ANALYSIS AND DESIGN OF A FORWARD POWER CONVERTER

by

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ABSTRACT

With advances in microelectronic fabrication and semiconductor devices, the size of electronic equipment is ever decreasing. Virtually, every piece of electronic equipment is powered from a DC power source, be it a battery or a DC power supply. The power supply constitutes to a main portion of any electronic device. Therefore, the key requirements for a power supply are lighter weight, reduced size, and reduced heat dissipation. There have been two technologies in power supplies: Linear and Switched mode. Though linear supplies have many desirable characteristics such as simplicity, low output ripple, excellent line and load regulation, fast response time to line or load changes, and low EMI, they suffer from poor efficiency and occupy large volumes. Switching power supplies are becoming popular because they offer better solutions to these problems.

Switching voltage regulation is the technique by which an unregulated source power is efficiently converted to regulated load power through the use of controlled power switching devices and energy transfer elements. DC to DC conversion is the heart of any switched mode power supply. There are many choices for DC to DC conversion but forward converter has dominated the market for the past 50 years for commercial power supplies in excess of 50W. At the beginning, this thesis describes the basic functional blocks of a switched mode power supply. Then the design of various elements that are generally used in any switched mode power supply is extensively covered. Later the theory and operating principles and limitations of a forward converter are discussed. Finally, the simulation results for a forward converter and test results on an engineering reference switched mode power supply are presented.

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CHAPTER – I

INTRODUCTION

With advances in electronics, the need for power supplies for use in integrated circuits and digital circuits has increased manifold. At present, realization of new portable electronic devices is one of the fastest growing segments in the electronics industry. The key portable system requirements are lighter weight, reduced size and reduced heat dissipation. The continuing advancements in semiconductor device technology have triggered the development of many smaller and lighter electronic products. Almost all the electronic devices e.g. computers and their peripherals, calculators, TV and hi-fi equipment, and instruments, are powered from a DC power source, be it a battery or a DC power supply. Electronic devices require not only a DC power source but also a source giving a well filtered and regulated DC voltage. Therefore, to meet these demands, the power supply itself has become more and more sophisticated.

A *power supply* is a buffer circuit that provides power with the characteristics required by the load from a primary power source with characteristics incompatible with the load. It makes the load compatible with its power source[1].

The regulated power supply technology can be divided into two distinct technologies; firstly, the *linear or series regulator* and, secondly, the *switched-mode conversion* technique. The linear voltage regulator behaves as a variable resistance between the input and the output as it provides the precise output voltage. One of the limitations to the efficiency of this circuit is due to the fact that the linear device must drop the difference in voltage between the input and output while simultaneously conducting current. Due to this dissipation, linear regulators have to be adequately cooled, by mounting them on heat sinks and the heat is transferred from the heat sinks to the surrounding air either by natural convection or by forced-air cooling. Heat sinks and provision for cooling makes the regulator bulky and large. While these supplies have many desirable characteristics, such as simplicity, low output ripple, excellent line and load regulation, fast response time to load or line changes and low EMI, they suffer from low efficiency and occupy large volumes. In applications where size and efficiency are critical, linear voltage regulators cannot be used.

A *switching-mode power supply* is a power supply that uses low loss components such as capacitors, inductors, and transformers and the use of switches that are in one of two states, ON or OFF[1]. The advantage is that the switch dissipates very little power in either of these two states and power conversion can be accomplished with minimal power loss, which equates to high efficiency.

Switch mode regulators overcome the drawbacks of linear regulators. Switched mode power supplies (S.M.P.S.) tend to have an efficiency of 80% or more. They can be packaged in a fraction of the size of linear regulators. Furthermore, by employing high switching frequencies, the sizes of the power transformer and associated filtering components in the S.M.P.S. are dramatically reduced in comparison to the linear. For example, an S.M.P.S. operating at 20 kHz produces a 4 times reduction in component size, and this increases to about 8 times at 100 kHz and above. Unlike linear regulators, switched power supplies can step the input voltage either up or down.

Switching voltage regulators are used where it is required to supply a stable output voltage from a varying input voltage with a minimum power loss even under varying load conditions. Though SMPS have everything a power supply needs to have, they are very complicated to design. Owing to circuit complexity, Electronic hobbyists or even electronics technicians cannot easily employ them in a casual fashion. A traditional 12V DC power supply providing 1A of current is well within the construction capabilities of every electronic hobbyist with some experience. A switched mode power supply on the other hand could prove a very daunting task.

Switching power supplies can be classified into three types according to their input and output voltages. 1) The AC/DC power supplies 2) The DC/DC Converter 3) The DC/AC inverter[2].

At the heart of the converter is the high frequency inverter section, where the input supply is chopped at very high frequencies (20 to 200 kHz using present technologies) then filtered and smoothed to produce dc outputs. A power supply which converters AC line voltage to DC power performs the following functions with high frequencies and at low cost.

- *Rectification*: Convert the incoming AC line voltage to DC voltage.
- *Voltage transformation*: Supply the correct DC voltage level(s).
- *Filtering*: Smooth the ripple of the rectified voltage.
- *Regulation*: Control the output voltage level to a constant value irrespective of line, load and temperature changes.
- *Isolation*: Separate electrically the output from the input voltage source.
- *Protection*: Prevent damaging voltage surges from reaching the output; provide back-up power or shut down during a brownout.

The circuit configuration, which determines how the power is transferred, is called the *TOPOLOGY* of the S.M.P.S., and is an extremely important part of the design process[2]. The topology consists of an arrangement of transformer, inductors, capacitors and power semiconductors (bipolar or MOSFET power transistors and power rectifiers). Presently, there is a very wide choice of topologies available, each one having its own particular advantages and disadvantages, making it suitable for specific power supply applications. Basic operation, advantages, drawbacks and most common areas of use for the most common topologies are discussed in the following sections.

This thesis work is divided into 5 chapters. Chapter II discusses the basic functional blocks of a switched mode power supply and the theory of operation of basic types of DC-DC converters. Chapter III describes the operation of a forward converter and its theoretical aspects and complete details of design of a forward converter based switch mode power supply. Chapter IV presents simulation results of a basic forward converter and single phase rectifier and testing results of a reference power supply. Chapter V gives conclusions and the future trends in switch mode power supply design.

CHAPTER – II

FUNDAMENTALS OF SWITCHED MODE POWER SUPPLY AND DC-DC CONVERTERS

2.1 Switched mode power supply functional block diagram

2.1.1 Introduction

Switch mode power supplies (SMPS) have been used for many years in industrial and aerospace applications and also in consumer products where good efficiency, light weight, low cost and small size were of prime concern. In addition to these, some power supplies require electrical isolation between the source and load, low harmonic distortion for the input and output waveforms, and high power factor if the source is ac voltage. Some special power supplies require controlled direction of power flow.

The building blocks of a typical high frequency off-the-line switching power supply are shown in figure 2.1.

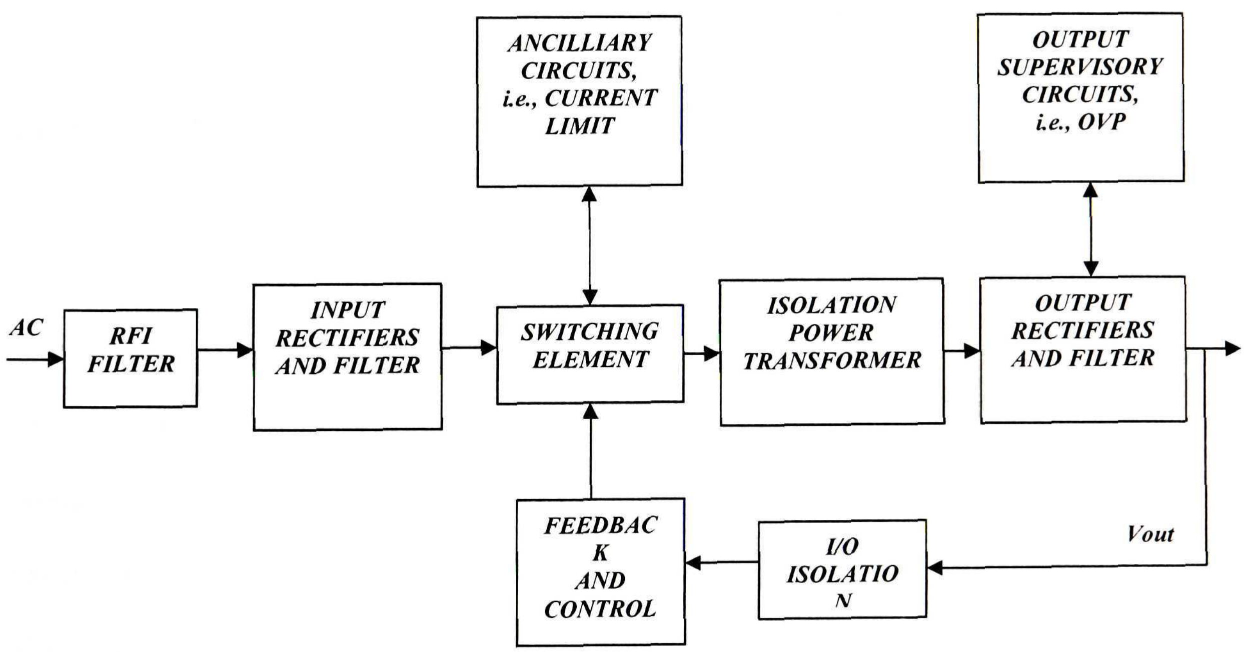


Figure 2.1. The building blocks of a typical off-line switched mode power supply[3]

2.1.2 The Input Section

The "direct-off-line" switch mode supply is so called because it takes power input directly from the ac power lines, without using the rather large low-frequency (60 to 50 Hz) isolation transformer normally found in linear power supplies[1]. The AC supply lines can be a source of large transient voltages. Power supplies are required to protect themselves and the end equipment from these large transient voltages. To meet this need requires special *Input transient voltage protection devices.* The most common suppression device used for this protection is the metal oxide varistor (MOV) type transient voltage suppressor, and it may be connected across the AC input line. Input fuse selection is also an important function for *Over load protection methods*. Switching power supplies are electrically noisy, and source of *Radio Frequency Interference (RFI)*. To meet various national and international RFI regulations for conducted-mode noise, a differential and common-mode noise filter is usually fitted in series with the line inputs. *Faraday screens* are used for high-frequency conducted-mode noise, which is normally caused by capacitively coupled currents in the ground plane or between input and output circuits.

2.1.3 Line Rectification and Capacitor Input Filters

The off line switched mode power supply rectifies the input ac supply without the use of a low frequency line isolation transformer between the ac mains and rectifiers. The rectified voltage is then filtered by the input reservoir capacitor to produce a rough dc input supply. This level can fluctuate widely due to variations in the mains. In addition, the capacitance on the input has to be fairly large to hold up the supply in case of a severe droop in the mains.

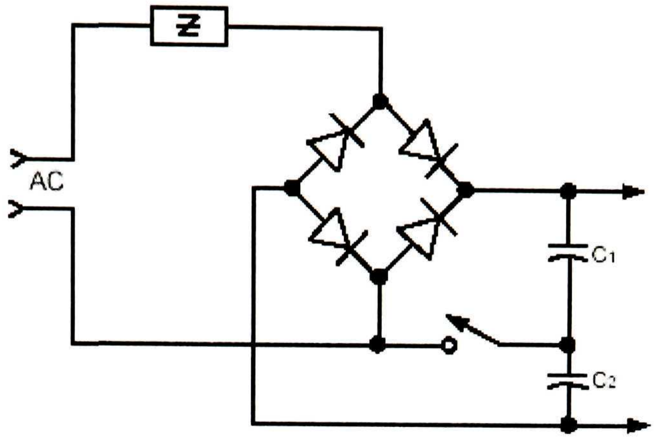


Figure 2.2 Basic line rectifier and filter operating from 110/220 input AC line[4]

Most of the switch mode power supplies have input line rectification and filter capacitor blocks as shown in figure 2.2. The input section uses a voltage doubler technique since today's most of the electronic equipment is manufactured to work with in all input voltage levels existing across the world. When the switch is closed, the circuit may be operated at a nominal line of 110V AC. The impedance shown in the figure may be a thermistor or resistor. The impedance limits the inrush current and it should be designed such that it causes less power loss across it. To decrease the series power loss, the impedance is often shunted by a triac after start-up. The input section also has two capacitors as part of the voltage doubling arrangement. The capacitors are connected in series when the circuit is operated with 220V AC input line.

During the positive half cycle of the AC, capacitor C1 is charged to the corresponding peak voltage (~170V DC) through the forward biased diode while during the negative half cycle, capacitor C2 is charged to 170V DC. Therefore the resulting output DC voltage from the rectifier will be the sum of two capacitors (~340V DC). With the switch open, the circuit configures into a full bridge rectifier capable of rectifying 240V AC line and giving out the same DC output voltage of 340 V DC. The input filter capacitor is an important element in switch mode power supply design because it prevents excessive transient voltage from appearing across the DC bus and also carries the line to neutral transient pulse current.

Designers have to be very careful while selecting bridge rectifier (or diodes) and the filter capacitor. For diodes, maximum forward rectification current capability, peak inverse voltage capability, and the surge current capability are the most important specifications. For the filter capacitor, normally high-grade electrolytic capacitors with high ripple current capacity and low equivalent series resistance (ESR) are used.

2.1.4 Input protective devices

Inrush limiting: In direct-off-line switched mode power supplies, there may be a high flow of currents into the input terminals when the supply is first switched on. Almost every SMPS uses semiconductor rectifiers and low impedance input electrolytic capacitors in input filter configuration. The capacitors are initially discharged and when the supply is turned on, charging of filter capacitor would cause very large surge currents into the SMPS. Therefore, it is a common practice to employ inrush limiting techniques to limit the surge currents. Two widely used methods for this purpose are a resistor-triac arrangement for high-power systems and negative temperature coefficient (NTC) thermistors for low-power systems. These devices when connected in series with the supply lines increases the input impedance during turn on.

In the resistor-triac technique, a resistor is placed in series with the AC line and it is shunted by a triac or SCR. This arrangement shorts the resistor when the input filter capacitors have been fully charged. The other method uses a NTC thermistor placed on either the AC lines or the DC buses after input AC line rectification. When the supply is switched on, the high impedance thermistor limits the inrush current. Input section with all protective devices is shown in figure 2.3.

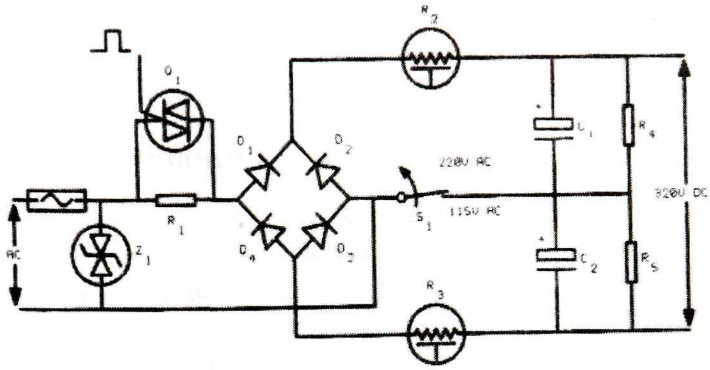


Figure 2.3 Input section of an SMPS showing basic components[3]

2.1.5 Switching element

The unregulated dc voltage produced after the input line rectification and capacitor filtering is fed directly to the central block of the supply, the high frequency power switching section. Fast switching power semiconductor devices such as MOSFETs and Bipolars are driven on and off, and switch the input voltage across the primary of the power transformer. The drive pulses for the switches are normally fixed frequency (20 to 200 kHz) and variable duty cycle. Hence, a voltage pulse train of suitable magnitude and duty ratio appears on the transformer secondaries.

The choice of a switch is influenced by many factors such as cost, peak voltage and current, frequency of operation, and heat sinking. The Bipolar Junction Transistor (BJT) is generally opted for low cost or high power switching applications but the maximum frequency of operation is less than 80 – 100 kHz. IGBTs also offer advantages such as low cost and high power applications but switching frequency is still a limitation. Power MOSFETs are used in the majority of applications due to their ease of use and their higher frequency capabilities. Each technology has its own merits and drawbacks. The designer has to make a choice between them depending on his application and other design criterion.

2.1.6 Isolation power transformer

Magnetic elements are mainstay for efficient operation of all switching power supply designs but are also the least understood. There are three types of magnetic components used in an SMPS: a forward-mode transformer or a flyback-mode transformer, an AC filter inductor, and a DC filter inductor or a magnetic amplifier, etc. These elements are used either for stepping-up or down a switched AC voltage, or for energy storage. In forward-mode topologies, the transformer is used for step up/down the AC voltage generated by the power switches and the output filter (output inductor and capacitor) is used for energy storage. In boost-mode topologies, the transformer is used for step up/down and energy storage as well.

The general procedure to design a magnetic element is as follows:

- Select an appropriate core material depending on the application and operating frequency.
- Select a core form factor that is appropriate for the application and that satisfies applicable regulatory requirements.
- Calculate the core cross-sectional area to meet the power ratings of system.

- Determine the airgap necessity and calculate the number of turns required for each winding. Make sure that the output specification is met with the choices made and whether the winding will fit into the selected core size.
- Wind the magnetic component using proper winding techniques.
- During the prototype stage, verify the components operation with respect to level of voltage spikes, cross-regulation, output accuracy and ripple, RFI, etc. and make correction if necessary.

2.1.7 Output section

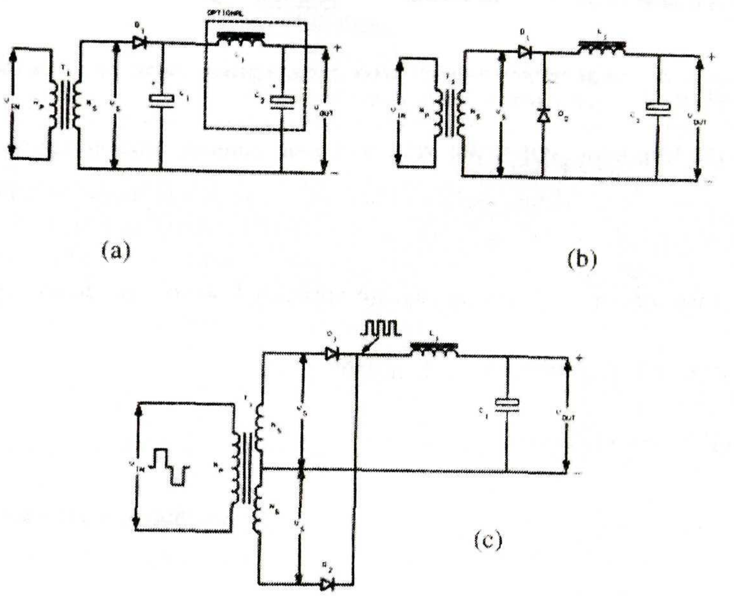


Figure 2.4 Output stages (a) Flyback type (b) Forward type (c) Push-pull/half-bridge/full-bridge[3]

The output section of an SMPS consists of high speed switching diode rectifiers and LC filters. The output section may have single or multiple DC voltages. The transformer secondary voltages are appropriately rectified, and then smoothed by the output filter, which is either a capacitor or capacitor / inductor arrangement, depending upon the topology used. This rectified and filtered power is used to drive electronic components and circuits. Typical output DC voltages are 5V, 12V, 15V, 24V, or 28V DC, and their power capability varies from a few watts to thousands of watts. This transfer of power has to be carried out with the lowest losses possible, to maintain efficiency. Thus, optimum design of the passive and magnetic components, and selection of the correct power semiconductors is critical.

The selection of the output section depends on the topology used. The conventional flyback converters output section is shown in figure 2.4a. The transformer in flyback converter acts as an energy storage element also. So, the output section in a flyback converter consists of a diode and capacitor only. For some practical applications an additional LC filter is also needed as shown in the figure to suppress high-frequency switching spikes.

The output section of a forward mode converter is shown in figure 2.4b. Compared to the flyback converter, the forward converter has an extra freewheeling diode and also a series connected inductor. Freewheeling diode provides current to the output during the switch-off period.

Figure 2.4c shows the output section used for push-pull, half-bridge, and full-bridge converters. In this topology, each of the two diodes provides current to the output for approximately half of the cycle. A freewheeling diode is not necessary in this scheme since either diode acts as a flywheel when the other one is turned off.

2.1.8 Output supervisory circuits

Output overvoltage protection: Loss of voltage control can cause excessive output voltages. In direct-off-line switched mode power supply, the transformer isolates output from the input. Therefore, most failures result in a low or zero output voltage. An independent signal level voltage clamp can be a satisfactory for overvoltage protection which acts on the converter drive circuit.

Output undervoltage protection: Excessive transient current demands and power outages can cause output under voltages. In switch mode supplies, the energy stored in input capacitors provides holdup of the outputs during short power outages. Nevertheless, transient current demands can still cause undervoltages as result of limited current ratings and output line voltage drop. Active undervoltage prevention circuits can be a solution for output undervoltages.

Overload protection (Input power limiting): Power limiting is usually applied to the primary circuit and it limits the maximum throughput power of the power converter. In multiple output converters the sum of the individual output current limits often has a total VA rating in excess of the maximum converter capability. Though output current limiting would prevent output overloading conditions, primary power limiting is often provided as additional backup protection.

Output current limiting: Each output line in an SMPS is independently current limited and it protects the supply under all conditions to short-circuit. Continuous operation in a current-limited mode should not cause over dissipation or failure of the power supply. Naturally, SMPS does not dissipate excessive power under short-circuit conditions and it does not cause problems like "lockout" under nonlinear or cross-coupled load conditions.

2.1.9 Input to Output (I/O) Isolation

Input/output isolation is essential to an off-the-line SMPS. The isolation may be optical or magnetic. These are often used to convey information from the secondary output circuits back to the input primary control circuits without compromising the galvanic isolation between the two. In transformer-coupled supplies, in order to keep the isolation barrier intact, the required electronic isolation in the feedback is usually achieved by using a small pulse transformer or an opto-isolator.

