SINGLE STAGE GRID-CONNECTED MICRO-INVERTER FOR PHOTOVOLTAIC SYSTEMS

by

Nikhil Sukesh

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Abstract

This thesis concentrates on the design and control of a single stage inverter for photovoltaic (PV) micro-inverters. The PV micro-inverters have become an attractive solution for distributed power generation systems due to their modular approach and independent Maximum Power Point Tracking (MPPT). Since each micro-inverter has an individual inverter section, it is essential to have small-sized power conversion units. Moreover, these inverters should provide large voltage amplification in order to connect to the utility grid because of the low voltages of the PV panels. In order to operate these inverters at high frequencies, the soft-switching of the power MOSFETs is an important criterion to minimize the switching losses during the power transfer.

A novel Zero Voltage Switching (ZVS) scheme to improve the efficiency of a single stage grid-connected flyback inverter is proposed in this thesis. The proposed scheme eliminates the need for auxiliary circuits to achieve soft-switching for the primary switch. ZVS is realized by allowing the current from the grid-side to flow in a direction opposite to the actual power transfer with the help of bi-directional switches placed on the secondary side of the transformer. The negative current discharges the output capacitor of the primary MOSFETs thereby allowing turn-on of the switch under zero voltage. In order to optimize the amount of reactive current required to achieve ZVS a variable frequency control scheme is implemented over the line cycle. Thus the amount of negative current in each switching cycle is dependent on the line cycle.
Since the proposed topology operates with variable frequency, the conventional methods of modeling would not provide accurate small signal models for the inverter. A modified state-space approach taking into account the constraints associated with variable switching frequency as well as the negative current is used to obtain an accurate small signal model. Based on the linearized inverter model, a stable closed loop control scheme with peak current mode control is implemented for a wide range of operation. The system incorporates the controllers for both the positive as well as negative peak of inductor current.

Simulation and the experimental results presented in the thesis confirm the viability of the proposed topology.
Acknowledgements

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Nomenclature

\( L_m \) Magnetizing Inductance of the flyback transformer (H)
\( L_f \) Output filter inductor (H)
\( L \) Magnetizing Inductance reflected to secondary of transformer (H)
\( C_f \) Output filter capacitor (F)
\( C_{dc} \) Input capacitor (F)
\( C_{sn} \) Snubber capacitor for primary switch (F)
\( I_{dc} \) DC input current from solar panel (A)
\( V_{dc} \) DC input voltage from solar panel (V)
\( i_{grid} \) Grid current (A)
\( v_{grid} \) Grid voltage (V)
\( i_{sm} \) Transformer primary current (A)
\( i_{sec} \) Transformer secondary current (A)
\( i_{sec\_avg} \) Average secondary current (A)
\( i_{Lm} \) Transformer Magnetizing current (A)
\( i_{cap} \) Snubber capacitor current (A)
\( i_{prim\_ref} \) Primary current reference (A)
\( i_{sec\_ref} \) Secondary current reference (A)
\( v_{ds\_sm} \) Drain-source voltage for primary switch (V)
\( v_{ds\_sec} \) Drain-source voltage for secondary bidirectional switch (V)
\( N_p \) Number of primary turns of the transformer
\( N_s \) Number of secondary turns of the transformer
\( T_s \) Time period of a switching cycle (s)
\( t_{on} \) Time duration for which primary switch is ON (s)
\( t_{off} \) Time duration for which primary switch is OFF (s)
\( f_{s\_avg} \) Average switching frequency over AC cycle (Hz)
\( T_{s\_avg} \) Average time period over the ac cycle (s)
\( P_{dc} \) Input DC power (W)
$P_{ac}$  Averaged output AC power (W)
$B_{max}$  Maximum flux density (Tesla)
$B_{sat}$  Worst case saturation flux density (Tesla)
$l_g$  Air gap length (cms)
$A_c$  Cross sectional area of the core (cm$^2$)
$A_W$  Cross sectional area of the conductor (cm$^2$)
$W_A$  Window area of the core (cm$^2$)
$K_u$  Window utilization factor
$\rho$  Resistivity of the conductor material (Ω-m)
$I_{tot}$  Sum of the rms winding currents referred to the primary (A)
$I_{max}$  Peak current through the winding (A)
$F$  Fringing Factor
$N_p^E$  Effective number of turns in primary due to fringing effect
$B_{max}^E$  Effective maximum flux density due to fringing effect (Tesla)
$v_c$  Averaged capacitor voltage (V)
$\langle i_L(t) \rangle$  Averaged inductor current (A)
$\langle i_{in}(t) \rangle$  Averaged secondary switch current (A)
$\tilde{v}_c$  Small AC variation in capacitor voltage (V)
$\tilde{i}_L$  Small AC variation in inductor current (A)
$\tilde{v}_{in}$  Small AC variation in input voltage reflected to the secondary (V)
$\tilde{i}_m$  Small AC variation in input current (A)
$\tilde{i}_{pk}$  Small AC variation in positive peak of inductor current (A)
$\tilde{i}_{npk}$  Small AC variation in negative peak of inductor current (A)
$\tilde{t}_{on}$  Small AC variation in on-time (sec)
$\tilde{t}_S$  Small AC variation in switching period (sec)
$D$  Steady state duty ratio
# Abbreviations

