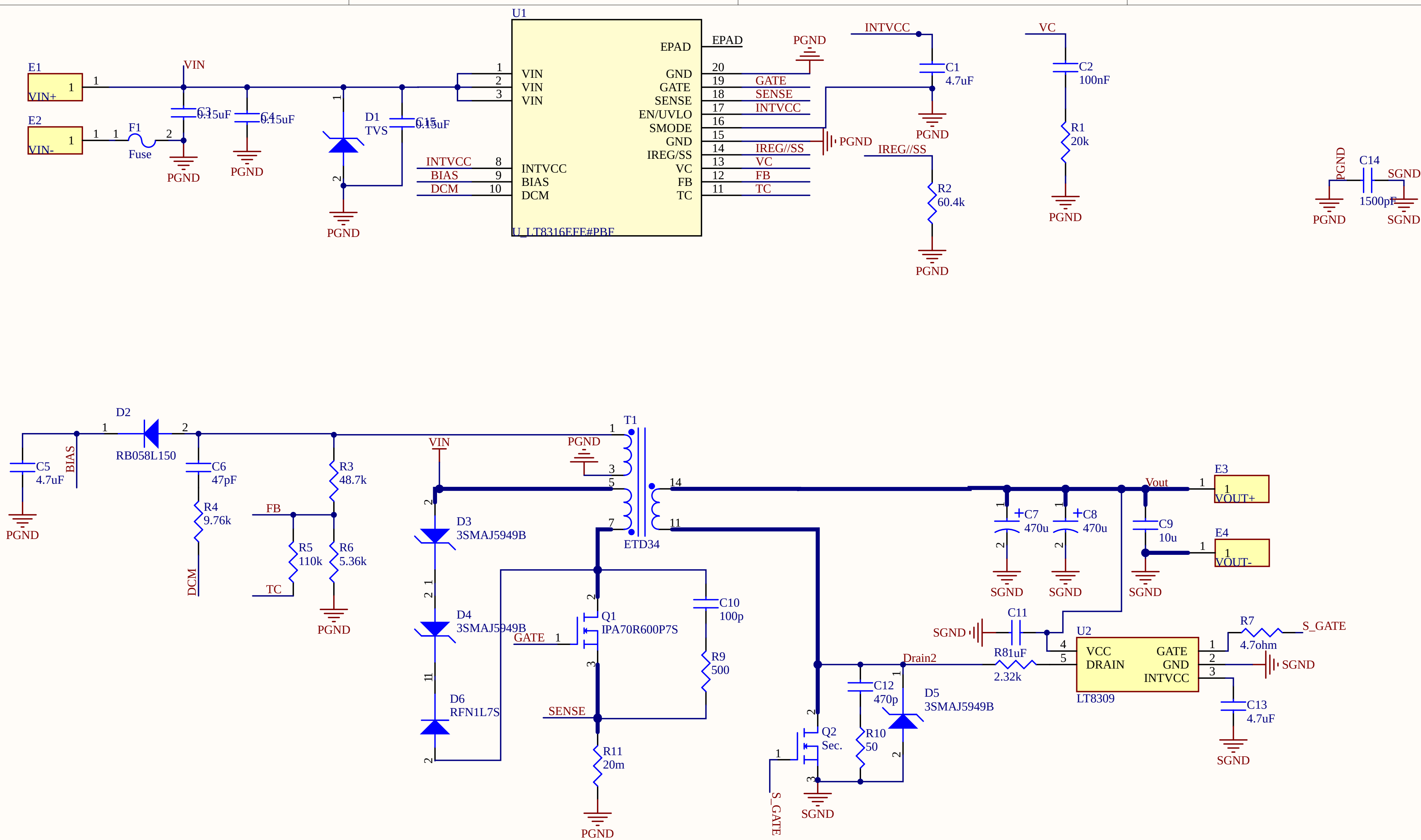
https://www.analog.com/media/en/technical-documentation/data-sheets/lt8316.pdf

[2] On Semiconductor, A Quick PCB Thermal Calculation for Power Electronic Devices with Exposed Pad Packages

https://www.onsemi.com/pub/Collateral/AND9596-D.PDF

[3] Beyond a Heatsink - Thermal Considerations in PCB Design. Webinar

https://youtu.be/wEdn9kPAX1I



Power traces are shown with thick wires. They will be short and wide

There are two ground planes. They are connected with a capacitor

Title		
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