2.1.10 Feedback and Control

Regulation of the output to provide a stabilized dc supply is carried out by the control / feedback block. Generally, most S.M.P.S. systems operate on a fixed frequency pulse width modulation basis, where the duration of the on time of the drive to the power switch is varied on a cycle-by-cycle basis. This compensates for changes in the input supply and output load. The output voltage is compared to an accurate reference supply, and the error voltage produced by the comparator is used by dedicated control logic to terminate the drive pulse to the main power switch/switches at the correct instance. Correctly designed, this will provide a very stable dc output supply. It is essential that delays in the control loop be kept to a minimum, otherwise stability problems would occur. Hence, very high-speed components must be selected for the loop.

2.2 Basic principles of DC to DC conversion and Topologies

DC to DC conversion is the heart of any switched mode power supply. The following discussion presents a brief overview of the DC to DC converter fundamentals, and a simplified analysis for input to output relationship in different topologies. There are two major modes of conversion: forward and flyback. Though the component arrangement in these two modes does not change much, their operating principles, advantages and applications may change considerably.

2.2.1 Forward Mode Converters

These converters have an L-C filter section directly after the power switch or after the output rectifier. The operation of the converter can be divided into two periods:

- Power switch ON period: During this period, the input voltage is presented to the L-C section, and inductor current ramps up linearly. Energy is stored in inductor.
- Power switch OFF period: During this period, the voltage at the inductor input point goes below ground since inductor can not support instant change in current. Then the freewheeling diode becomes forward biased. The energy that was stored in inductor is dumped on to the load. The inductor current is negative linear ramp as shown in figure 2.5.

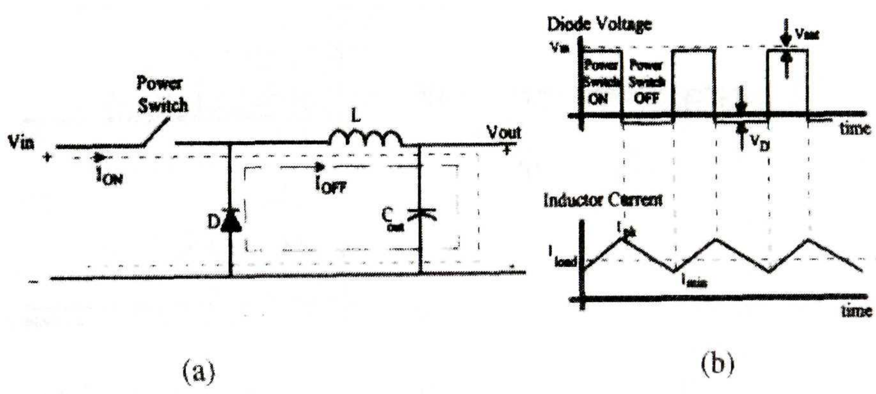


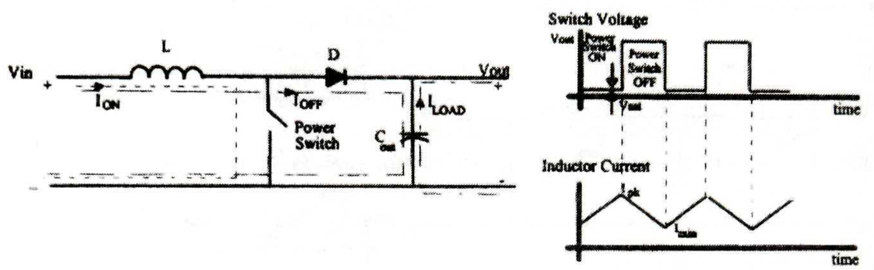
Figure 2.5 The forward mode converter (a) The basic circuit (b) Associated waveforms[3]

Forward mode converters have lower output peak-to-peak ripple voltages, and higher levels of output power.

2.2.2 Flyback mode converters

The inductor is placed between the input source and the power switch.

- Power switch ON period: During this period, current flows through inductor, power switch back to the input source. The inductor current is a positive ramp and energy is stored in the inductors core.
- Power switch OFF period: During switch turn-off, the inductor voltage flies back above the input voltage and forward biases the diode. Then the inductor voltage is clamped at the output voltage. The inductor current is a negative ramp.



(a) (b)

Figure 2.6 The flyback mode converter (a) The basic circuit (b) Associated waveforms[3]

The converter is said to be operating in continuous conduction mode if the inductor current does not go to zero during flyback period. The core's flux of the inductor is not completely emptied during the switch-off period and there will be a residual flux remaining after every cycle of operation. This might lead to instability in control of the converter. Therefore, flyback mode converters are preferred to be operation in discontinuous conduction mode. Moreover, the output filter capacitor is the only element of energy storage. This makes flyback converters to have higher output ripple voltage compared to forward mode converters. The power output is also lower because of higher peak currents that are generated when the inductor voltage flies back. Despite these drawbacks, flyback mode converters are popular in low-to-medium power applications owing to the fewest number of components.

2.2.3 The volt-second balance for inductors

Power electronics circuits can be characterized by the following five laws:

- The average voltage across an inductor over a complete switching cycle in the steady state is zero.
- The average current through a capacitor over a complete switching cycle in the steady state is zero.
- Voltage across a capacitor is a continuous function. (Capacitors do not make discontinuous voltage changes.)
- Current through an inductor is a continuous function. (Inductors do not make discontinuous current changes.)
- Energy balance is always true.

The input-output relationships for all the DC to DC converters can be derived from the volt-second balance for inductors. The initial and final inductor current within a cycle should be the same for steady state operation. It means the net energy stored in each inductor within one switching cycle should be zero.

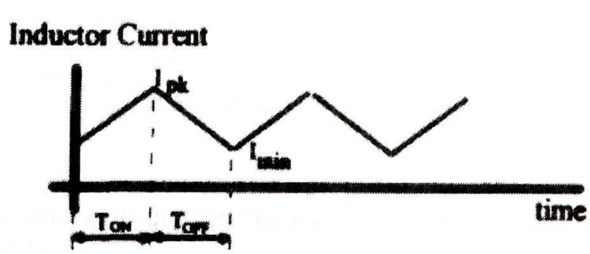


Figure 2.7 A typical inductor current waveform[3]

A typical inductor current waveform is illustrated in figure 2.7. Let the positive slope of the current be m_1 and m_2 be the negative slope. For the initial and final currents to be the same:

$$m_1 T_{on} - m_2 T_{off} = 0. \quad 2.1$$

The voltage across an inductor L is given by L.di/dt, therefore

$$V_{Lon} = L\, m_1, \quad 2.2$$

$$V_{Loff} = -\, L\, m_2, \quad 2.3$$

where V_{Lon} and V_{Loff} are the voltages across the inductor during the switch ON and OFF periods respectively.

From the above three equations,

$$V_{Lon}\, T_{on}/ L + V_{Loff}\, T_{off} / L = 0. \quad 2.4$$

The duty cycle of the converter is given by,

$$D = T_{on} / (T_{on} + T_{off}). \quad 2.5$$

Combining equations 2.4 and 2.5 gives,

$$V_{Lon}\, D + V_{Loff}\, (1 - D) = 0. \quad 2.6$$

The output voltage of the forward mode converter is given approximately by $V_{out} = D.V_{in}$. Therefore a forward mode converter always performs step-down operation. The output voltage of the flyback mode converter is given by $V_{out} = V_{in} / (1 - D)$. Therefore a flyback converter always performs step-up operation.

2.3 Converter topologies

Topology refers to the arrangement of components within the converter. There are two major categories in converters: transformer-isolated and non-transformer-isolated. Each category has many different topologies, with some available in both the forms.

Non-transformer-isolated topologies

These topologies are used when there is a 50-60 Hz transformer or bulk power supply provides the DC isolation and protection. Many state of art power supplies use the following three non-transformer-isolated converters: the buck, the boost, and the buck-boost. These are the simplest configurations possible with lowest component count, requiring only one inductor, capacitor, transistor, and diode to generate single DC output. But at the same time they are prone to failure due to lack of DC isolation. If required, an isolation transformer can be inserted before the converter.

2.3.1 The Buck Converter

This is the most elementary forward-mode converter. The topology and associated waveforms are shown in figure 2.8.

When the switch is ON, the input voltage is applied to the inductor and power is delivered to the output. When the switch is OFF, the voltage across the inductor is reversed. This makes the freewheeling diode forward biased which allows the energy stored in inductor during switch ON period to be delivered to the output. This continuous current is then smoothed by the output filter capacitor.

Applying equation 2.6 to the buck converter waveforms shown in figure 8 gives the input to output relation.

$$(V_{in} - V_{out})D - V_{out}\,(1 - D) = 0. \qquad 2.7$$

Therefore,

$$V_{out} = D.V_{in.} \qquad 2.8$$

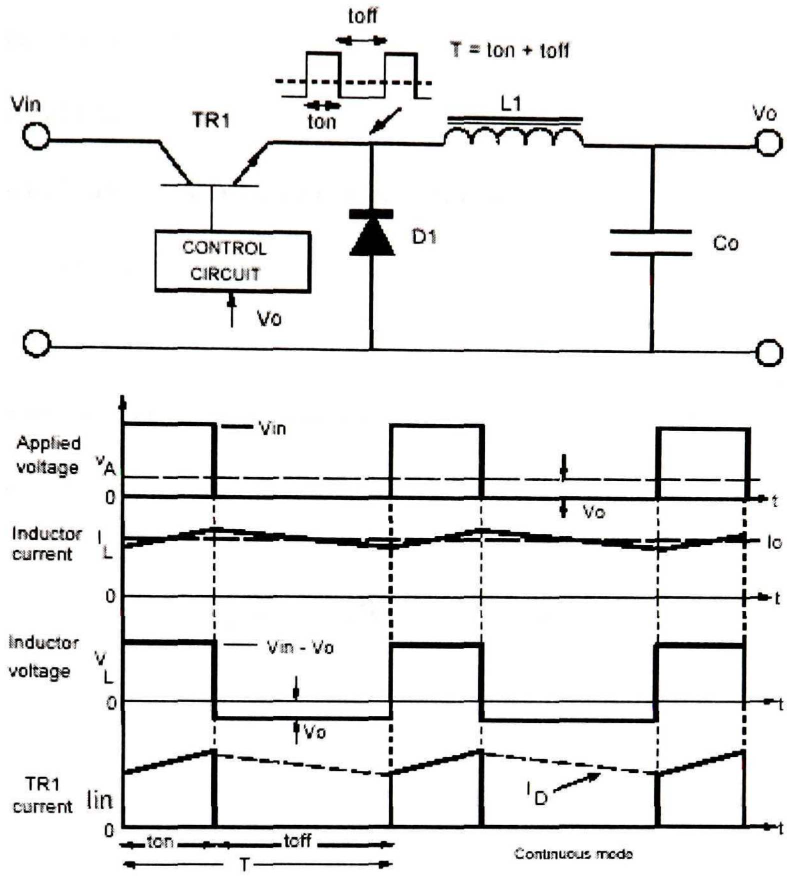


Figure 2.8 Buck converter topology and associated waveforms[5]

The buck is a step-down transformer since D never reaches one. Output voltage is controlled by varying the duty cycle. The L-C filter section provides effective smoothing of the inductor current. So, the buck and its derivatives have very low output ripple characteristics. Buck converter is normally operated in continuous conduction mode where the peak currents are lower and the filter capacitor requirements are small.

The buck converter has some limitations also which will put some restriction to be considered as the popular choice among experienced designers:

- The input voltage must be always 1 or 2V higher than the output voltage to maintain regulation.

- The finite reverse recovery time of the diode presents an instantaneous short circuit across the input source when the power transistor is switched ON. This results in adding more stress to the power switch and diode.
- If the switch fails in short-circuited conditions, the input is short circuited to load. Additional protective measures are needed to prevent this hazard.

2.3.2 The Boost Converter

The Boost converter is an elementary flyback mode converter. The converter topology and waveforms are shown in figure 2.9.

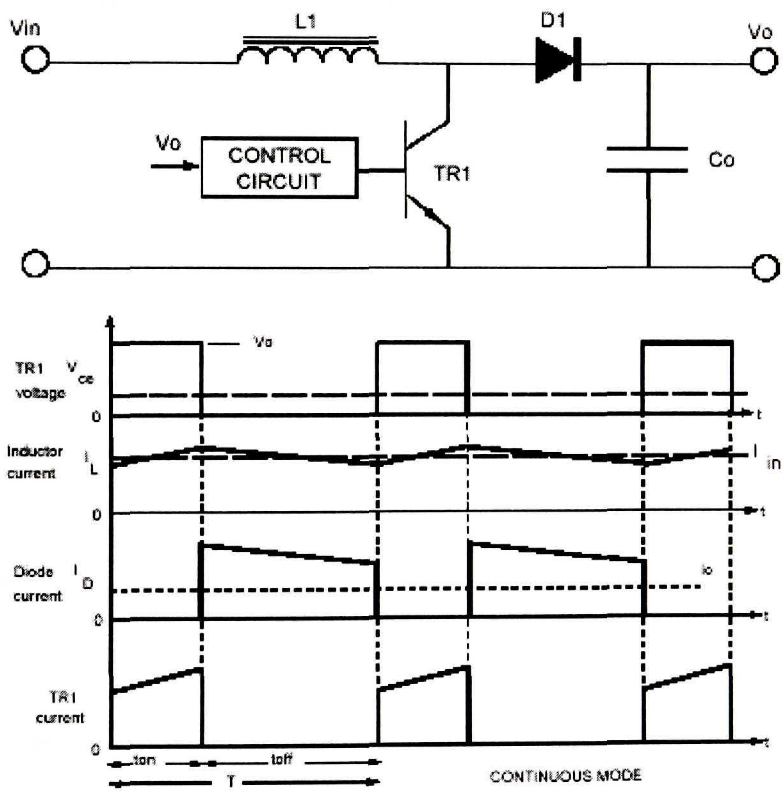


Figure 2.9 The boost converter and associated waveforms[5]

When the switch is turned ON, the diode is reverse biased, and the input voltage is applied to the inductor. The inductor current ramps up to a maximum point from zero in discontinuous conduction mode, and from an initial value in continuous conduction mode. When the switch is turned OFF, the voltage across the inductor reverses causing the diode to forward bias. Then the energy stored in inductor during switch on period and the input supply energy is delivered to the filter capacitor and load.

Equation 2.6 can also be applied to boost converter to derive the input-to-output relation.

$$V_{in}\, D + (V_{in} - V_{out})\,(1 - D) = 0. \qquad 2.9$$

Therefore, $$V_{out} = V_{in} / (1 - D). \qquad 2.10$$

The Boost converter always steps-up the input voltage since D never reaches one. The output voltage is controlled by changing the duty ratio of the switch. The current supplied to the filter capacitor is diode that is always discontinuous as shown in the figure 9. Hence, the filter capacitor has to be large with low ESR to lower the output ripple to acceptable level. On the other hand, the boost input current is the continuous inductor current with lower input ripple characteristics. This makes the boost converter a popular choice for capacitive load applications and power factor improvement circuits as the pre-regulator.

Boost has problems in both continuous and discontinuous modes of operation. In discontinuous mode, the peak transistor and diode currents will be higher, and a very large output capacitor is needed. Moreover the output voltage becomes dependent on the load, resulting in poorer load regulation. When used in continuous mode, major control and regulation problems will arise due to a complex second order characteristic in the small signal control response caused by pseudo LC filter. These problems can be eliminated with discontinuous mode of operation.

2.3.3 The Buck-Boost Converter (Non-Isolated Flyback Regulator)

The topology is shown in figure 2.10. It is a flyback-mode converter whose operation and waveforms are similar to that of boost converter except that the transistor switch now has to support the sum of both input and output voltages across it. And also, the positions of the inductor and the switch are reversed in the buck-boost compared to the boost converter. Energy is delivered to the output only during switch OFF period in this topology.

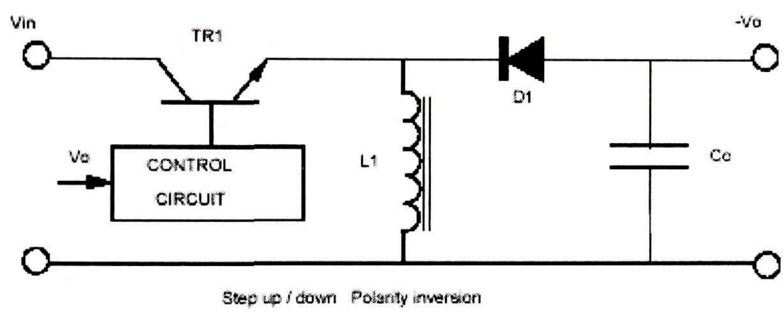


Figure 2.10 The Buck-Boost converter[5]

When the switch is ON, the inductor stores energy and releases it through the diode into the output capacitor and load during the switch OFF period. This results in negative voltage whose value is controlled by varying the duty cycle of the switch. Both the input and output current waveforms must be discontinuous.

Analyzing the circuit using equation 2.6,

$$V_{in} D + V_{out} (1 - D) = 0, \qquad 2.11$$

$$V_{out} = - V_{in} D / (1 - D). \qquad 2.12$$

The duty ratio can be varied below 50% for voltage step-down and above 50% for step-up. It is also called as inverter topology since the polarity of the output is opposite to that of input voltage. The buck-boost topology also suffers from the same continuous mode control problems as the boost, and discontinuous mode is often favored.

Transformer-Isolated Converter Topologies

Transformer isolation in DC to DC converters has the following advantages to their non-transformer-isolated versions:

- Input to output isolation is provided
- Outputs can be widely different from the input by selecting proper transformer turns ratio. Duty cycle can be optimized and the peak currents flowing minimized.
- Multiple outputs can be added to the power supply without using separate regulators. The polarity of each output can be selected, depending upon the polarity of secondary with respect to the primary.

These features make the isolated topologies always a better choice. But they have some disadvantages such as additional size, weight, and power loss. The generation of voltage spikes due to leakage inductance may also be a problem.

2.3.4 The Flyback Converter

The flyback converter topology and waveforms are shown in figure 2.11. Single-ended flyback converter is the simplest of all isolated converters. The use of a single switch makes the converter asymmetrical i.e. the transformer can be driven only in one direction, which requires a large core size. The flyback converter does not really contain a transformer but a coupled inductor which performs the dual function of transformer and inductor.

When the switch is turned ON, the full input is applied to the primary winding of transformer and the current flowing through it increases linearly storing energy in the core. During the switch OFF period, the diode starts conducting, delivering the stored energy to the capacitor and the load. During this flyback period, the secondary current is a negative ramp. The flyback converter can operate in either the continuous mode or the discontinuous mode.

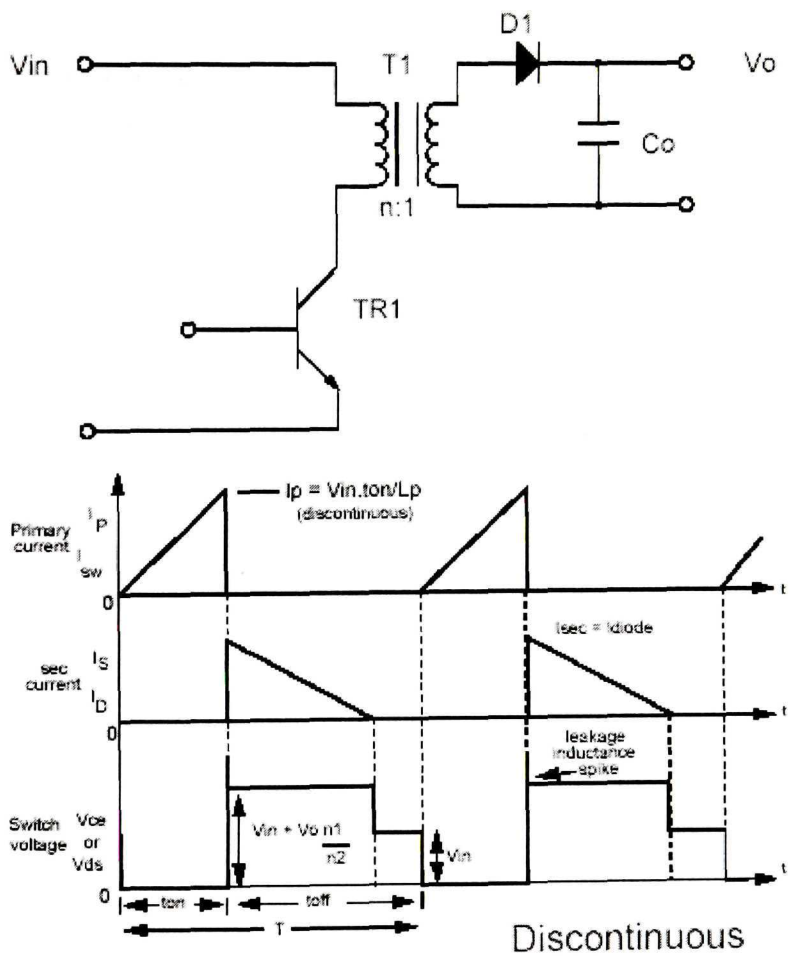


Figure 2.11 The flyback converter topology and associated waveforms[5]

The DC input and output relations are given by,

- Continuous mode

$$V_{out} = n\, V_{in}\, D/\,(1 - D). \qquad 2.13$$

- Discontinuous mode

$$Vout = DVin\sqrt{\frac{R_L T}{2L_P}} \qquad 2.14$$

Advantages

- No filter inductor is needed in the output circuit since the transformer secondary inductance is in series with the diode when current is delivered to the load. This makes the flyback converter suitable for generating large output voltages without needing large filter inductor whose function is to reduce the ripples in output currents.
- Each output requires only one diode and capacitor contributing to lowest component count and hence low cost. Multiple outputs are also possible with this configuration.
- Very good cross regulation in multiple output sections due to the absence of output choke.