<table>
<thead>
<tr>
<th>Abbreviation</th>
<th>Description</th>
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<tbody>
<tr>
<td>PV</td>
<td>Photovoltaic</td>
</tr>
<tr>
<td>MPPT</td>
<td>Maximum Power Point Tracking</td>
</tr>
<tr>
<td>ZVS</td>
<td>Zero Voltage Switching</td>
</tr>
<tr>
<td>PLL</td>
<td>Phase Locked Loop</td>
</tr>
<tr>
<td>EPIA</td>
<td>European Photovoltaic Industry Association</td>
</tr>
<tr>
<td>DC</td>
<td>Direct Current</td>
</tr>
<tr>
<td>AC</td>
<td>Alternating Current</td>
</tr>
<tr>
<td>DCM</td>
<td>Discontinuous Conduction Mode</td>
</tr>
<tr>
<td>CCM</td>
<td>Continuous Conduction Mode</td>
</tr>
<tr>
<td>BCM</td>
<td>Boundary Conduction Mode</td>
</tr>
<tr>
<td>PFM</td>
<td>Pulse Frequency Modulation</td>
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<tr>
<td>MLT</td>
<td>Mean Length per Turn</td>
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<tr>
<td>ePWM</td>
<td>Enhanced Pulse Width Modulation</td>
</tr>
<tr>
<td>PSIM</td>
<td>Power SIMulator</td>
</tr>
<tr>
<td>PLL</td>
<td>Phase Locked Loop</td>
</tr>
<tr>
<td>ANF</td>
<td>Adaptive Notch Filtering</td>
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<tr>
<td>ESR</td>
<td>Equivalent Series Resistance</td>
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Chapter 1

Introduction

Fossil fuels such as oil, coal and natural gas are the major sources of energy all over the world. The price of the non-renewable sources such as oil and natural gas are increasing as the demand for energy worldwide has grown appreciably over the past few decades. Extensive use of fossil fuels has also resulted in growing concerns of environmental problems, forcing mankind to look for alternate and clean sources of energy. In order to preserve our earth for future generations from the long lasting harmful effects of the fossil fuels, various ways to improve the electricity generation from renewable energy sources such as sun and wind are being extensively studied. Even though the renewable energy sources are inexhaustible in nature, the seasonal nature of the sources makes it difficult to operate a power system solely based on renewable energy source.

Among the various renewable energy sources, solar energy has been an attractive solution for the growing energy demand due to its availability in abundance. The growth of PV technology is evident from the rate of increase in the installed PV capacity globally as seen in Figure 1.1. According to the European Photovoltaic Industry Association (EPIA), the cumulative installed PV capacity reached almost 40GW in 2010 [1]. The major reason for the increased use of PV technology is the reduction in prices of the PV
inverters by more than 50% during the last decade. It has been possible due to the implementation of advanced technologies both in terms of the PV cell as well as the power electronics used for the power conversion [2].