Disadvantages

- The output capacitor has to smooth out large pulsating output currents since it is supplied only during the switch OFF period. This needs a large output capacitor with very low ESR.
- Higher output ripples and peak currents, larger capacitors and transformers limit the flyback converter to lower output power applications in the 20 to 200W range.

2.3.5 The Push-Pull Converter

The push-pull is a double ended topology i.e. two transistors share the switching function. The transformers flux swing is fully utilized in this topology by symmetrical operation of its core. This feature is useful in using much smaller transformer sizes and in providing higher output powers. The circuit and waveforms are shown in figure 2.12.

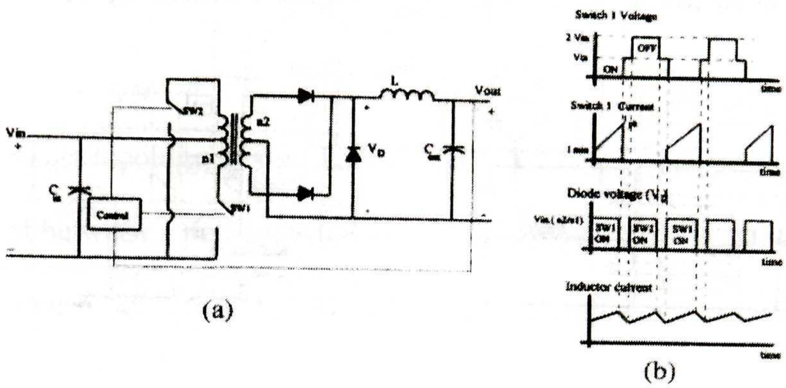


Figure 2.12 The Push-Pull converter topology and associated waveforms[3]

The input is applied to the center tapped arrangement and each transistor switch is driven alternately, driving the transformer in both directions. Power is transferred to the buck type output circuit during each transistor conduction period. No core reset arrangement is needed since the current flows in opposite directions in the primary winding during each switch's ON period. There is a dead time provided where neither transistor is conducting, to allow for the finite time necessary for the power switches to turn off. Push-pull is normally operated in continuous conduction mode.

The push-pull topology offers compact design of the transformer and output filter, while producing very low output ripple. Each switch operates with nearly 45% duty ratio giving the total switching frequency of the converter a very high operating frequency of 90%. Thus, push-pull converters are excellent for high power density, low ripple outputs.

Nevertheless, push-pull topology has some disadvantages also. Each transistor in the converter has to block twice the input voltage due to doubling effect of the center-tapped primary. Another major problem is that neither the two halves of the winding nor the two power switches can be absolutely identical due to real-world factors. This causes flux symmetry imbalance.

2.3.6 The Half-Bridge Converter

The half-bridge topology is also double ended. It has only one primary winding which is connected between a pull-up/pull-down configuration of power switches and at the center node between two series capacitors. The topology and waveforms are shown in figure 2.13.

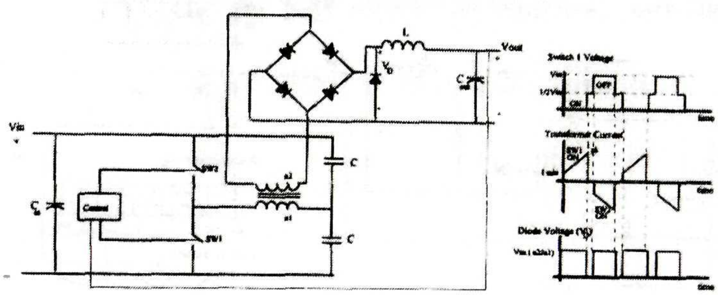


Figure 2.13 The Half-Bridge converter and associated waveforms[3]

The capacitor center node is fixed approximately at half of the input voltage and the other end of the primary is alternately supplied with the input voltage and ground by the power switches. Hence, only half of the input voltage is applied to the primary winding at any time. Power is transferred directly to the output on each transistor conduction time and a maximum duty cycle of 90% is available. A full wave output buck filter is implemented rather than an half wave since the primary is driven in both directions.

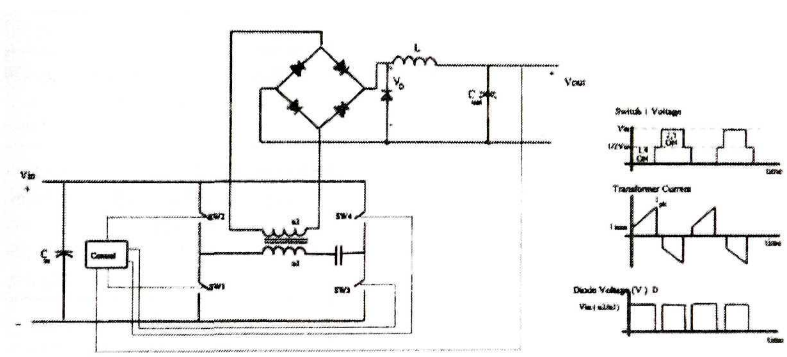
The input and output relation is given as:

$$V_{out} = n.D.V_{in.} \tag{2.16}$$

The major advantages with this topology compared to push-pull converters are reduced stress on transistors, reduced saturation problems, and provision for implementing a doubling circuit. The major drawbacks are need for two 50/60 Hz input capacitors and requirement of an isolated gate drive for the top transistor.

2.3.7 The Full-Bridge Converter

The full-bridge converter topology and associated waveforms are shown in figure 2.14. The balancing capacitors in the half bridge are replaced with another pair of switches to form a full-bridge topology. The transistors are driven in pairs and the full input voltage is applied to the primary winding. The output rectification function is similar to those of push-pull and half-bridge topologies.



(a) (b)

Figure 2.14 The Full-Bridge converter and associated waveforms[3]

The input to output relation is given as:

$$V_{out} = 2.n.D.V_{in.} \qquad 2.17$$

The advantages of this topology are increased output power levels, and the need for only one main smoothing capacitor compared to two for the half-bridge. The major disadvantage is the need for four transistors out of which two need isolated gate drives.

CHAPTER – III

FORWARD CONVERTER DESIGN AND THEORETICAL ANALYSIS

3.1 Introduction

The single transistor forward converter is an isolated version of normal buck converter with added transformer and an extra diode in the output circuit. The forward converter is a popular choice for power supplies in the 100–300W range and finds itself as the best solution for DC–DC applications in industrial controls, telecommunication systems, digital feature phones, and in distributed power systems. Due to the transformer, the forward topology can be used as either an up or a down converter, although the most common application is down conversion.

3.2 Topology and operating principles

The topology of single ended forward converter with auxiliary winding reset is shown in figure 3.1. Unlike the flyback converter, forward converter employs a separate transformer and output filter inductor for voltage isolation/conversion and energy storage respectively. This arrangement permits more favorable trapezoidal current waveforms and lower output current and voltage ripple, which consequently gives reduced noise and decreased stress on semiconductors and capacitors. Due to the nonpulsating characteristic of its output current, the forward converter is well suited for applications involving high output currents.

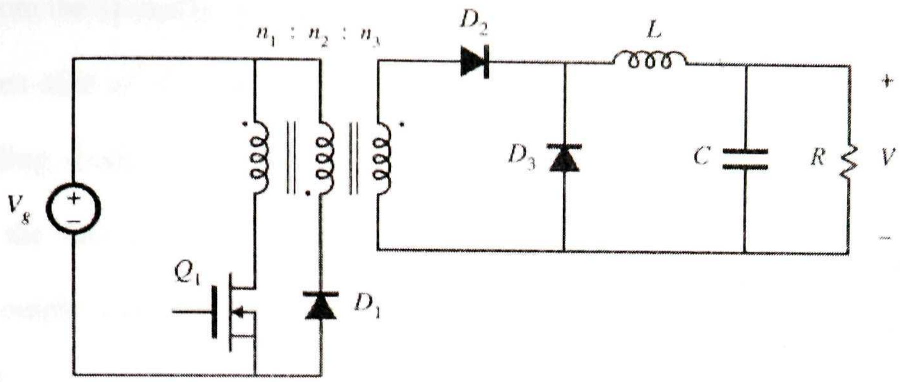


Figure 3.15 Single transistor forward converter[6]

Useful Power Level: 10 to 250 Watts
Switch Voltage Stress: 2 x V_{in}
Switch Power Stress: P_{in}
Transformer Utilization: Fair-good/no taps required
Duty Cycle: <0.5
Output Ripple Frequency: F
Relative Cost: Low

When the power switch is turned ON, energy transfer occurs across the isolation transformer. During the switch ON period, the primary voltage is reflected on to the secondary winding, and this voltage is rectified by the forward diode. The rectified voltage is then filtered and stored in the output LC section of the circuit. In the conventional three winding transformer single ended forward converter, the transformer reset occurs after the power transfer cycle, requiring the power transistor to block a minimum of twice the input voltage. Thus, when the primary switch turns off, the primary clamp winding conducts through the clamp diode by clamping the drain voltage of power transistor at twice the input voltage. During this transformer reset period, the energy from the magnetizing inductance of the transformer is returned to the primary bus. The driven side of the output inductor is clamped at 0.7 volts below the ground by freewheeling diode. The output inductor and the output capacitor store energy and integrate the duty cycle so that the output voltage is proportional to the product of the rectified output voltage and duty ratio. The waveforms for the converter are shown in figure 3.2.

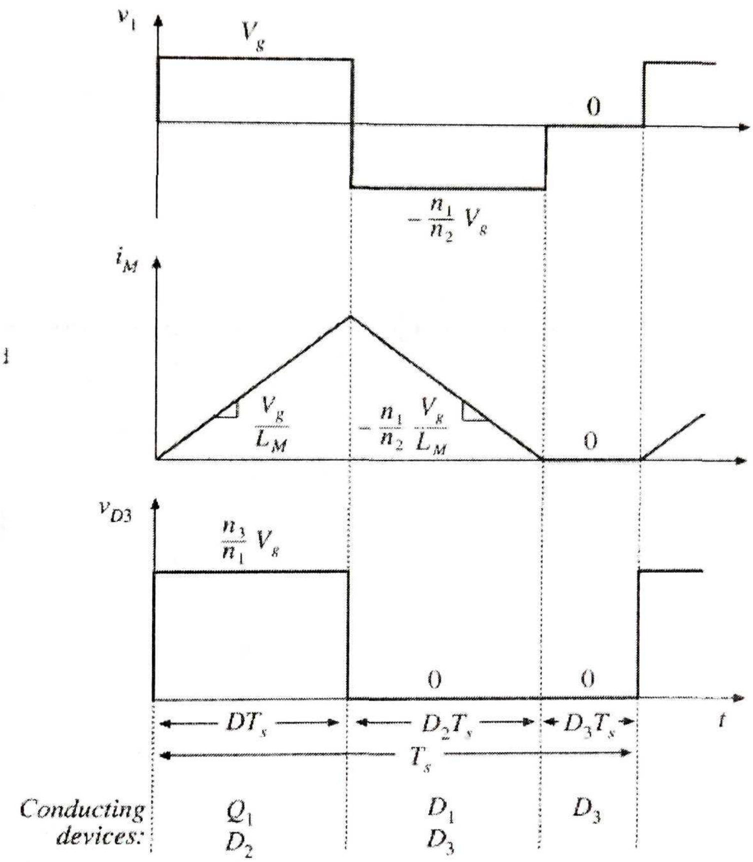


Figure 3.2 Waveforms of the forward converter[6]

If the primary winding and the reset windings are not wound at the same time, there will be a substantial unclamped leakage inductance on the primary. This leakage inductance in addition with the any associated stray inductance stores energy which is not reset by the reset winding. This energy storage in the transformer might lead to the saturation of its core. In this case, additional circuitry might be needed to dissipate the unwanted energy stored in the transformer. Protective snubber networks are often used for this purpose.

3.3 subintervals of operation

More insight into the forward converter operation and core reset is obtaining by replacing the three winding transformer in figure 15 with the equivalent circuit shown in figure 3.3.

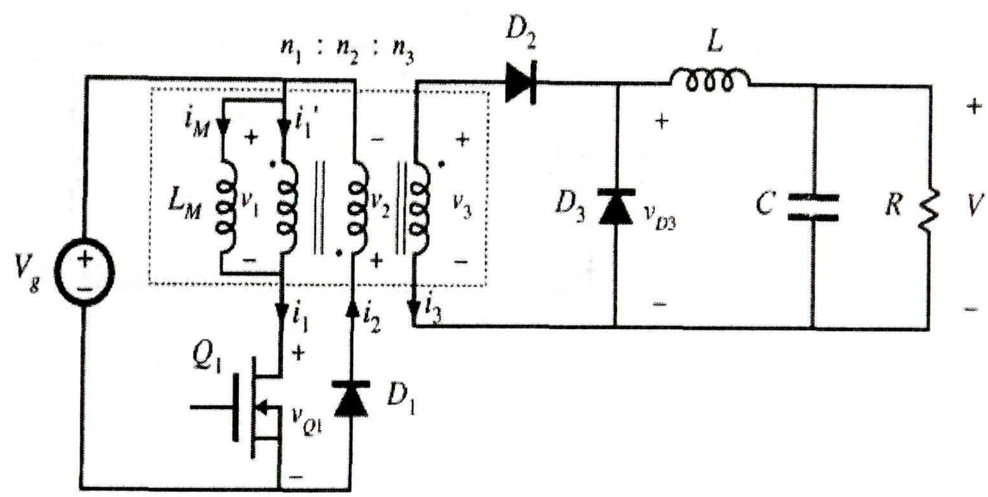


Figure 3.3 Forward converter, with transformer equivalent circuit model[6]

The magnetizing inductance of the transformer L_M is shown separately in parallel with the primary winding inductance. The magnetizing inductance must operate in the discontinuous conduction mode together with the clamping diode D_1. The output filter inductor L and the freewheeling diode D_3 can be operated in either continuous or discontinuous conduction mode.

The switching period of the converter can be divided into three subintervals.

Subinterval 1: Transistor Q_1 conducts and diode D2 is forward biased. The other diodes, the clamping diode and the freewheeling diode are reverse biased. The input DC voltage (Vg) is applied to the primary winding. The transformer magnetizing current $i_M(t)$ increases with a slope of V_g/L_M as illustrated in figure 3.2. The voltage across the output filter capacitor is the voltage V_g multiplied by the transformer turns ratio n_3/n_1.

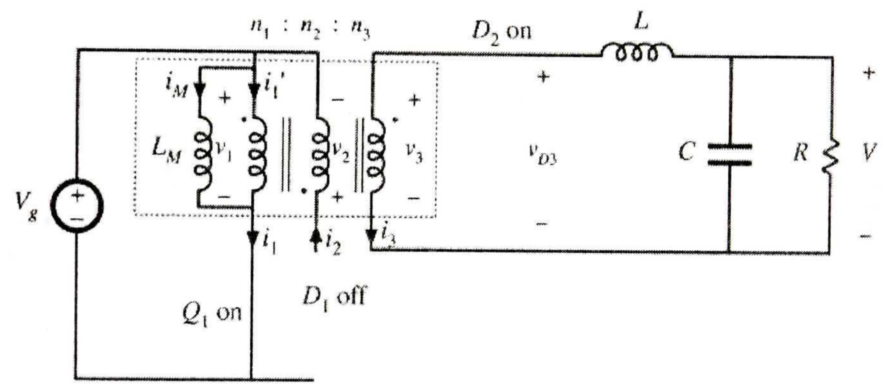


Figure 3.4 Equivalent circuit during subinterval 1 [6]

Subinterval 2: This interval starts when the power transistor is turned OFF. The equivalent circuit for this interval is shown in figure 3.5. The magnetizing current $i_M(t)$ is positive at this time and it has to continue to flow. The transistor Q_1 is OFF during this interval and hence the only path for the magnetizing current to flow is through the primary winding of the ideal transformer. It is evident from the equivalent circuit shown in figure 19 that the ampere-turns $n_1 i_M$ flow out of the polarity mark of the primary winding. According to ampere-turns balance principle, an equal number of total ampere-turns must flow into the polarity marks of the other windings. Diode D_2 prevents flow of current into the polarity mark of the winding 3. Therefore, the only possibility left for the ampere-turns balance principle is, a current $i_M n_1/n_2$ has to flow into the polarity mark of the winding 2. Diode D_1 is then forward-biased to allow this current flow, while diode D_2 is reverse-biased. The supply voltage V_g is applied to winding 2, and hence the voltage across the magnetizing inductance is $- V_g n_1/n_2$, referred to winding 1. The negative voltage across the magnetizing inductance forces the magnetizing current to decrease, with a slope of $- V_g n_1/n_2L_M$. During this interval, the freewheeling diode D_3 is forward biased to conduct the output inductor current.

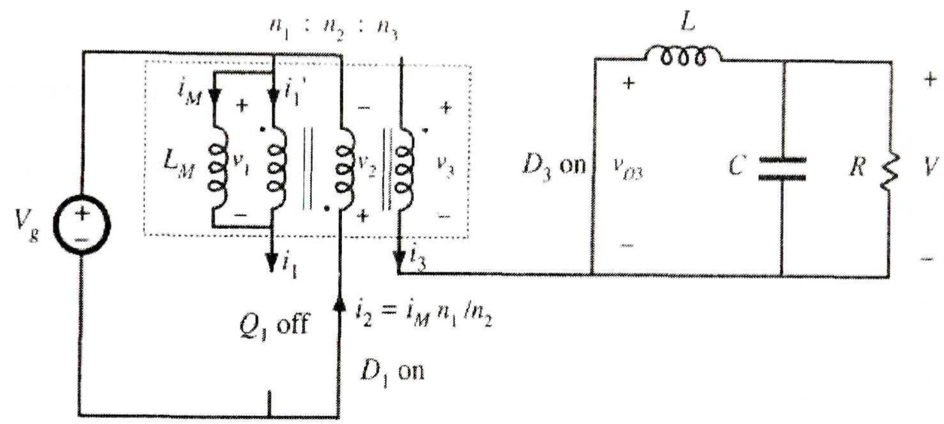


Figure 3.5 Equivalent circuit for subinterval 2 [6]

Subinterval 3: This interval begins when the magnetizing current reaches zero. Diode D_1 becomes reverse biased. The equivalent circuit is shown in figure 3.6. The transistor Q_1, and diodes D_1 and D_2 are all operate in OFF state. The magnetizing current remains at zero through out the subinterval 3.

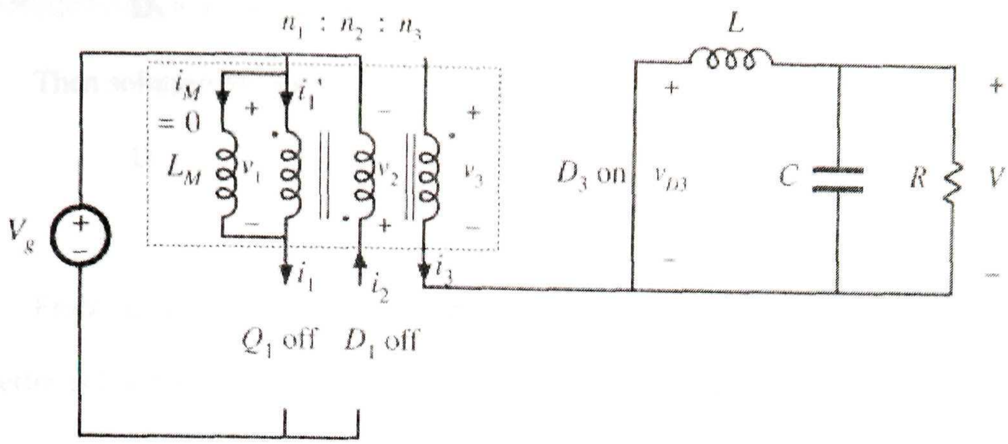


Figure 3.6 Equivalent circuit for subinterval 3 [6]

3.4 Limitation of maximum duty ratio

According to the inductor volt-second balance principle, the average voltage across every winding of the transformer must be zero. Applying this principle to the magnetizing inductance of the transformer, the average primary winding voltage $v_1(t)$ must be zero. From the $v_1(t)$ voltage waveform shown in figure 16, the average voltage can be equated to zero as,

$$\langle v_1 \rangle = D\,(V_g) + D_2\,(-\,V_g\, n_1/n_2) + D_3\,(0) = 0. \tag{3.1}$$

Solving for D_2 gives,

$$D_2 = (n_2 / n_1)\, D. \tag{3.2}$$

It is also know that,

$$D + D_2 + D_3 = 1. \tag{3.3}$$

Now we can write,

$$D_3 = 1 - D - D_2 \geq 0. \tag{3.4}$$

From the equations 3.2 & 3.4,

$$D_3 = 1 - D\,(1 + n_2/n_1) \geq 0. \qquad 3.5$$

Then solution for duty ratio D yields,

$$D \leq 1/\,(1 + n_2/n_1). \qquad 3.6$$

From the equation 3.6, it is evident that the maximum duty ratio for the forward converter is limited. The maximum duty ratio could be increased by decreasing the turns ratio n_2/n_1. This would cause the magnetizing current to decrease more quickly during second subinterval, resetting the transformer faster. But on the other hand, the decrease in n_2/n_1 ratio increases the voltage stress on the transistor. The maximum voltage is applied to the transistor during subinterval 2. Solution of the circuit of figure 19 for this voltage gives

$$\text{Max}\,(v_{Q1}) = V_g\,(1 + n_1/n_2). \qquad 3.7$$

For the normal choice of $n_1 = n_2$, the voltage applied to the transistor during second subinterval is $2V_g$. in practical converters, this voltage stress may be a bit higher due to the ringing associated with the transformer leakage inductance.

Therefore, the maximum duty cycle for $n_1 = n_2$ is given by,

$$D \leq 0.5.$$

If this limit is not satisfied, then the transistor off-time is not long enough to reset the transformer magnetizing current to zero before the end of the switching period. Then the core saturation might occur which might lead to the total failure of the converter circuit.