![Figure 1.1: Evolution of global cumulative installed capacity PV 2000-2010 (Reproduced from [1])](image)

The unpredictable and seasonal nature of solar energy makes it unsuitable to build electrical power-grid based completely on solar energy. Integrating the PV system with the existing electrical grid provides more reliable and a better quality power to the customers [3]. The grid-connected PV system could either be used for large scale power generation (output power range from few kilowatts to megawatts) or for residential
applications (low output power range). Since the grid-connected PV systems are becoming more prominent all over the world, this thesis concentrates on grid connected topologies.

### 1.1 Power Electronics in Solar Power Generation

![Growth of Power Electronics for renewable energy generation](image)

**Figure 1.2: Growth of Power Electronics for renewable energy generation [4]**

The major role of power electronics in PV systems is to transfer the energy harnessed by PV cells to the grid with optimum possible efficiency and performance. A grid connected PV network consists of a PV panel and an inverter with controller stage to provide maximum power to the utility grid. Power Electronics, in the past few decades, has undergone a remarkable technological evolution due to the development of fast and high power semiconductors and real time controllers capable of handling the complex control algorithms. These latest advancements in the ratings of the switching devices and
components have resulted in a reduction of the cost and improvement in efficiency of the power electronic devices used in PV network. The growth in the market of power electronics for PV applications can be seen in Figure 1.2[4]. It can be observed that there is a steady increase in the investment on power electronics in past few years and is projected to have a significant increase over the coming years.

1.2 Challenges in PV System Design

Regardless of the power topology, there are certain challenges that need to be overcome for an efficient grid-connected PV system. Some of these challenges that need to be taken care are described in the following sub-sections.

1.2.1 Maximum Power Point Tracking

The power delivered by a PV module depends on the irradiance, temperature and the current drawn from the cells. The characteristics of the solar cell vary significantly with the variation of the insulation level and the temperature of the cell. The output voltage and current of the solar cell decide the operating point of the cell and thus specify the output power generated by the cell. With the varying atmospheric conditions, the power output of the PV module changes as the operating point of the solar cells shifts. Hence, it is imperative to extract the maximum power from the PV modules when there is a change in the operating conditions for efficient use of PV technology. Maximum Power Point Tracking (MPPT) is a technique employed to extract maximum power from the
non-linear PV cells throughout the day by varying the operating point of the modules[5-7].

1.2.2 Partial Shading

A single PV cell can provide only a limited amount of current and voltage when exposed to standard solar irradiance. Hence, a number of identical PV cells are interconnected and encapsulated to form a single PV module which provides sufficient current and voltage to meet the energy demands. Though the PV cells are identical in terms of their electrical behavior, they tend to exhibit different characteristics when exposed to different levels of the solar irradiance. Shading of a part of the solar panel is a major cause of a mismatch. Partial shading could occur due to tree shadow, cloud covering a portion of the panel, snow or the orientation of the PV arrays on the roof. Partial shading results in a variation of the I-V characteristics of the PV module causing a large reduction in the power generated by the panel. The effect of partial shading on the PV panel/array characteristics have been studied in the past. [8-10] The extent of the problem due to shading depends on the PV system configuration being employed. In order to overcome this problem, various MPPT algorithms are used for each solar cell. The losses due to partial shading can also be reduced by using different orientations for PV module [9].
1.2.3 Power Decoupling

In case of a single phase grid connected PV system, the power flow to the grid is time varying while the power output of PV panel must be constant to maximize its efficiency. This causes instantaneous power mismatch between the input and the output stages of the inverter. The output power oscillates with twice the grid frequency while the input to the inverter is DC. The oscillating output power is reflected to the input resulting in a deviation of operating point of the PV panel. As the PV module cannot operate at a constant power, there would be power loss which reduces the efficiency of the PV system. In order to overcome this problem, usually a large electrolytic capacitor is placed between the two stages to achieve the power decoupling [11]. However, size of the electrolytic capacitors is an issue and its lifetime is quite low thus reducing the life of a solar panel. Research is continuously going on to replace the large electrolytic capacitors with smaller film capacitors by including power decoupling circuits in the system. [11]

1.2.4 Grid Synchronization

The interconnection of the renewable energy based power generation systems to the utility grid can lead to grid instability due to the controllability issues associated with them. With stringent standards for the interconnection of these systems to the utility network, grid synchronization has become an important design challenge [12]. The main task of grid synchronization algorithm is to detect the phase angle, amplitude and frequency of the grid voltage which can be used to synchronize the control variables of
the system. The grid synchronization scheme should ideally provide a unity power factor correction by synchronizing the inverter output current with the grid voltage such that it generates a clean synchronization signal even in the presence of utility distortions. There are numerous reasons for the distortion of the utility waveforms. It could be due to the harmonics generated by the grid load appliances or the faults caused by lightning or short-circuit. Hence, all these factors should be considered while designing the grid synchronization scheme.