3.5 Magnetization current and output voltage

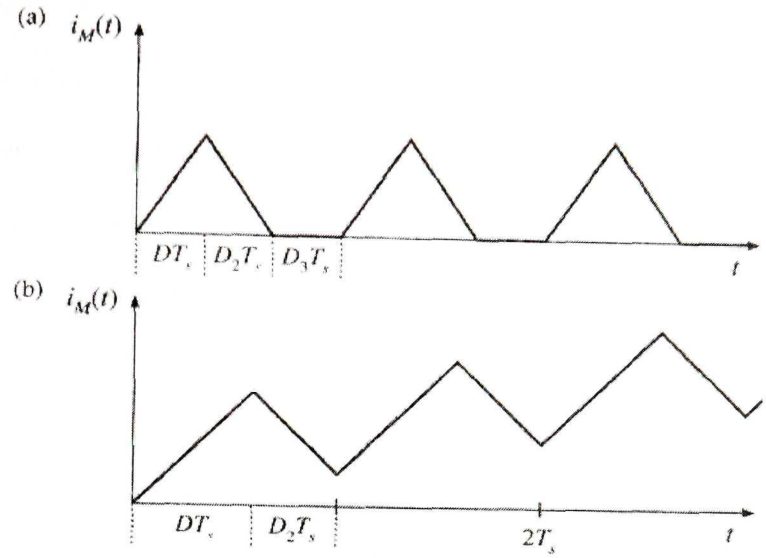


Figure 3.7 Magnetizing current (a) DCM, D < 0.5(b) CCM, D > 0.5 [6]

The transformer magnetizing current waveforms for two different ranges of duty cycles are shown in figure 3.7. The primary and reset winding turns are equal. Figure 21(a) illustrates operation with D≤0.5. The magnetizing inductance and the diode D_1 operate together in discontinuous conduction mode, and magnetizing current is reset to zero before the end of each switching cycle. Figure 3.7 illustrates the behavior of the magnetizing current when duty ratio exceeds 0.5. For duty ratios above 0.5, there will be no third subinterval in which the power transistor, the clamping diode, and the forward diodes are all OFF. The magnetizing inductance operates in continuous conduction mode. Apart from all these, the second subinterval is not long enough to reset the magnetizing current to zero. Therefore, the magnetizing current continues to increase by each switching period. At some point of time, the magnetizing current will reach a value where it can saturate the core.

The converter output voltage can be derived by applying the volt-second balance principle to the output inductor L. The average voltage across the inductor must be zero at the end of each switching period. For the forward converter output section shown in figure 14, the dc output voltage V is equal to the dc component of the diode D_3 voltage $v_{D3}(t)$, which is shown in figure 16.

Therefore, the output voltage in continuous conduction mode for $D \leq 0.5$ is given by,

$$V = \langle v_{D3} \rangle = (n_3/n_1)\, D\, V_g . \qquad 3.8$$

3.6 Switch and transformer utilization

Inclusion of isolation transformer into any converter topology leads to reduced switch utilization. The switch utilization U(D) and maximum switch utilization Max U(D) for normal buck converter and forward converter are compared in the table below.

Converter	U(D)	Max U(D)	Max U(D) occurs at D=
Buck	$\sqrt{D}$	1	1
Forward($n_1 = n_2$)	$\frac{1}{2}\sqrt{D}$	$\frac{1}{2\sqrt{2}} = 0.353$	0.5

The transformer isolated versions of the buck converter should be designed to operate at duty ratios as higher as possible. Even when the forward converter is operated at the upper limit of the duty ratio, the switch utilization is limited to $U \leq 0.353$. It implies that the switch stress is increased by a factor of approximately 2.8 as compared with the non-isolated buck converter at D=1. On the other hand, in isolated topologies the transformer turns ratio can be chosen to match the load voltage to the input voltage and better optimize the converter [6].

The transformer utilization in forward converter topology is good. The transformer magnetizing current cannot be negative and hence only half of the core B-H loop can be used in forward converter. This means the size of the transformer core in forward topology is twice as large as those of full or half-bridge converters. Nevertheless, in high-frequency converters, the flux swing is constrained by core loss rather than by the core material saturation flux density. Eventually, the forward converter transformer utilization can be as good as in the full- or half-bridge configurations. Since there is no center tapped windings in forward converter, utilization of the primary and secondary windings is better than in the full-bridge, half-bridge, and push-pull topologies. All of the available winding copper is used for energy transformation to the load during the first subinterval. There is no unnecessary flow of currents during second and third subintervals. The magnetizing current is small compared to the reflected load current in the primary, and has negligible effect on the transformer utilization. So, the transformer core and windings are effectively utilized in forward converters operating at higher frequencies.

3.7 Advantage and disadvantages

Some of the major advantages and disadvantages for the forward converter are as follows:

Advantages

- The peak currents in primary and secondary windings are lower (the inductance is higher, as no airgap is required) compared to flyback converter. Therefore, the copper losses are also lower. Though, this results in smaller temperature rise in the transformer, in many cases this improvement may not allow smaller cores to be used.
- The ripple in secondary is too low compared to flyback converter. The output inductor and flywheel diode maintain a reasonably constant current in the output load and reservoir capacitor.
- The energy stored in the inductor is available to the load. Hence, the reservoir capacitor can be made quite small so that its main function is to reduce output ripple voltages. Moreover, the ripple current rating for this capacitor will be much lower than that required for flyback converter.
- The power transistor has a lower peak current due to the high inductance of primary winding of the transformer.
- The output ripple voltages are lower.

Disadvantages:

- Cost is increased because of the extra output inductor and freewheeling diode.
- Under light load conditions, when the output inductor is reverted into discontinuous conduction mode, excessive output voltages will be produced, particularly on auxiliary outputs. This can be avoided by specifying minimum loads or by providing the ballast resistors.

3.8 Forward Converter Design

The step-by-step design procedure for the single ended forward converter with auxiliary winding reset is presented in the following section.

3.8.1 Step – I: Design specifications

The design begins with the definition of requirements. The output specifications are the biggest factor in determining the system parameters. Single output topology is always the simplest to design, while a requirement for several outputs with complex loading needs a careful consideration. Single output topology is considered in this thesis. The major design parameters are presented in the following discussion.

Input voltage range:

The actual input voltage range is greater than the specified range. Therefore, the designer has to take care that the converter works properly at a voltage that is lower than the specified minimum voltage. Similarly, the converter must be designed to operate at a voltage higher than the specified maximum input voltage.

Output voltage:

The output voltage can be maintained to ±4% over the range of line, load, and operating temperature ranges with normal feedback circuit. the proper choice of frequency compensation can control the transient response. The size of the output inductor and the choice of output capacitors greatly influence the output ripple and noise.

Output rectifiers:

The low cost discrete schottky diodes or high efficient synchronous rectifiers can be used for output rectification. Ultra fast PN junction diodes can be used at high operating frequencies.

Efficiency:

Higher efficiencies are always desired in any circuit design. Designing DC-DC converters has many engineering tradeoffs with efficiency against cost and complexity of the system. The designer has to make a good compromise between these tradeoffs to get the optimum design.

Temperature:

DC-DC converters may be needed to operate at higher temperatures compared to normal electronic devices. The characteristics of the passive elements might change at these higher operating temperatures to give unexpected and undesirable behavior. Particular attention must be paid while designing the circuit to achieve specified performance through out the temperature range.

Apart from these, output current and maximum output voltage ripple are also considered to be important design specifications.

3.8.2 Step – II: Transformer Design

Knowledge of system parameters allows the designer to determine the power transformer specifications. The transformer in forward converter is used for coupling and isolation, in which energy storage is undesirable. Detailed procedure for transformer design includes determining flux density excursions, selecting the core, designing the windings, and calculating losses and temperature rise.

Determining the flux density excursion

The first step in transformer design is to determine the flux density swing (ΔB) in normal steady state operating conditions. The flux density swing (ΔB) should be as large as possible to achieve fewer turns in the winding, increased power rating, and lower leakage inductance. In practice, ΔB is limited either by core saturation (B_{sat}), or core losses.

The transformer is driven symmetrically in many bridge, half-bridge, and full-wave center-tapped configurations so that the flux swing is symmetrical around zero on B-H curve. This means that these configurations have a maximum theoretical flux swing of 2 times of B_{sat}. But in the forward converter like in many single ended topologies, the core is driven in first quadrant of B-H characteristic from zero to B_{sat}. This limits the maximum flux swing to B_{sat} instead of 2 B_{sat}, so that the transformer in forward converter has only half the power handling capability.

In voltage fed circuits, the volt-seconds applied to the primary establish ΔB in accordance with Faraday's law. With normal steady state operating conditions, the primary volt-seconds will be constant, equal to $V_{in(min)}.t_{on(max)}$ or $V_{in(max)}.t_{on(min)}$. In simple duty cycle controlled converters, it is possible to have nearly twice the normal primary volt-seconds $V_{in(max)}.t_{on(max)}$, during start-up or after a large step increase in load current. Here it is assumed that $V_{in(max)}$ is nearly twice the $V_{in(min)}$. Therefore, the normal steady-state maximum flux density (B_{max}) cannot be greater than one-half B_{sat} or the core will saturate under transient conditions. On the other hand, in current-fed circuits ΔB is governed by the volt-seconds on the secondary windings which are clamped to the output voltages. Since the volt-seconds are totally independent of input voltage variations, current fed circuits can operate with B_{max} close to B_{sat} without the need for volt-seconds clamping unlike in voltage fed circuits.

Core selection

The second step in transformer design is to select a core which will house the windings. The core size is determined on the basis of transmitted power and it must support the required volt-seconds without saturating and with acceptable core losses and winding losses. The selection of core size can be accomplished with iterative methods but approximate solution can be obtained by relating the area product, AP, of the core window and magnetic cross section to the requirements of the application.

$$AP = A_W A_e = ((\frac{P_{in}.10^4}{K_t.K_u.K_p.450.\Delta B.2f})^{1.143} = (\frac{11.1P_{in}}{K.\Delta B.f})^{1.143} cm^4. \qquad 3.9$$

Where $p_{in} = p_o/\acute{\eta}$ = power output/efficiency

$K_t = I_{in}/I_p$, topology factor

$K_u = A_W{}^1/A_W$ = primary area factor

$K = K_t.K_u.K_c$

J = Current density (450 A/cm^2)

Equation 3.9 is based on the assumptions that the windings occupy 40% of the window area, the primary and secondary winding areas are proportioned for equal power density and the windings are operated at a current density appropriate for a 30^{o}C temperature rise with natural convection cooling.

For the forward converter the K factors are $K = 0.141$, $K_t = 0.71$, $K_u = 0.40$, and $K_p = 0.50$

Designing windings

The minimum number of primary turns, N_p, required to support the normal volt-seconds is given by the following equation.

$$t_{on(max)} = D_{max}/f = 0.5/f \text{ sec} \qquad 3.10$$

$$N_p > \frac{V_{in(\min)} t_{on(\max)}}{\Delta B A_e} .10^4 = \frac{5000.V_{in(\min)}}{\Delta B.A_e.f} \qquad 3.11$$

The transformation ratio is calculated at minimum V_{in} and maximum duty cycle. V_f is the rectifier forward drop. The factor 0.9 allows for inaccuracies and transistor storage times:

$$n = \frac{N_p}{N_s} = \frac{0.9[Vin(\min) - V_{CE(sat)}]D}{V_o + V_F} \qquad 3.12$$

Then the number of turns for secondary winding is calculated.

N_s = Integer (N_p/n).

The primary turns is recalculated.

$N_p = n\ N_s$.

The rms current in the primary winding, I_p, is:

$$I_{p(\max)} = I_{in(\max)} / K_t = \frac{P_{in(\max)}}{V_{in(\min)}.K_t} A. \qquad 3.13$$

The maximum current density for 30^0C rise with natural convection cooling for the core AP found:

$$J_{max} = 450\ AP^{-.125} \quad \Omega/cm^2 \qquad 3.14$$

The minimum primary wire area, A_{xp}, is:

$$A_{XP} = I_{p(max)}/J_{max}\ cm^2 \qquad 3.15$$

The maximum rms secondary current, I_s, occurs at 50% duty cycle:

$$I_{s\ (max)} = I_{o\ (max)}/1.414\ A. \qquad 3.16$$

Minimum secondary wire area, A_{xs}, is:

$$A_{XS} = I_{s(max)}/J_{max} \quad cm^2 \qquad 3.17$$

The primary and secondary AWG wire sizes can be looked in wire table. Wires larger than AWG 18 should be avoided to prevent eddy current losses and to make winding easier. When a larger wire is required, multiple paralleled turns of finer wire with equivalent total cross-sectional area can be used. For higher current secondaries, thin copper strip is often used.

3.8.3 Step – III: Output filter design

Output inductor

The inductor value is mainly influenced by the amount of output ripple current. Higher ripple current means smaller the output inductor both electrically and physically but it demands for more output capacitance with lower equivalent resistance (ESR). Moreover, higher ripple current translates to higher peak current for the power transistor for a given output power which means greater loss and obvious lower efficiency. A design parameter, output ripple factor, defined as the ratio of the peak-to-peak ripple current to the average current in the inductor is used in output inductor selection. Ripple factor can be determined as shown in figure 3.8.

The ripple factor,

$$K_{RF} = \Delta I / 2I_o. \tag{3.18}$$

Where Io is the maximum output current. For most practical design, the ripple factor is often set to 0.1~0.2.

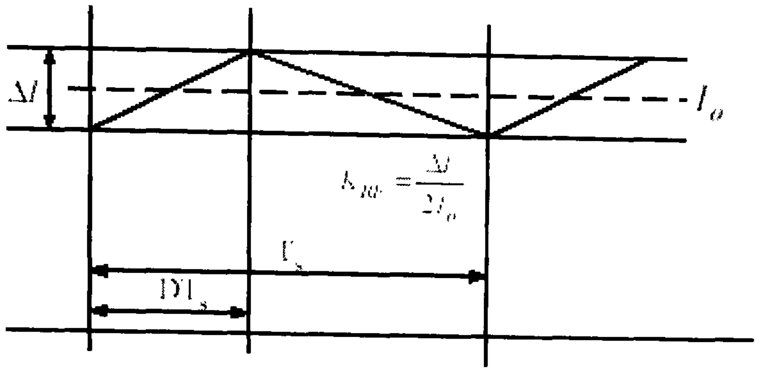


Figure 3.8 Output inductor current and Ripple factor[10]

The filter inductor calculation is based on the maximum switch OFF time:

$$D_{min} = D_{max}.\ V_{in(min)}/V_{in(max)}. \qquad 3.19$$

$$t_{off(max)} = (1 - D_{min})/f. \qquad 3.20$$

The inductance required to prevent discontinuous conduction mode of operation depends on the minimum load current

$$\Delta I_{max} = 2.\ K_{RF}.\ I_{o\,(min)}. \qquad 3.21$$

And output inductor value L is given as:

$$L = \frac{(V_o + V_F)t_{OFF(\max)}}{\Delta I_{\max}} \qquad 3.22$$

Output capacitor

The ripple current in the output inductor generates a voltage ripple on the output capacitors. Part of the ripple voltage comes from the integration of the current by the capacitance, and part comes from the voltage that appears across the capacitors ESR. The value of the output capacitance must be high enough and the ESR is low enough to give acceptable voltage ripple with the chosen output inductor.

The output capacitance can be calculated with the following equation:

$$C = \frac{1}{2} . \frac{\Delta I_{max}}{2} . \frac{1}{2f} . \frac{1}{V_o} \quad 3.23$$

And the maximum ESR of the capacitor is given by,

$$ESR = V_o / \Delta I_{max} \quad 3.24$$

3.8.4 Step – IV: Selection of the power semiconductors

Power switch selection

The first criterion in selecting a power transistor for switching is the peak current capability. Once the turns ratio for transformer and peak current in the output inductor are known, estimate the peak current through the power switch. The magnetization current can be neglected for this calculation. Switch should be selected such that it has a minimum current limit that is at least 10% greater than the maximum primary current.

The peak and rms currents for the MOSFET are obtained with the following equations.

$$I_{ds}^{peak} = I_{EDC}(1 + K_{RF}) \quad 3.25$$

$$I_{ds}^{rms} = I_{EDC}\sqrt{(3 + K_{RF}^2)\frac{D_{max}}{3}} \quad 3.26$$

Where,

$$I_{EDC} = \frac{P_{in}}{V_{DC}^{min} . D_{max}} . \quad 3.27$$

The second criterion for the selection is V_{DSS} rating of the MOSFET. In one transistor forward converters, the switch will see twice the maximum input voltage plus any spikes caused by winding leakage inductance and rectifier forward and reverse characteristics. Therefore, the minimum V_{DSS} rating for the MOSFET is

$$V_{DSS\,(min)} = 2\,(V_{in\,(max)} + V_{clamp\,(estimated)}). \qquad 3.28$$

Another major criterion is the heat generated by the device. The $R_{DS\,(ON)}$ and the drive circuit have the greatest influence on this. Multiplication of the $R_{DS\,(ON)}$ by the square of the RMS current in the primary gives a reasonable estimate of the power dissipation in the switch and hence the generated heat can be estimated.

Output rectifier

For the output rectifier diode, the maximum reverse voltage and peak output current are the deciding factors. These are calculated with the equations shown below and appropriate device is selected.

The maximum reverse voltage is:

$$V_{r\,(min)} = V_{in\,(max)}\,(N_s/N_p) \qquad 3.29$$

The peak output current is:

$$I_{o\,(peak)} = 2.8\, I_{o\,(max)} \qquad 3.30$$

Reset diode

The maximum voltage and rms current of the reset diode are given by:

$$V_{Dreset} = V_{DC}^{\max}\,(1 + \frac{N_r}{N_p}) \qquad 3.31$$

$$I_{Dreset}^{rms} = \frac{V_{DC}^{\min}.D_{\max}}{L_m f_s}\sqrt{\frac{D_{\max}}{3}} \qquad 3.32$$

CHAPTER – IV

SIMULATIONS AND TESTING

4.1 Simulations

The circuit used for the simulation of a forward converter is shown in Figure 4.1. The objective of this circuit is to produce a regulated 5V output voltage from 40V input voltage at 4A and 20W. This circuit can be modified to produce the desired output values. It uses a custom PWM generator block and a three winding transformer. The switch used here is an ideal switch. The values L, C are chosen in accordance with the analysis given in Chapter III. The time period of the saw tooth voltage is 20μs, thus the frequency of the converter is 50 KHz. The software used for these simulations is Pspice schematics student version (DesignLab Eval 8).

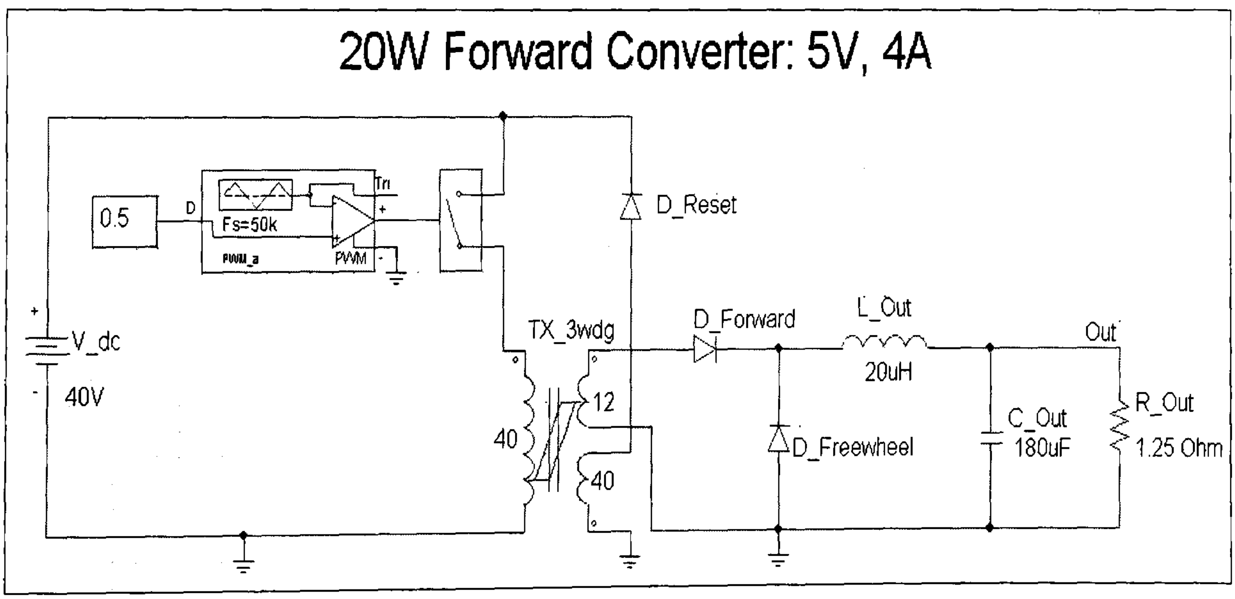


Figure 4.1 Schematic of a forward converter

The output waveforms of the forward converter are shown below.