1.3 Thesis Contribution

The objective of the thesis is to develop a novel Zero Voltage Switching (ZVS) approach in a grid connected single-stage flyback inverter without using any additional auxiliary circuits. Since the elimination of switching losses in the primary switch is important for increasing the efficiency of the flyback inverter, an efficient scheme of power injection to the utility grid without increasing the size of the inverter is the main motive for the proposal of this ZVS scheme. The soft-switching of the primary switch is achieved by allowing negative current from the grid-side through bidirectional switches placed on the secondary side of the transformer. A variable switching frequency approach is adopted to optimize the amount of reactive current required to achieve ZVS for the primary switch over the line cycle.

Additionally the modeling of the proposed inverter scheme taking into account the effect of negative peak of the inductor current provides a comprehensive dynamical model of the inverter system. The conventional modeling techniques would result in an
inaccurate model leading to an unreliable compensator design. Further, compensator
design based on these conventional models would result distortion on the grid current.
Hence, this thesis introduces a modified state space modeling technique taking into
account these subtleties in order to obtain optimize the amount of reactive current from
the grid.

1.4 Thesis Outline

The contents of this thesis are organized into 5 Chapters.

Chapter 1 provided general information on the importance of renewable energy in
the future, the benefits of solar energy and the role of power electronics in accelerating
the growth of PV all over the world. Some of the design challenges associated with PV
systems and the objective of this thesis were also discussed.

Chapter 2 gives an insight about the current scenario of the system topologies that
are being widely used for grid-connected PV inverters. The advantages and disadvantages
of the different PV configurations are discussed. A literature review on the existing
converter topologies for the micro-inverters is presented in the latter half of the chapter.
Lastly, a basic idea about the proposed topology is presented.

Chapter 3 provides a detailed explanation of the operation of proposed single
stage inverter topology. This chapter focuses on the main circuit topology to achieve soft
switching. The design expressions and the selection criteria of the circuit parameters are
presented along with the simulation and experimental waveforms for the same.
In Chapter 4, the procedure for modeling of the proposed inverter topology is explained. The transfer function of the inverter for the compensation of the system is derived. The grid synchronization technique for unitary power factor operation of the inverter is presented in this chapter.

Chapter 5 summarizes all the features and the contributions of the proposed single stage inverter topology. The chapter concludes with possible suggestions for the future work that could be done to achieve an improvement in the proposed topology.
Chapter 2

Single Phase Grid Connected Inverter: Review

2.1 PV System Configuration

One of the major tasks in generation of solar power is the interfacing of the PV modules to the utility grid. Depending on the power level of operation, a number of configurations have been proposed over the past few decades. A brief overview of a few of these configurations is explained in the subsequent sections.

2.1.1 Centralized Configuration

In the past, photovoltaic (PV) power generation systems were based on the centralized configuration, where a number of PV panels were interfaced to the utility grid through a high power converter. The PV panels were connected in series to form arrays such that each array generated an amplified DC voltage. These series-connected arrays of PV modules are then connected in parallel through string diodes to increase the power level. The total power output from these arrays is processed by a centralized inverter for the DC to AC conversion, before connecting to the utility grid as shown in Figure 2.1(a).

Although this configuration provides a robust, efficient and inexpensive scheme for a PV system, it has a few disadvantages which have forced researchers to look for different schemes. Since there is no option of providing MPPT for independently operating
sections of the PV array, the overall output power is reduced due to mismatch in the operating condition of different sections of the array. Power losses in the string diodes, lack of flexibility in the design and the need of high voltage dc cabling between the PV arrays and the centralized inverter lower the benefits of such a configuration [13, 14].