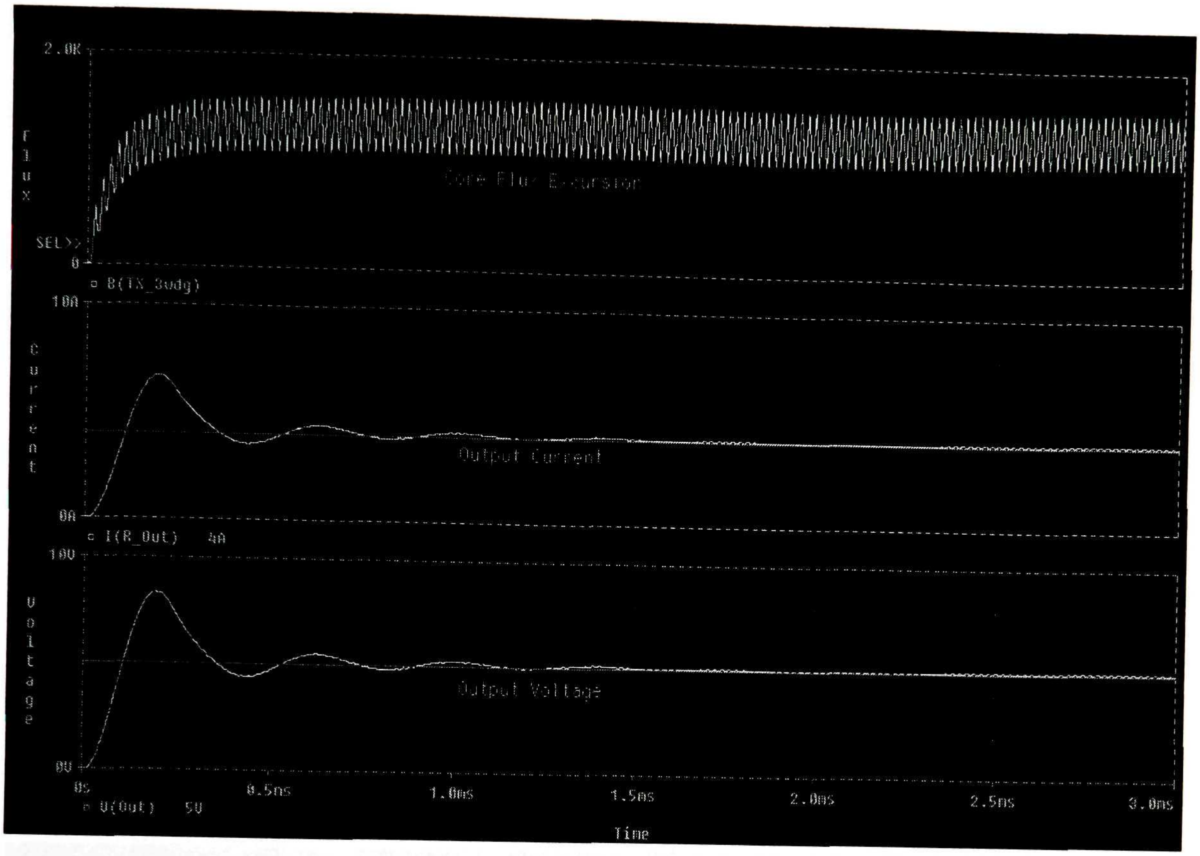


Figure 4.2 Transformer core flux, output current and voltage waveforms

The transformer core flux excursion shows that the transformer is operated in the first quadrant of the B-H curve. It has a dc offset value but it shows that it is driven in both the directions about the offset value so that it does not reach saturation. The output current and voltages have desired steady state values of 4A and 5V respectively. The initial transients can be reduced by incorporating proper RC circuit.

The voltages across the primary, secondary and reset winding are shown in figure 4.3. The primary winding is supplied with an AC voltage varying between +40 and -40V and the secondary winding produces a voltage varying between +12 and -12V according to the transformers turns ratio. The reset winding operates exactly as the primary winding but with opposite polarity. This confirms that the transformer is being reset after every cycle.

Figure 4.3 Transformer primary, secondary, and reset winding voltage waveforms

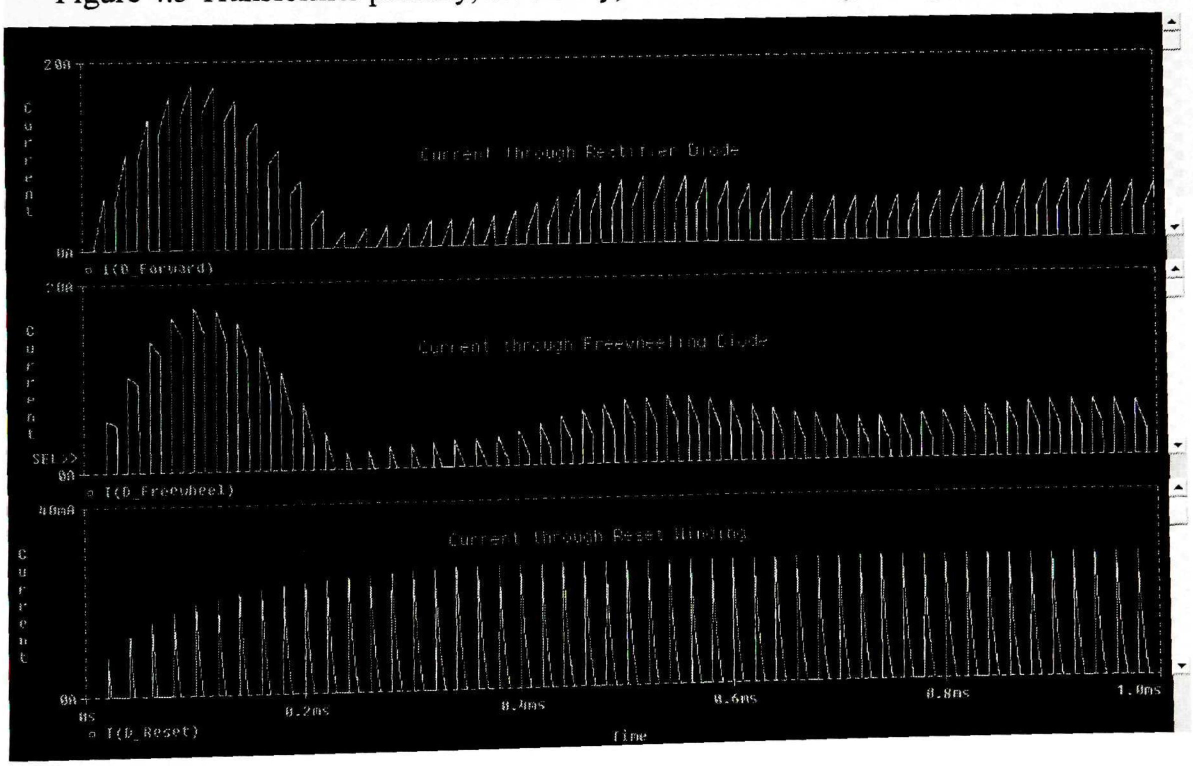


Figure 4.4 Rectifier diode, freewheeling diode and reset diode current waveforms

The currents through forward rectifier diode, freewheeling diode and reset diode are shown above in figure 4.4. The negative of the DC supply current, rectified voltage after the rectifying diode and PWM generation block waveforms are shown below in figure 4.5.

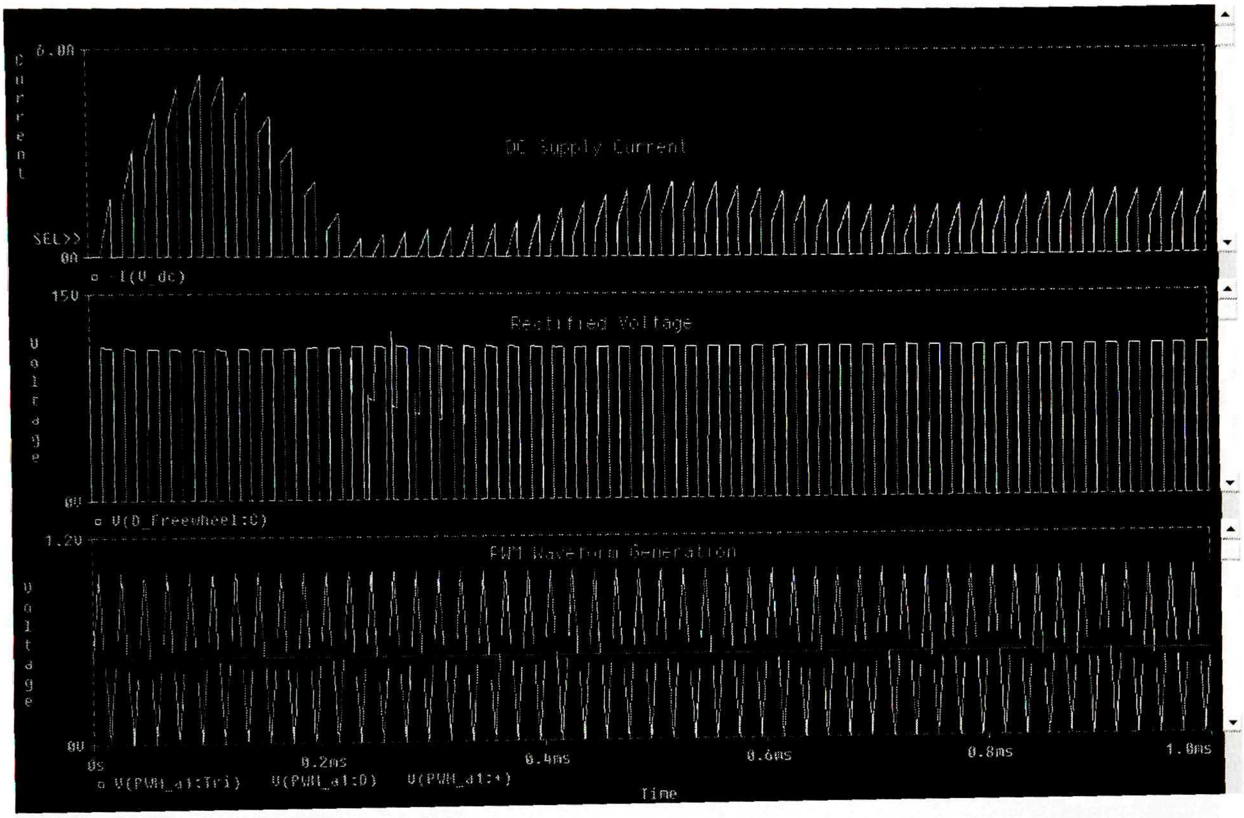


Figure 4.5 DC supply current, Rectified voltage and PWM waveforms

The schematic of a single phase bridge rectifier is shown in figure 4.5. This circuit can be used to supply the DC voltage for the forward converter simulated earlier. Here an attempt is made to produce a DC voltage from an input AC source which is further used in a DC to DC converter to produce the final desired DC output voltage. The rms and average power and the AC supply voltage and current waveforms are shown in figure 4.6. The difference between the apparent power and real power is the power factor of the circuit.

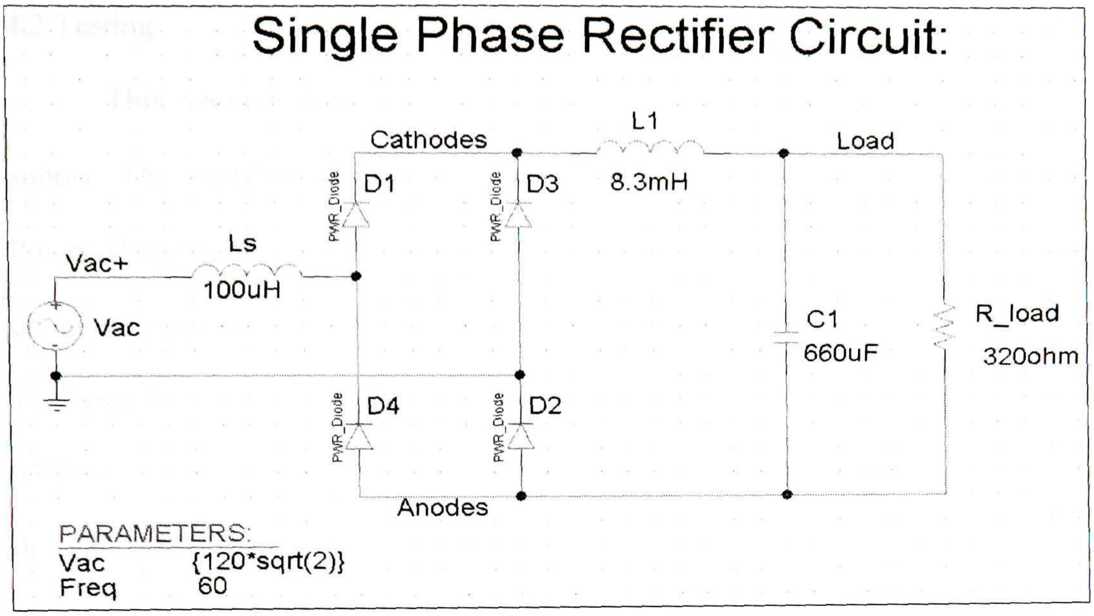


Figure 4.6 Schematic of a single phase bridge rectifier

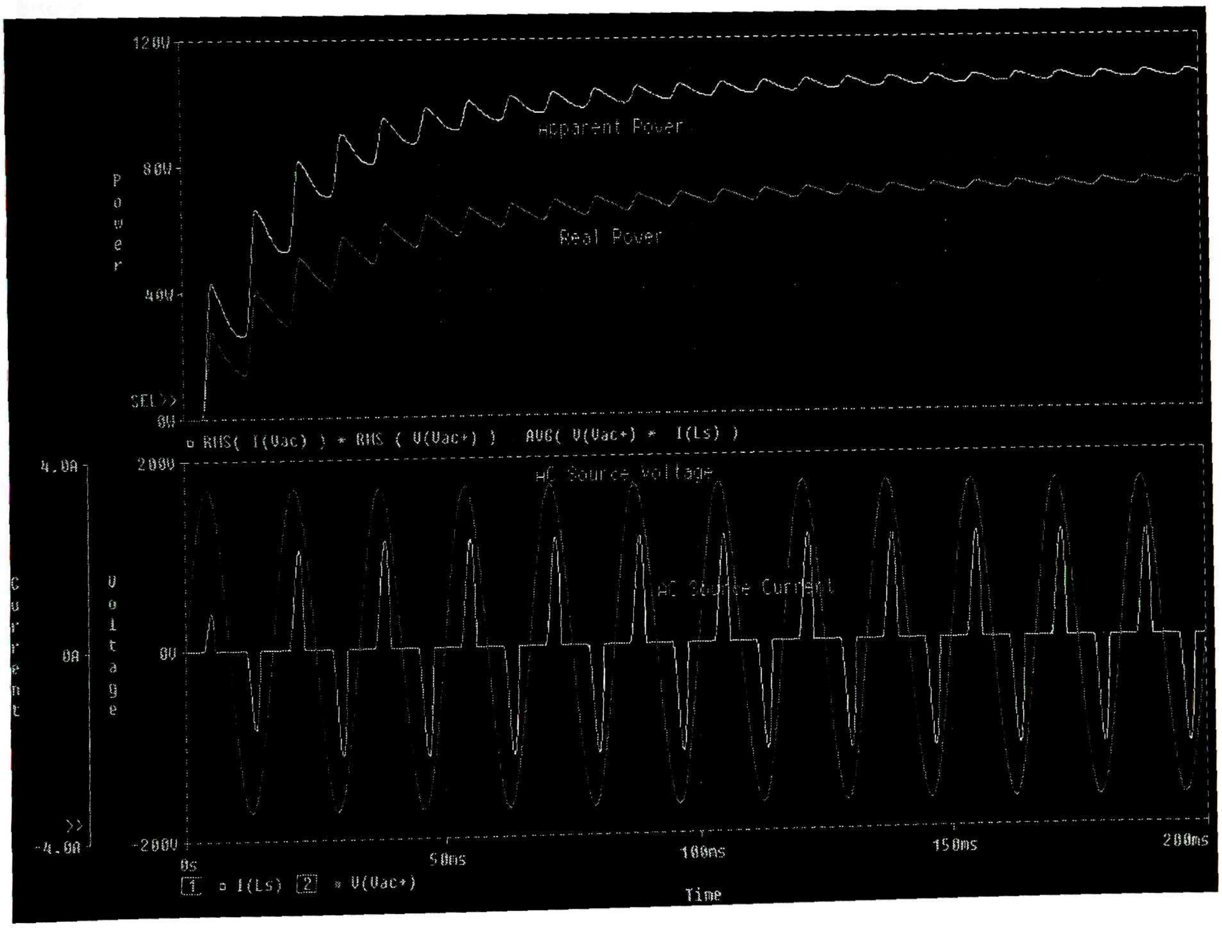


Figure 4.7 AC supply apparent and real power and voltage and current waveforms

4.2 Testing

This section describes testing of an engineering prototype computer power supply. The tested power supply is a Design Accelerator Kit (DAK-12) supplied by Power Integrations. DAK-12 consists of samples, a fully functional reference board and complete documentation. The power supply tested is an AC-DC power supply using TOPSwitch-GX in a single-ended forward converter. It addresses single input voltage 230V AC or doubled 115V AC input, but does not address universal input (85V to 265V) designs.

The specifications of the reference supply[12]:

Description	Symbol	Min	Typ	Max	Units	Comment
Input						
Voltage	V_{IN}	90	115	132	VAC	Doubler Input
Frequency	f_{LINE}	47	50/60	63	Hz	
Standby Input Power (115 VAC)			0.91	1	W	0.5 W Output From Standby
Blue Angel Input Power (240 VAC)			4.2	5	W	2.5 W output From Standby
Output						
Output Voltage 1	V_{OUT1}	3.17	3.30	3.43	V	±4%
Output Ripple Voltage 1	$V_{RIPPLE1}$			50	mV	20 MHz Bandwidth
Output Current 1	I_{OUT1}	0.5		12	A	
Output Voltage 2	V_{OUT2}	4.75	5.00	5.25	V	±5%
Output Ripple Voltage 2	$V_{RIPPLE2}$			50	mV	20 MHz Bandwidth
Output Current 2	I_{OUT2}	0.4		15	A	
Output Voltage 3	V_{OUT3}	11.16	12.0	12.84	V	±7%
Output Ripple Voltage 3	$V_{RIPPLE3}$			120	mV	20 MHz Bandwidth
Output Current 3	I_{OUT3}	0.05		3	A	5 A, 15 s Surge
Output Voltage 4 (standby)	V_{OUT4}	4.75	5.00	5.25	V	±5%
Output Ripple Voltage 4	$V_{RIPPLE4}$			50	mV	20 MHz Bandwidth
Output Current 4	I_{OUT4}	0		2.0	A	2.5 A, 15 s Surge
Total Output Power						
Continuous Output Power (main)	P_{O_main}		150	175	W	
Continuous Output Power (s/b)	$P_{O\_s/b}$		10	12.5	W	
Efficiency						
Main Converter	η_{main}	65	71		%	Measured at P_{O_main} =150 W
Environmental						
Conducted EMI						Meets CISPR22B / EN55022B
Safety						Designed to Meet IEC950, UL1950 Class I
Surge		4			kV	1.2/50 µs Surge, IEC 1000-4-5 12 Ω Series Impedance. Differential and Common Mode
Surge		4			kV	100 kHz Ring Wave, 500 A Short Circuit Current. Differential and Common Mode
Ambient Temperature	T_{AMB}	0		50	°C	Free Convection Sea Level

Schematics

Main Forward Converter primary side

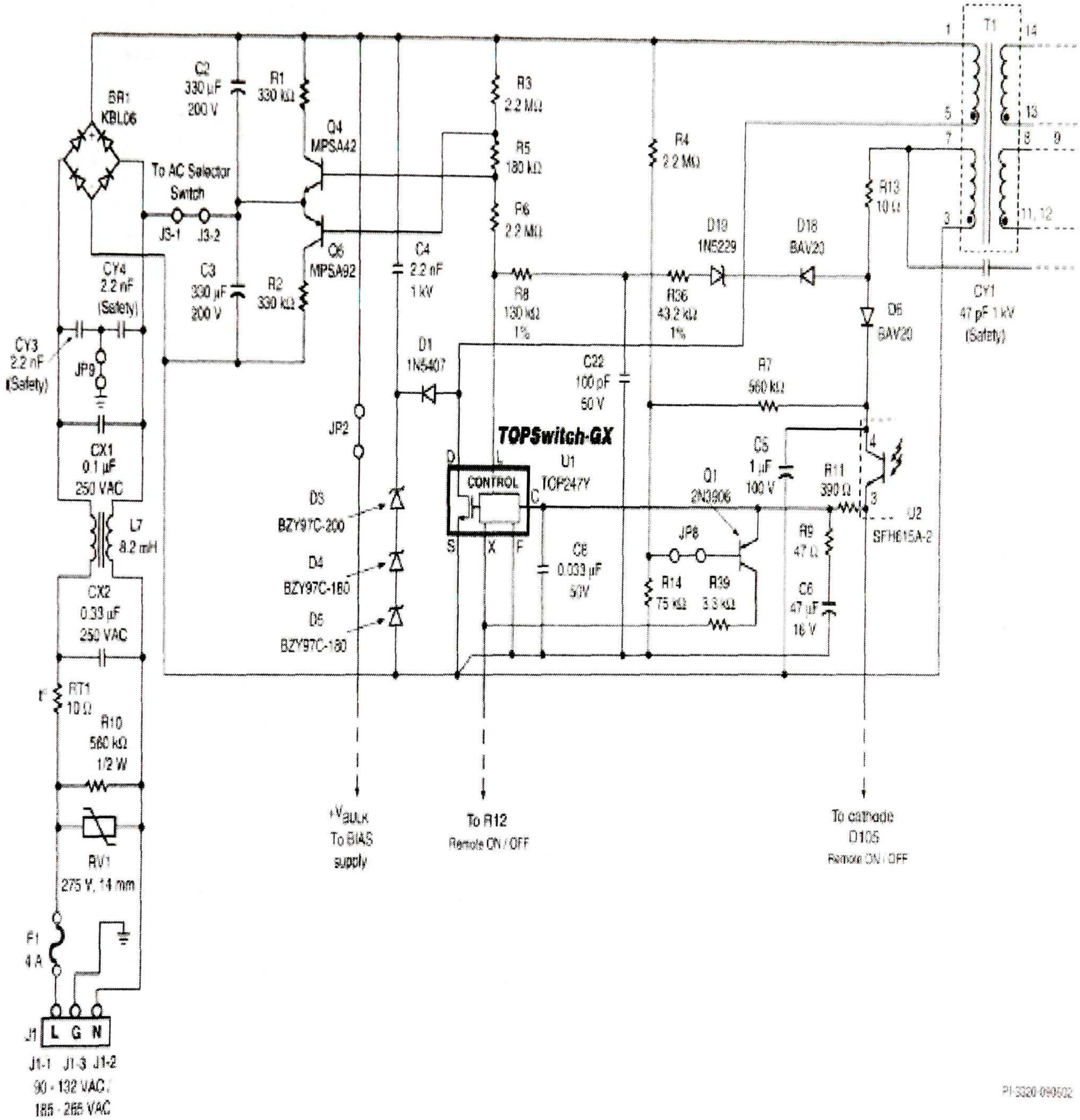


Figure 4.8 Main forward converter primary side[12]

Main Forward Converter secondary side

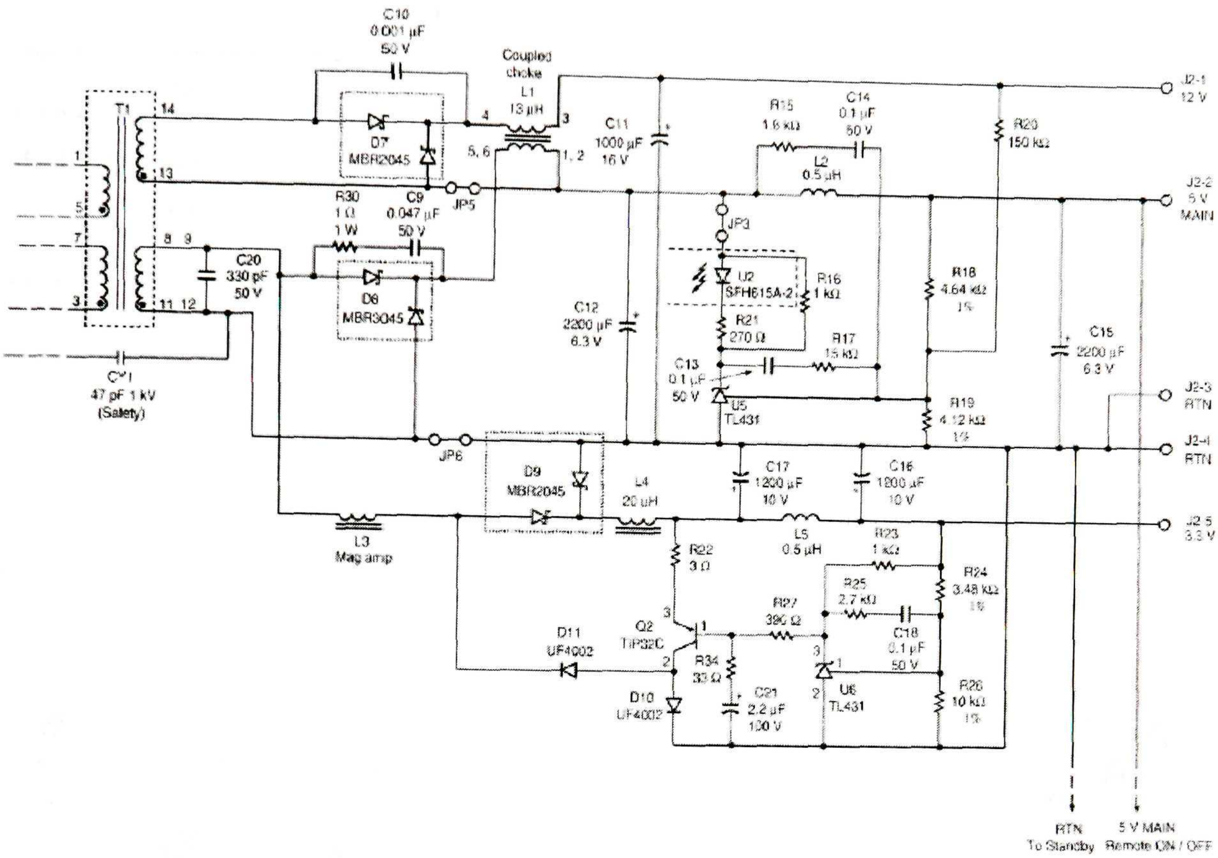


Figure 4.9 Main forward converter secondary side[12]

Standby Flyback Converter

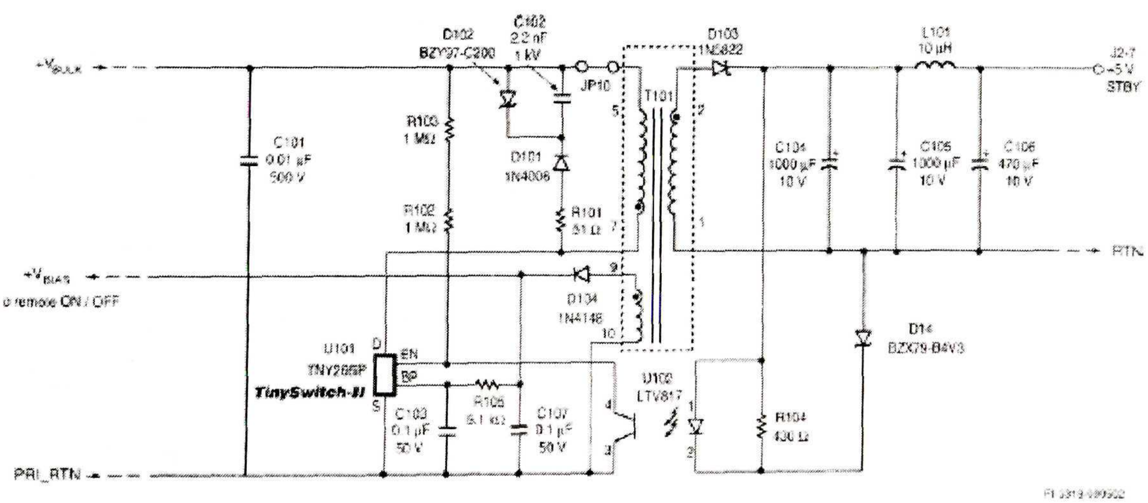


Figure 4.10 Standby Flyback converter[12]

Remote ON/OFF Interface

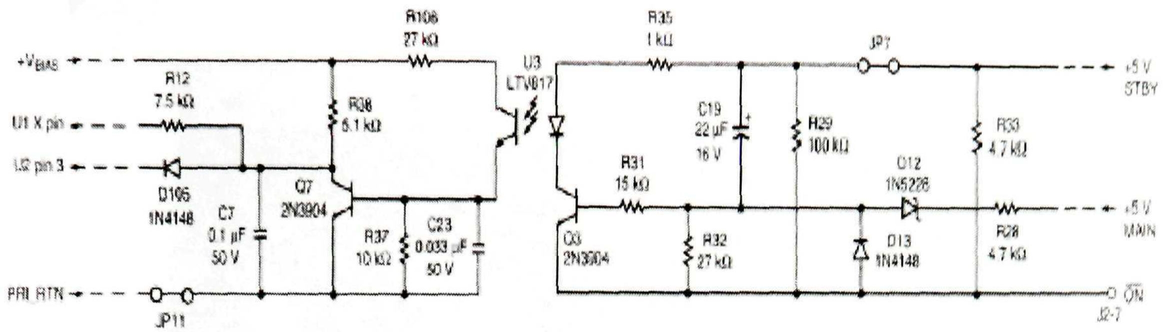


Figure 4.11 Remote ON/OFF interface[12]

PCB Layout

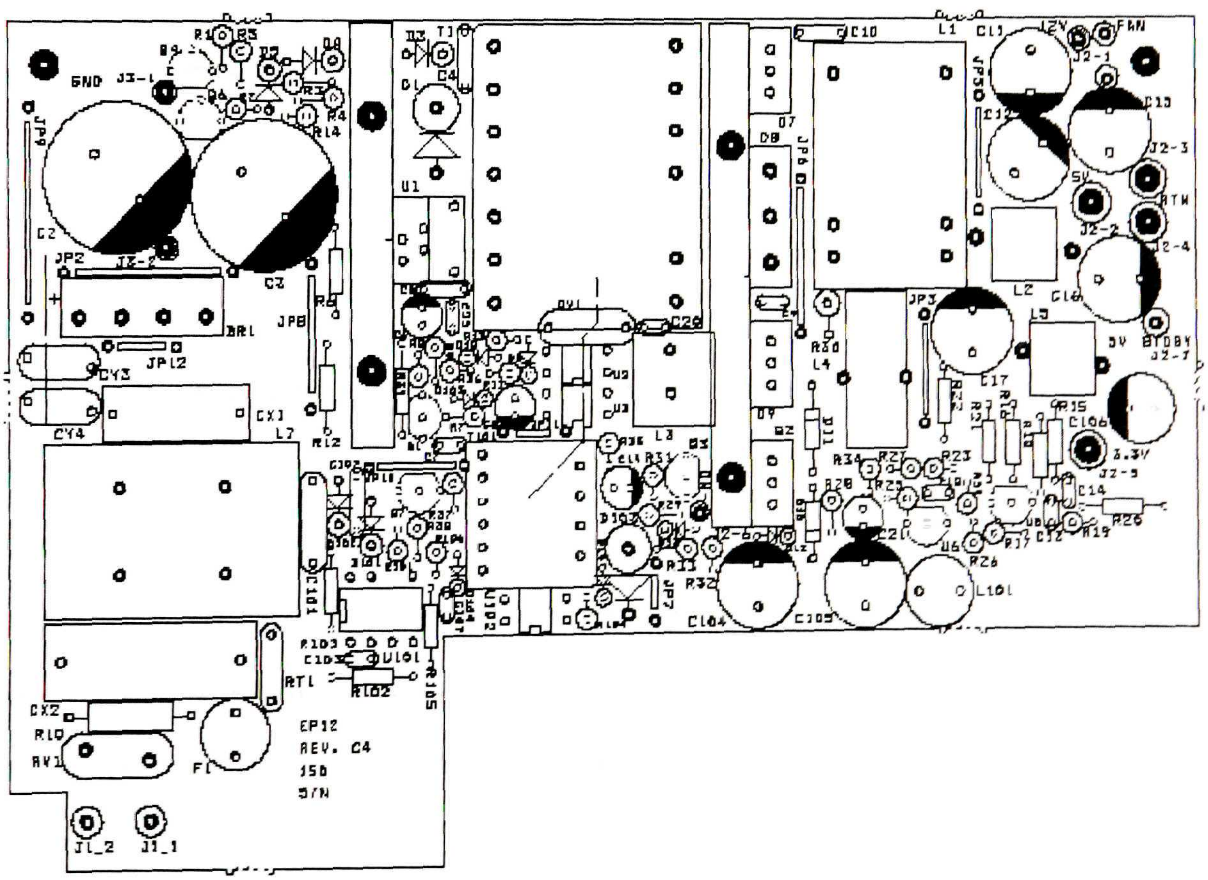


Figure 4.12 Assembly Diagram[12]

Power Supply Circuit Board

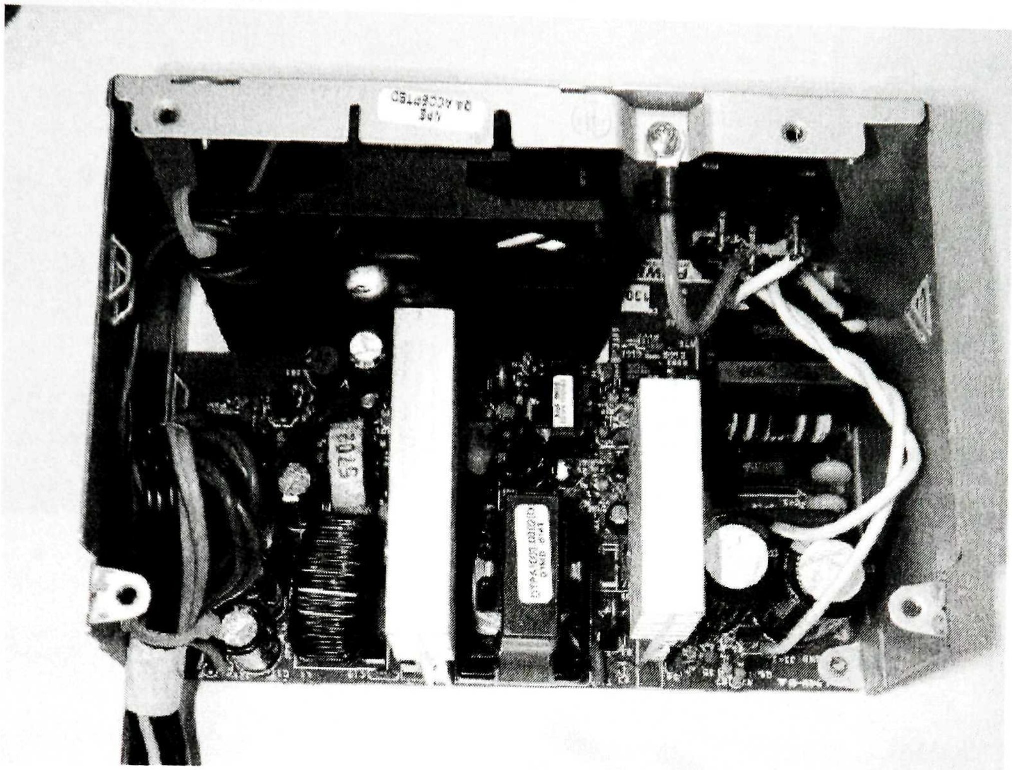


Figure 4.13 Populated circuit board

The Chroma 6310 series programmable DC electronic load is used for loading the power supply as shown in figure 4.14. The Chroma 6310 series Programmable DC Electronic Load is suitable for the test and evaluation of multi-output AC-DC power supply, DC-DC converter, and charger and power electronics components. The 6310 is actually a mainframe that controls two 63103 and two 63105 modules. These modules contain the electronics, which provide the load and also sense the voltage and current measurements. The difference between the two modules is the power, current and voltage ratings. The load can be operated in constant current, constant voltage, and constant resistance. In this testing, the loads were operated in constant current mode. The power supply is loaded with 3.3V/6A, 5V/6A, and 12V/1A throughout the testing.

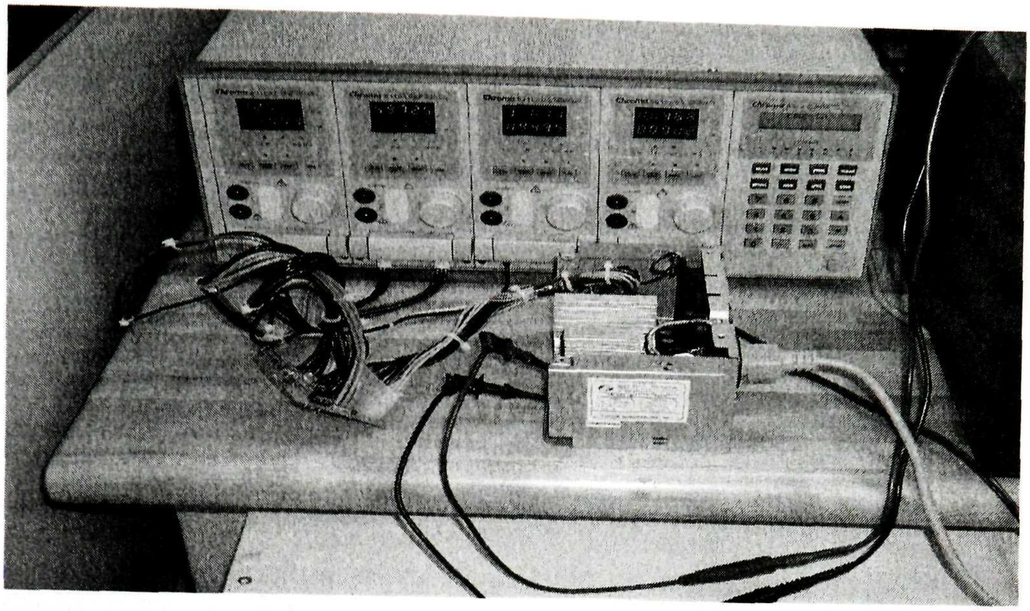


Figure 4.14 Reference supply with electronic load[14]

Figure 4.15 Tektronix P5200 Active Differential Probe[15]

The probe used for testing is a Tektronix P5200 active differential probe shown in figure 4.15. The P5200 can be used with any oscilloscope and provides safe measurements of floating circuits with oscilloscope grounded. The P5200 Active Differential Probe converts floating signals to low voltage ground referenced signals that can be displayed safely and easily on any ground referenced oscilloscope.

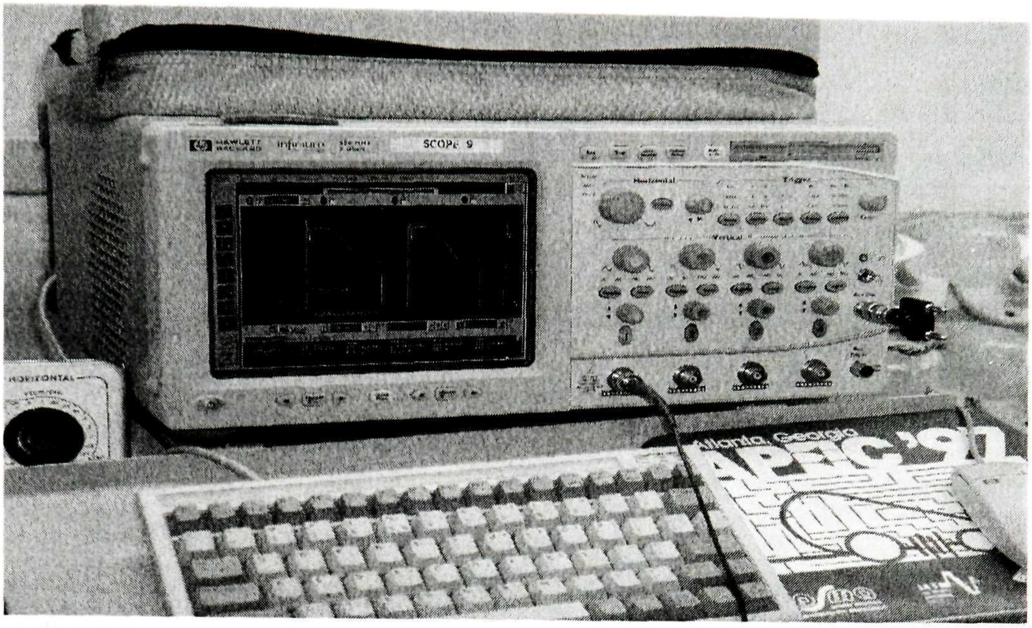


Figure 4.16 Infinium oscilloscope for measurements[16]

The HP/Infinium 54830A Oscilloscope is used for measurements as shown in figure 4.16. It offers the versatility and performance needed for general purpose design and troubleshooting. 4 channels, 500 MHz bandwidth, and a sample rate of 2 GSa/s on all four channels simultaneously ensures fast, accurate capture of waveforms. The 54830A's easy, familiar user interface makes it easy to take advantage of the performance inside. It has simple, analog-like front panel, dedicated vertical controls for each channel, windows-based graphical user interface, drag and drop measurements, built-in information system, advanced Triggering, and waveform math with FFTs and histograms.

Figure 4.17 is used to show the difference in size between a linear power supply and a switched mode power supply. The 10A linear supply is lot bulkier and heavier compared to switched mode supply. The large heat sink and the low frequency transformer add to size and weight to the linear supply. The high frequency operation of SMPS makes it size smaller and lighter.

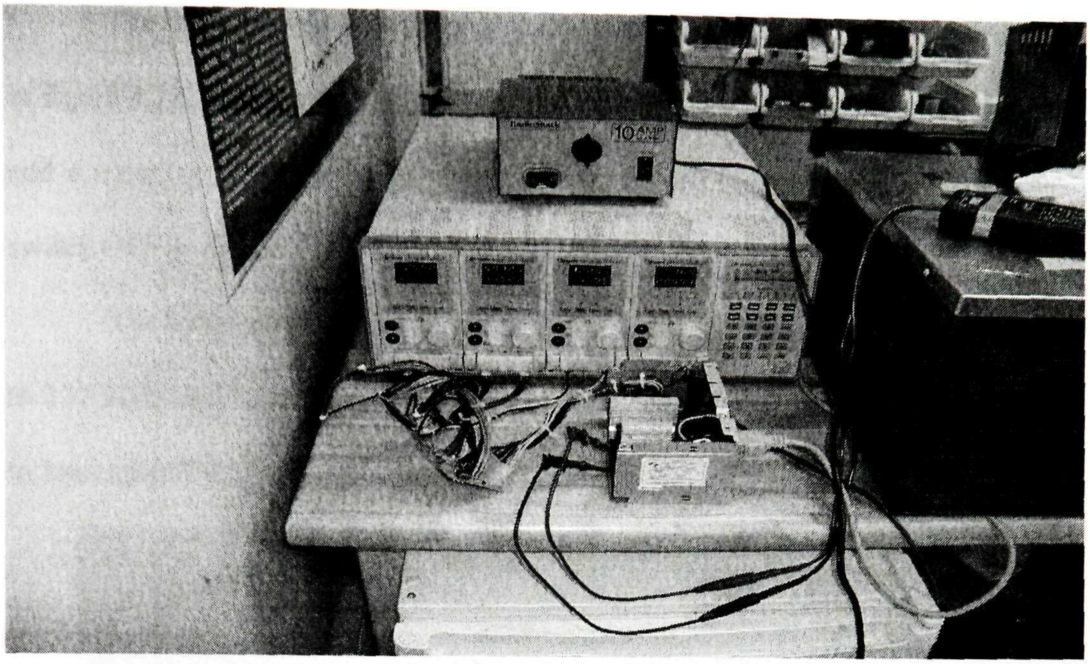


Figure 4.17 Linear and SMPS supplies are compared to demonstrate size difference

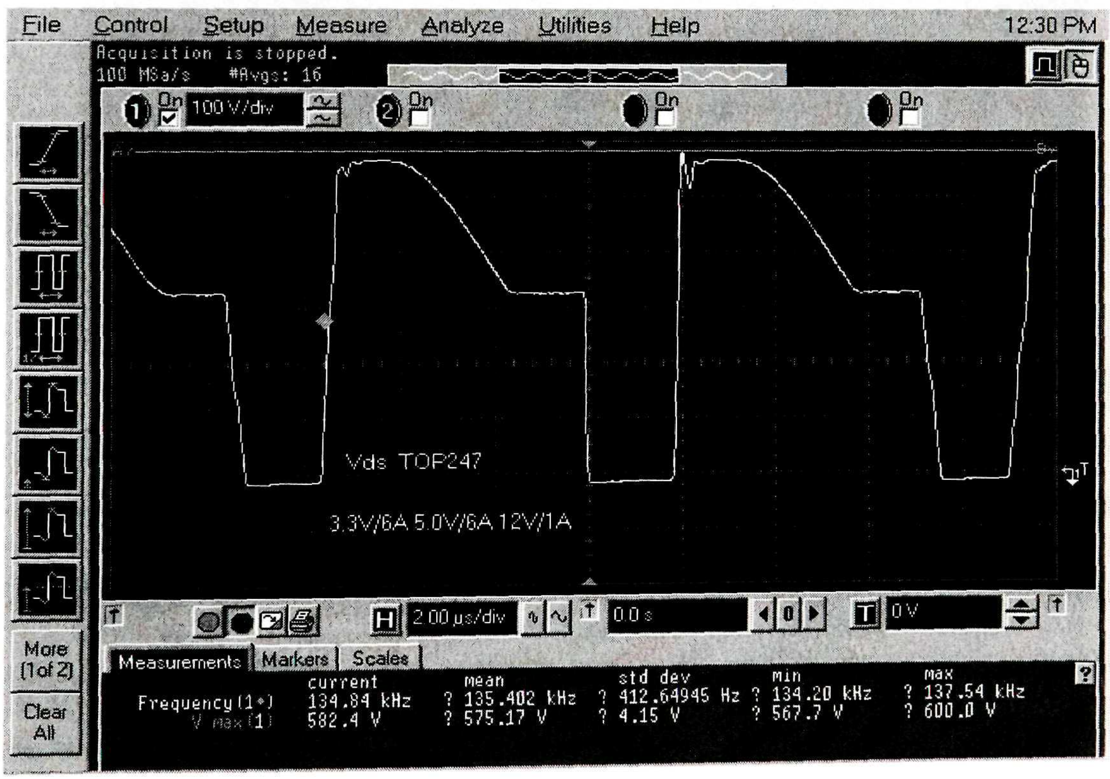


Figure 4.18 Drain to source voltage of the Topswitch

The drain to source voltage of the Topswitch in main forward converter is shown in figure 4.18. It has a switching frequency of 135.4 KHz and maximum voltage of 600V and a mean of 575V. It blocks almost twice the voltage of the DC bus voltage during switch OFF period.

The frequency jitter is shown in figure 4.19. The frequency is varied between 134 to 137 KHz and has a mean value of 135 KHz. Frequency jitter is induced on a purpose to lower the EMI.

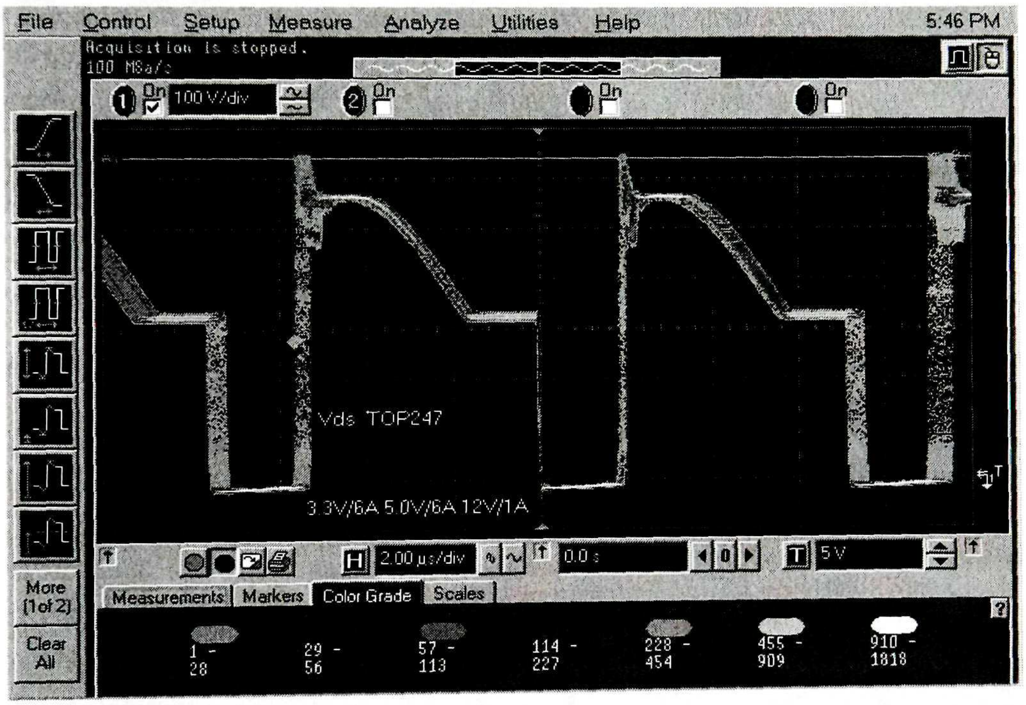


Figure 4.19 Frequency jittering for lower EMI

The voltages across the transformer terminals 1 and 5, 7 and 3, 1 and 3, 8 and 11, and 14 and 13 are shown in Figures 4.20, 4.21, 4.22, 4.23, and 4.24 respectively. The voltage across terminals 1 and 3 is DC bus voltage after the bridge rectifier and the rest of the voltages are AC voltages having zero average. These voltages are all according to the design and required specifications.

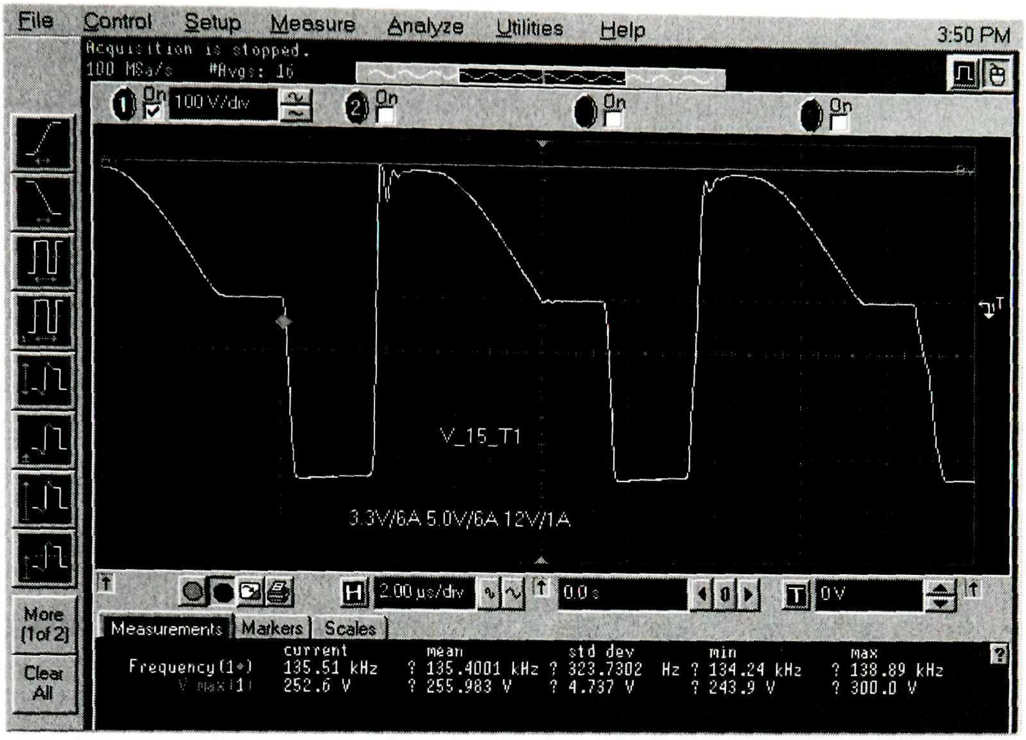


Figure 4.20 Voltage across transformer terminals 1 and 5

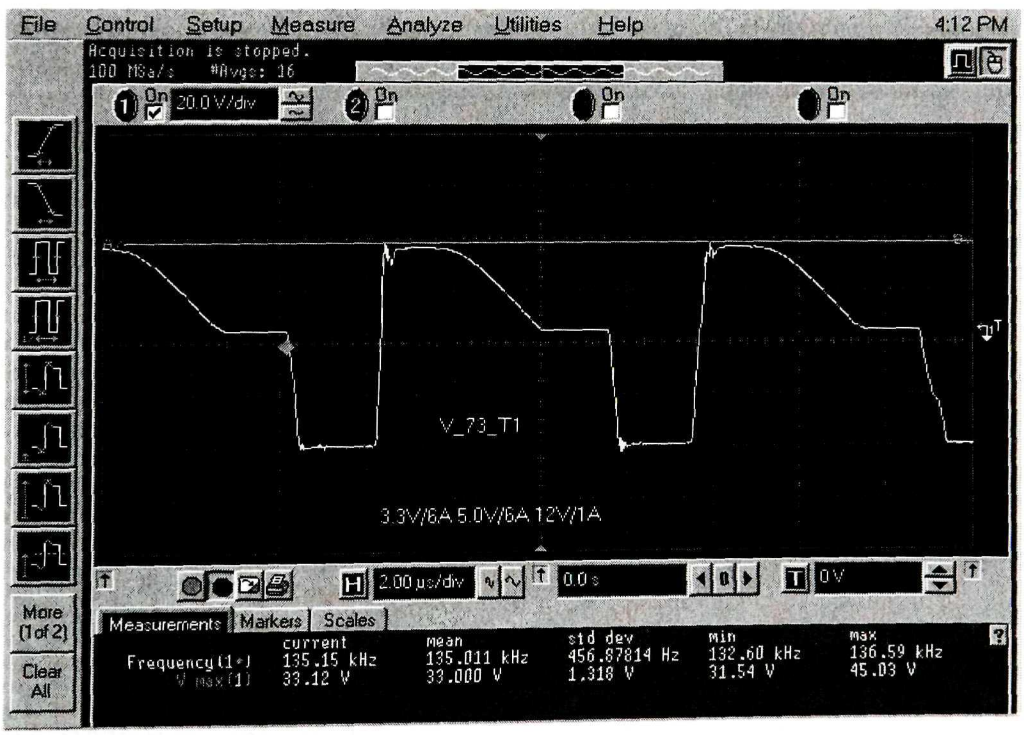


Figure 4.21 Voltage across transformer terminals 7 and 3

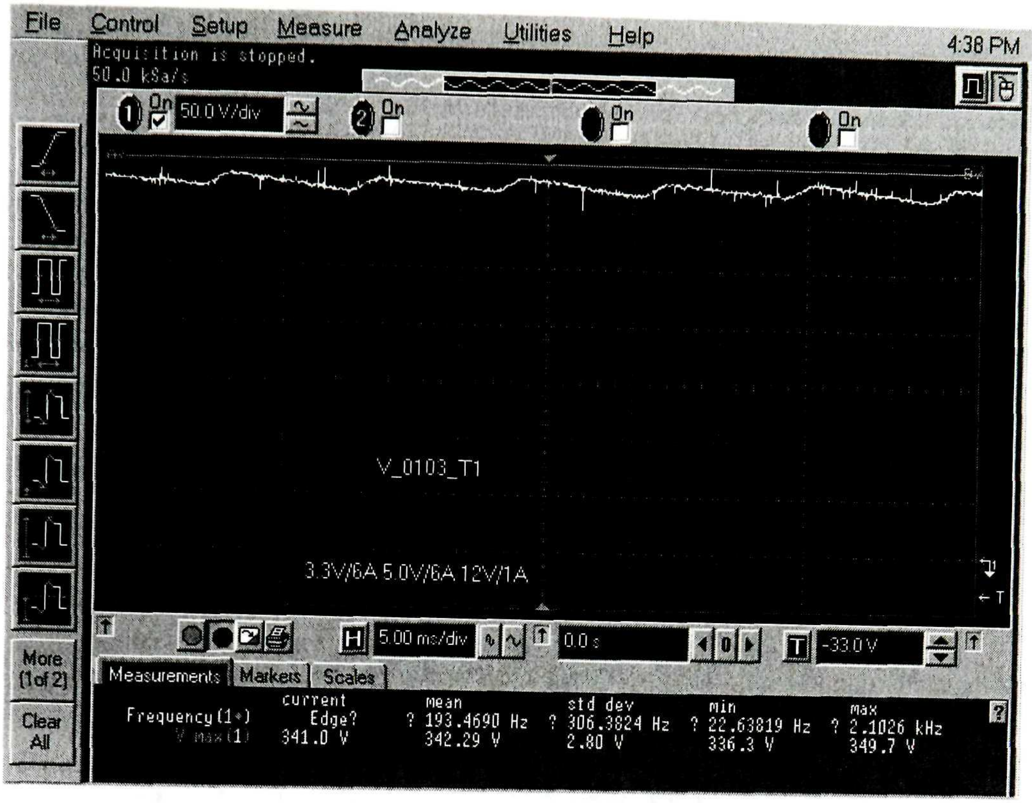


Figure 4.22 DC bus voltage after the bridge rectifier

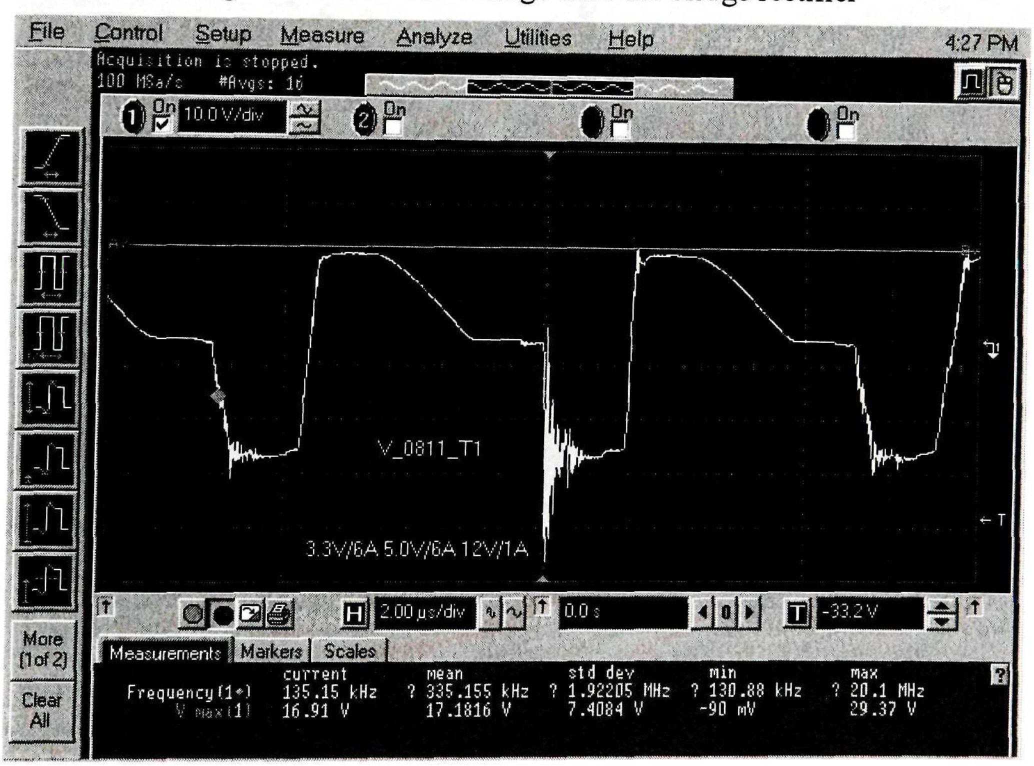


Figure 4.23 Voltage across transformer terminals 8 and 11

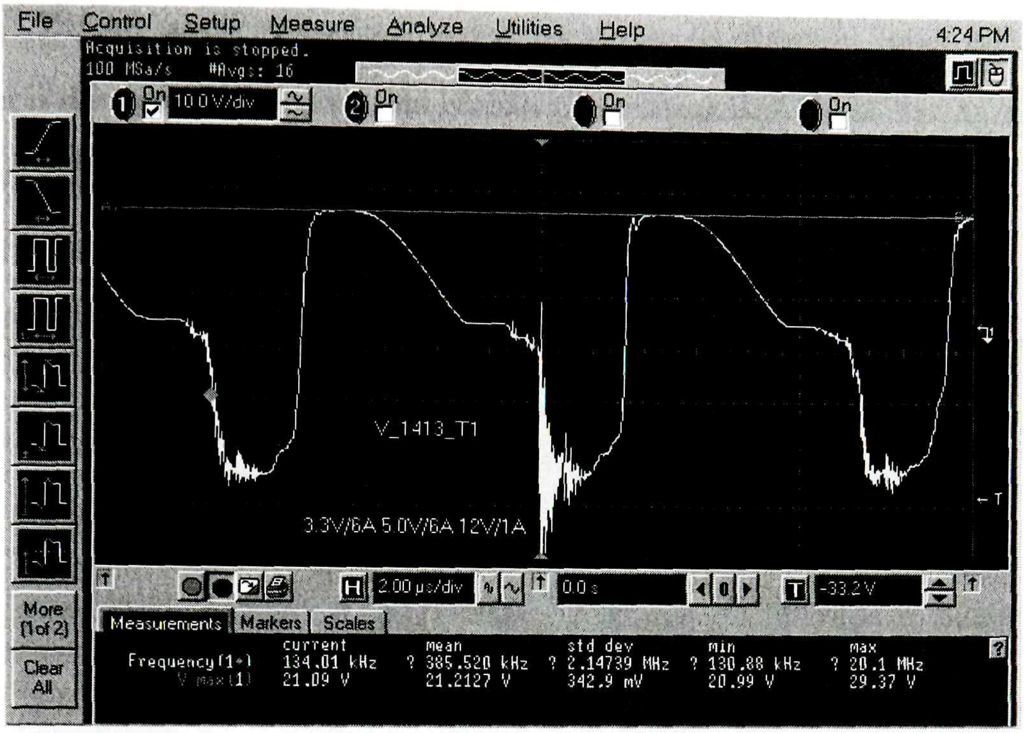


Figure 4.24 Voltage across transformer terminals 14 and 13

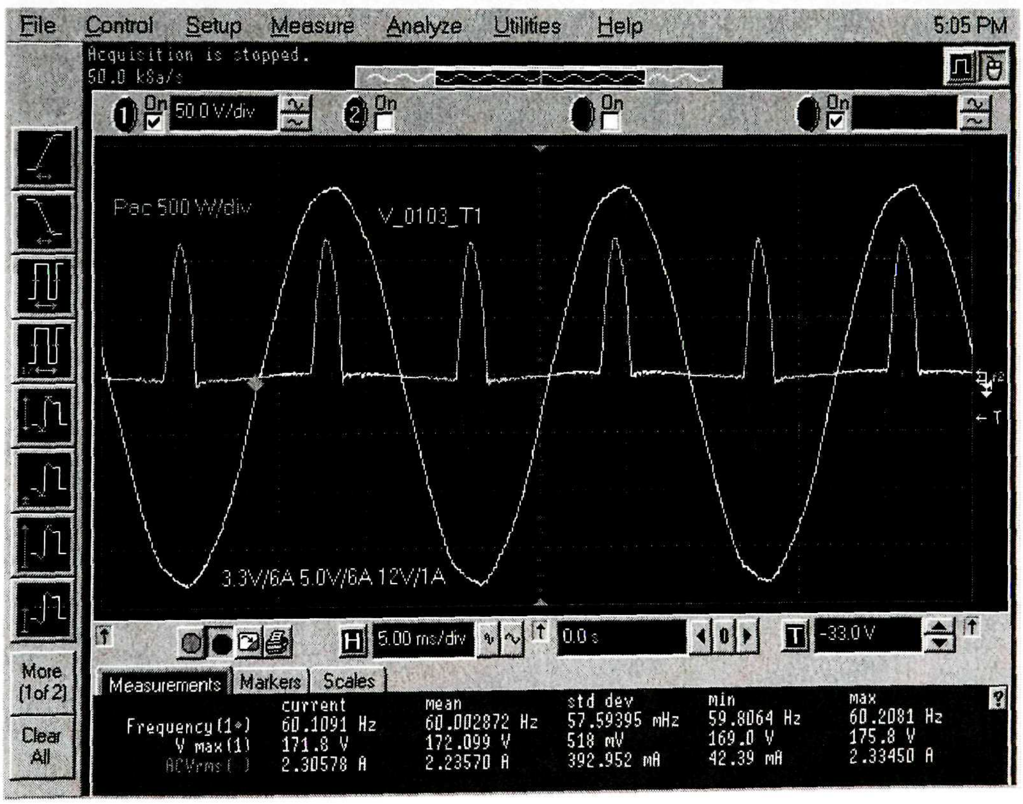


Figure 4.25 AC supply voltage, current and power waveforms

The AC supply voltage and current waveforms are shown in Figure 4.25. The power is shown as the function of the product of the voltage and current. The AC supply current and its FFT are shown in figure 4.26.

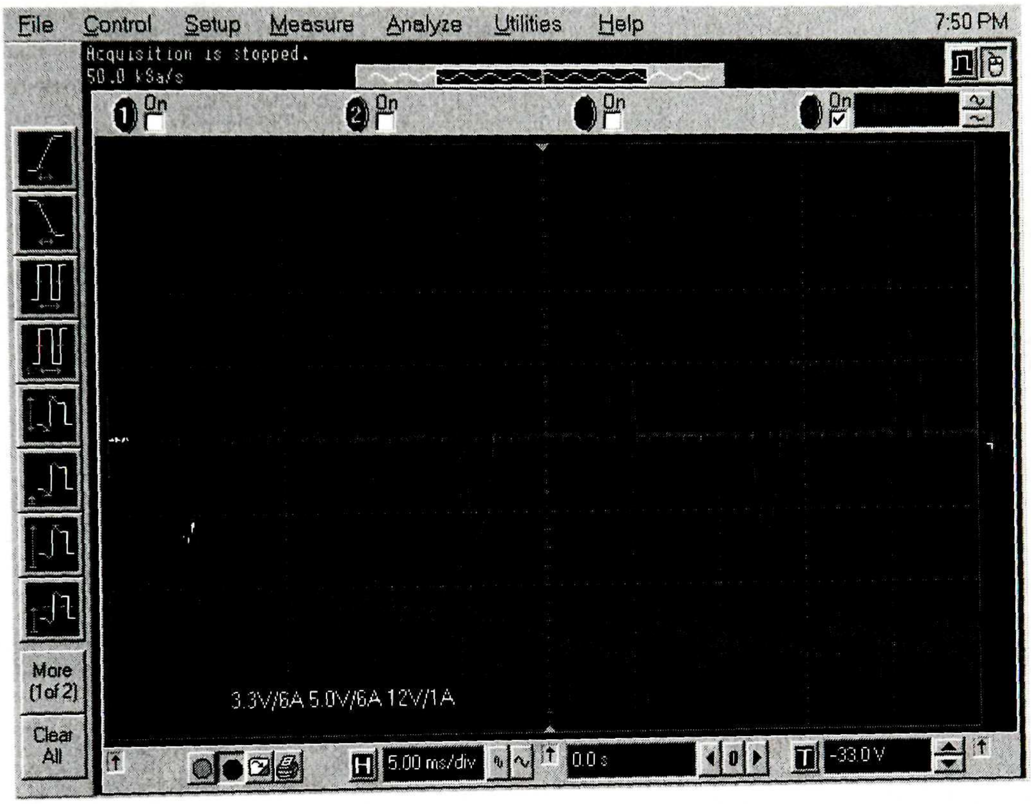
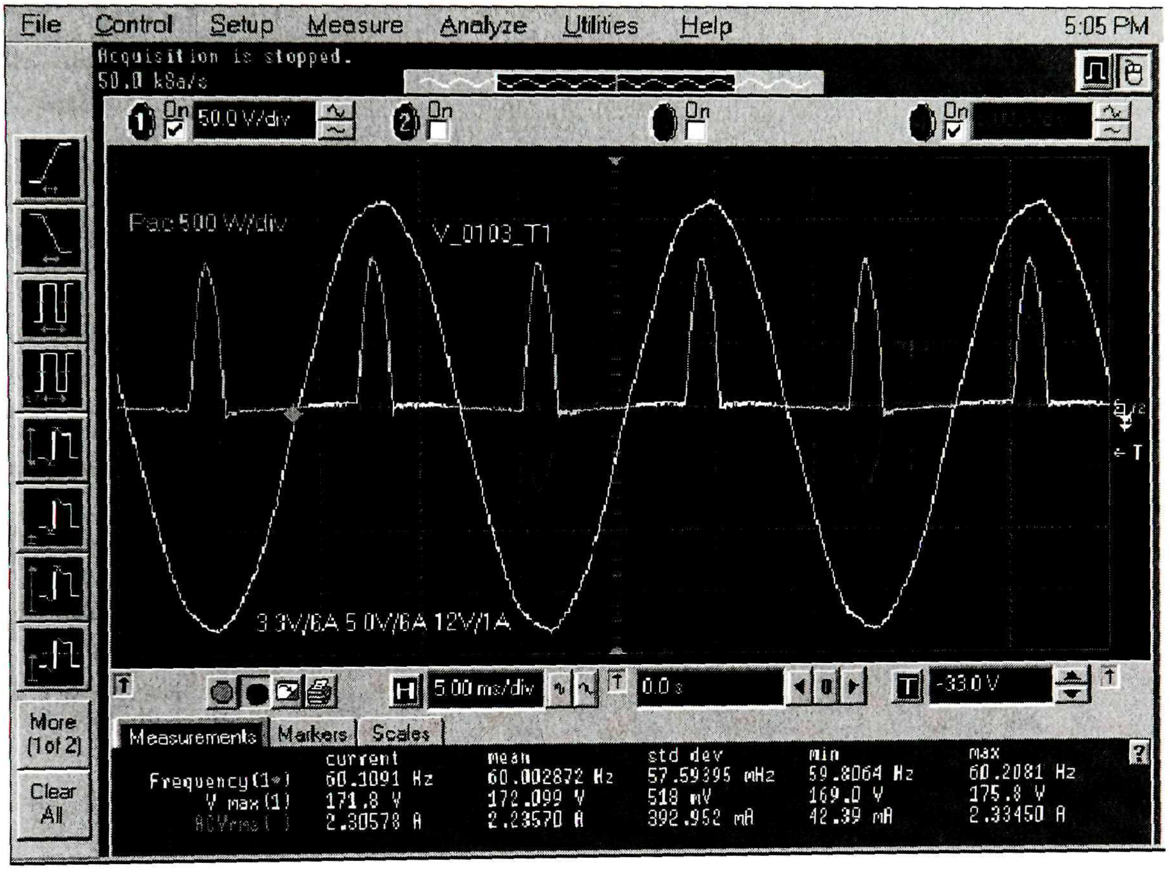


Figure 4.26 AC supply current waveform and its FFT

4.3 Calculations

The data from the test results is extracted into Mathcad and calculations are done which are shown in the following section.

Data from Infinium Scope: Forward Converter PC Power Supply: Jul-30-2004: $ms \equiv 10^{-3} \cdot sec \quad \mu s \equiv 10^{-6} \cdot sec$



Read Data from .prn File: $Pwr_{AC} := READPRN("AC\_Pwr.prn")$

Moving AVG on vector x, window n, index i:

$$AVG(x,n,i) := \begin{cases} x_0 & \text{if } i = 0 \\ \dfrac{1}{i+1} \cdot \displaystyle\sum_{k=i}^{0} x_k & \text{if } i < n \\ \dfrac{1}{n} \cdot \displaystyle\sum_{k=i}^{i-n+1} x_k & \text{otherwise} \end{cases}$$

Separate Data Array into Vectors:

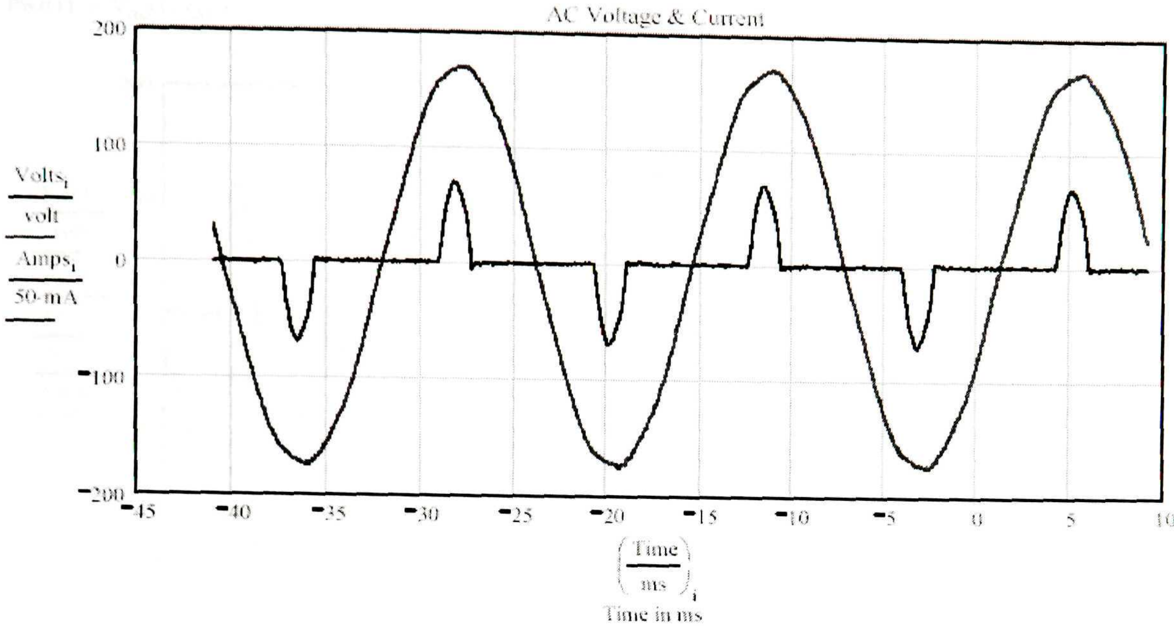
$Time := Pwr_{AC}^{\langle 0 \rangle} \cdot sec$ $\quad n := last(Time) \quad n = 2504$

$Volts := Pwr_{AC}^{\langle 1 \rangle} \cdot volt$ $\quad m := last(Volts) \quad m = 2504$

$Amps := \dfrac{Pwr_{AC}^{\langle 2 \rangle}}{2} \cdot amp$ $\quad o := last(Amps) \quad o = 2504$

$i := 0 .. n$

Figure 4.27 Data from scope



Calculating Sampling Time Step:

$TimeStep := Time_1 - Time_0$ $\qquad TimeStep = 20\,\mu s$

Calculating total Time:

$Duration := Time_{last(Time)} - Time_0$ $\qquad Duration = 50.08\,ms$

Defining Voltage & Current as functions of time:

$$V_{ac}(t) := Volts_{\left(round\left(\frac{t - Time_0}{TimeStep}\right)\right)}$$

$$I_{ac}(t) := Amps_{\left(round\left(\frac{t - Time_0}{TimeStep}\right)\right)}$$

$T_{start} := -40 \cdot ms$ $\qquad T_{stop} := 10 \cdot ms$

$t := T_{start}, T_{start} + TimeStep .. T_{stop}$

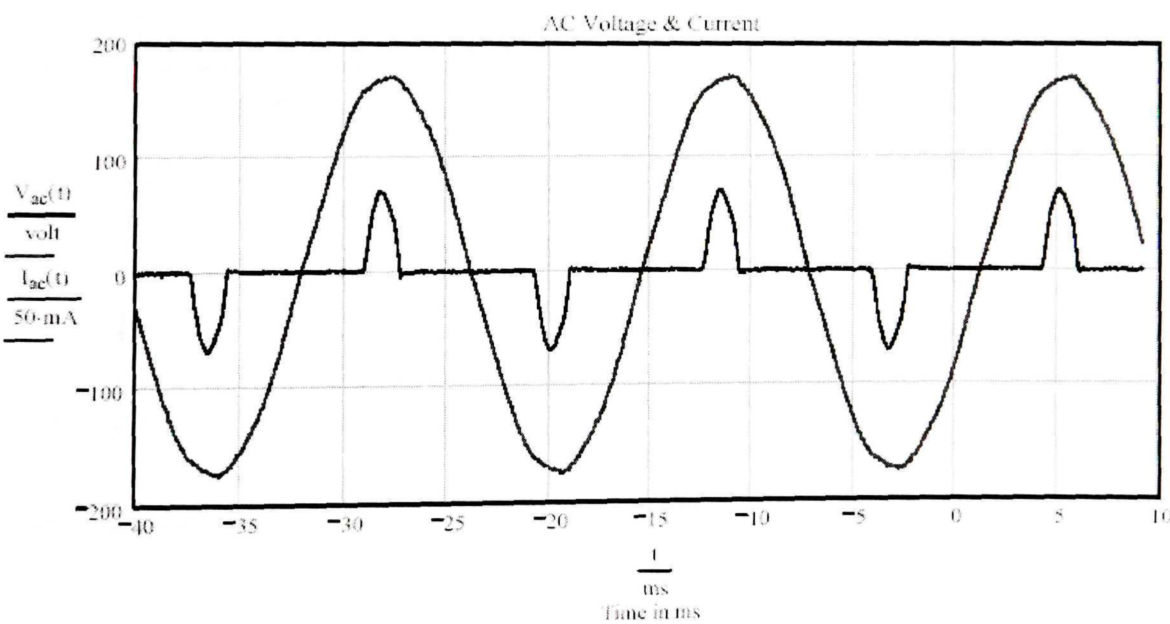
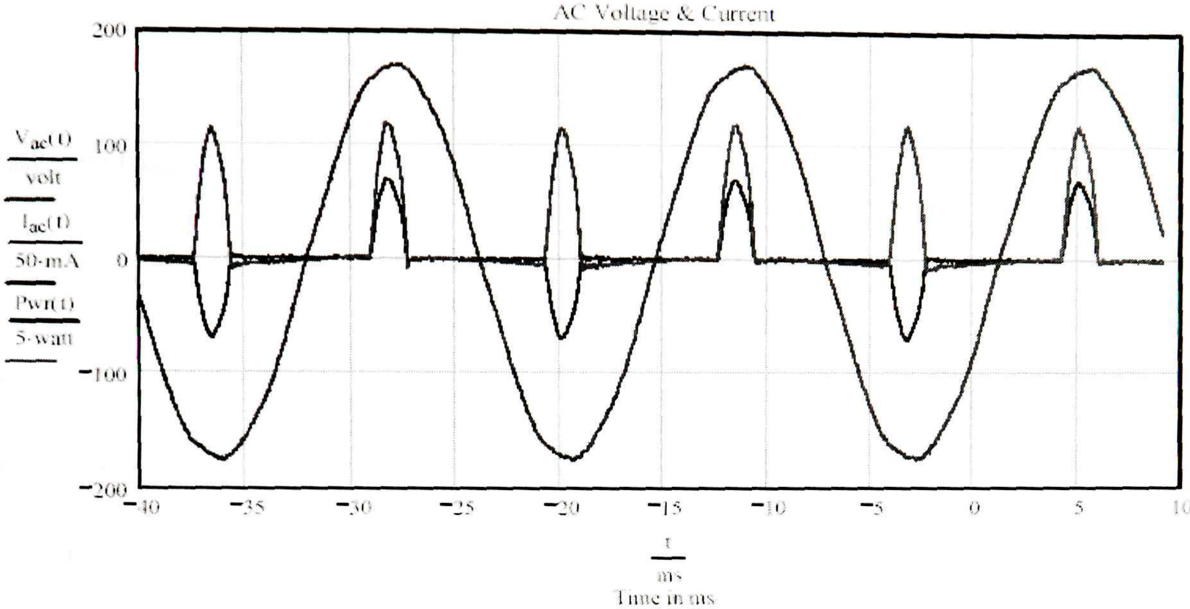


Figure 4.28 Voltage and Current waveforms

$Pwr(t) := V_{ac}(t) \cdot I_{ac}(t)$



$VA \equiv volt \cdot amp$

$$P_{avg} := \frac{\int_{-32.5 \cdot ms}^{2 \cdot ms} Pwr(t)\,dt}{2 \cdot \left(\frac{1}{60 \cdot Hz}\right)} \qquad P_{avg} = 81.1\ watt$$

$$P_{out} := (5 \cdot volt \cdot 6 \cdot amp) + (3.25 \cdot volt \cdot 6 \cdot amp) + (12 \cdot volt \cdot 1 \cdot amp) \qquad P_{out} = 61.5\ watt$$

$$V_{rms} := \sqrt{\frac{\int_{-32.5 \cdot ms}^{2 \cdot ms} V_{ac}(t)^2\,dt}{\left[2 \cdot \left(\frac{1}{60 \cdot Hz}\right)\right]}} \qquad V_{rms} = 122.8\ volt$$

$$I_{rms} := \sqrt{\frac{\int_{-32.5 \cdot ms}^{2 \cdot ms} I_{ac}(t)^2\,dt}{\left[2 \cdot \left(\frac{1}{60 \cdot Hz}\right)\right]}} \qquad I_{rms} = 1.2\ amp$$

$$V_{rms} \cdot I_{rms} = 142.2\ VA$$

$$Eff := \frac{P_{out}}{P_{avg}} \qquad Eff = 75.9\%$$

$$PF := \frac{P_{avg}}{V_{rms} \cdot I_{rms}} \qquad PF = 57\%$$

Figure 4.29 Power, Efficiency, and PF calculations

$$f_{Sample} := \frac{1}{TimeStep} \qquad f_{Sample} = 50\,kHz$$

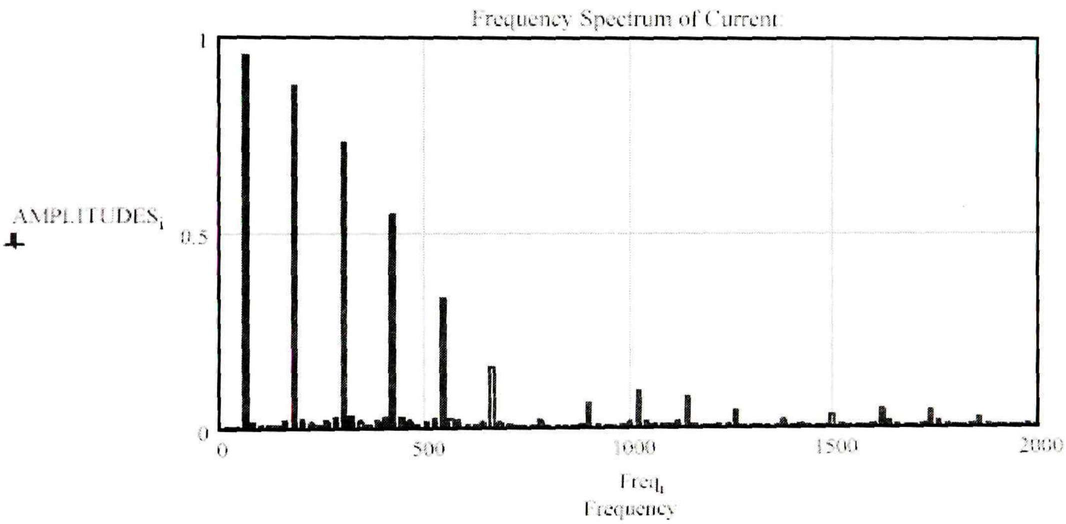
$$f_{max} := 0.5 \cdot f_{Sample} \qquad f_{max} = 25\,kHz \qquad f_{res} := \frac{1}{Duration} \qquad f_{res} = 19.968\,Hz$$

$$i := 0..n-1 \qquad 3 \cdot f_{res} = 59.904\,Hz$$

$$SPECTRUM := CFFT\left(\frac{Amps}{amp}\right) \qquad AMPLITUDES_0 := \left|SPECTRUM_0\right| \qquad last(SPECTRUM) = 2504$$

$$Freq_0 := \frac{0}{sec} \qquad i := 1..\frac{n}{2} \qquad AMPLITUDES_i := 2 \cdot \left|SPECTRUM_i\right| \qquad Freq_i := \frac{i}{Duration}$$

$$AMPLITUDES_0 = 0.085 \qquad AMPLITUDES_3 = 0.95$$



$$THD := \frac{\sqrt{\sum_{i=0}^{\frac{last(SPECTRUM)}{2}} \left(AMPLITUDES_i\right)^2 - \left(AMPLITUDES_3\right)^2}}{AMPLITUDES_3} \qquad THD = 139.844\,\%$$

Figure 4.30 Harmonic distortion calculations

CHAPTER – V

CONCLUSIONS AND FUTURE WORK

The motivation for this thesis is to study the complete theoretical, analytical and practical considerations of a forward converter. Forward converter is the most popular choice for computer power supplies. The forward converter has dominated the commercial power supplies market for the past 50 years but it is not covered extensively in any academic text books. This thesis has discussed the need for power supplies and why and how switched mode power supplies are better choice compared to linear power supplies for electronic devices. Basics of a switched mode power supply and various DC to DC converter topologies were discussed in detail. Practical design considerations and design equations for a switched mode power supply based on the forward converter were presented. The behavior of the characteristics of a single ended forward converter with reset winding was demonstrated by simulations. A reference power supply manufactured by power integrations was tested in the lab. The simulation results and the test results were demonstrated to be well in agreement with the theoretically expected results. But the reference power supply has power factor of 57% and total harmonic distortion of 139.8%. These values are not acceptable even though they don't affect the efficiency and other performance parameters of the power supply. Other measures like power factor correction/improvement and wave shaping circuits must be employed to improve power factor and THD. The objective of the thesis was fulfilled by extensive study of the forward converter design, basic forward converter simulations, and testing of a reference power supply.

The forward converter can be further simulated by including control, protective and supervisory circuits. At this stage, all the information obtained from theoretical design and simulations can be used to build a circuit that operates at nominal conditions for evaluation on the bench. The circuit can be tested at all the limits of specified operation and changes are made to the original assumptions if necessary after evaluating the performance of the circuit. A successful design can be obtained after repetition of the procedure with parameters adjusted from measurements on the hardware.

Future trend in the field of power supply design is toward distributed power architectures for high-power, high-reliability applications. Distributed architecture uses two levels of conversion: a front-end bulk power source and a point-of-load source for end-use voltages, either at the subsystem or board level. Multiple, on-card voltages are also delivered by multiple point-of-load power supplies. Higher efficiency multiple outputs are gained by building in synchronous rectification with proper signal timing.

Other differentiating features include:

Current sharing: Two or more power supplies are operated in parallel to achieve a higher current output at the same voltage.

Power factor correction: It has become a built-in option for all the power supplies, especially for higher-power supplies.

Universal input voltage: This feature allows power supplies to be connected to the full ranges of input voltages, typically from 95 to 265 VAC, without changing jumpers or switches.

Hot swap/hot plug capability: A board or other electronics can be removed or inserted from a "live" system, allowing redundant power for mission-critical applications.

Higher power densities: Power densities can be increased for DC to DC converters through designs that enable high efficiency (e.g., synchronous rectification) and effective heat removal techniques.

Over voltage and short circuit protection: These features are common technical requirements for AC to DC switching power supplies.

Accordingly, these trends in technology continue to shape the design of next generation AC to DC switching power supplies and DC to DC converters

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