### 2.1.2 String Configuration

String inverter configuration is an evolved version of the centralized inverter configuration. In this case, a number of PV modules are connected in series to form a string of panels providing amplified DC voltage. These series connected PV panels are connected to a grid-tied inverter such that each PV string has its own inverter as shown in Figure 2.1(b). Unlike the centralized configuration, it does not need the string diodes thereby minimizing the losses. The presence of individual inverters for each PV string results in dedicated MPP tracking for each string [13].

Multi-String configuration is a similar approach, as shown in Figure 2.1(c), where each string of PV panels is first connected to a DC-DC converter. These converters are then interfaced to a common grid-tied DC-AC inverter. Each PV string can have individual MPPT while the use of a single grid-tied inverter reduces the cost as compared to the string configuration. These configurations are useful for medium power applications (2kW – 5kW range) [14].

In spite of the advantages, the string and multi-string configurations have DC wiring in order to attain higher voltage levels. Secondly, the MPP tracking is not
available for individual PV modules. As a result, these configurations are not feasible for low power residential applications.

**Figure 2.1: Different Configurations for PV Inverters**
2.1.3 AC Modules / Micro-inverters

In order to overcome the drawbacks associated with the centralized approach, the concept of the AC modules has been introduced in the recent past. The need of long dc wires could be overcome by using grid tied DC-AC inverters for each PV panel as shown in Figure 2.1(d). The AC module is the integration of the PV panel and a small grid-tied inverter in one electrical device to harvest the energy [15]. In this modular approach, the AC module performs maximum power point tracking (MPPT) of the PV panel, voltage amplification, galvanic isolation, and injection of the high quality AC current to the utility grid. The individual MPPT for each module provides an efficient scheme to overcome the detrimental effects of partial shading and the mismatch losses [15]. In addition, the modular structure provides a very easy and practical solution for expansion of the power generation system. The compact nature, high efficiency and scalability make micro inverters a good candidate for low power (below 500W) residential applications. [14][15][16]

Despite the numerous advantages of the modular approach, there are a few complications that should be considered for the design of a micro-inverter.

- The size and number of components used in the design of micro inverter is a major concern as the inverter is installed at the back of each PV module. In order to achieve individual MPP tracking, a large electrolytic capacitor is required for decoupling. Hence the limitation of space on the back of the modules is usually a major concern in this regard.
• The electrolytic capacitor reduces the overall life span of the micro inverter. The design of the inverter tends to become complicated on trying to replace the electrolytic capacitor as additional circuits are needed to achieve power decoupling.

• In case of a fault in the inverter, replacement of the inverter is a difficult and costly proposition. Even though the installation cost is lower as compared to other configurations, the maintenance cost of micro-inverters is quite high. [14].

2.2 Review of Single Phase Micro-inverter Topologies

The major tasks performed by a grid connected micro-inverter include extraction of DC power from the panel, MPPT, voltage amplification to achieve grid voltage and injection of the sinusoidal current to the grid. It is possible to categorize the inverter topologies based on the schemes used for distribution of these tasks. One of the common classifications is based on the number of stages of power processing. The DC to AC power conversion within the PV module could be carried out either in a single stage or multiple stages. The block diagram for both multi-stage and single-stage inverter topologies are shown in Figure 2.2. The subsequent section takes a look at the features of various existing topologies commonly used for micro-inverters.
2.2.1 Multi-Stage Micro-inverters

In multi-stage topologies, the tasks performed by the grid connected system are distributed into two stages. Extraction of the maximum power from the PV panel and voltage amplification is usually carried out by a DC-DC converter in the first stage. The second stage is an inverter, which injects high quality current to the utility grid. The output of the converter in the first stage could be either a pure DC voltage or a rectified AC current. In case of a pure dc output voltage, DC-DC converter has to only handle the nominal power and inverter in the second stage controls the grid current by means of pulse width modulation. While the DC–DC converter has to handle twice the nominal power in case of a rectified AC current at the output of the first stage and the inverter stage switches at line frequency to unfold the rectified current.

The multi-stage topology has a boost type DC-DC converter section which provides the necessary voltage amplification at the DC link. The increased voltage at the DC link helps in reduction of the capacitance value at the link. A lower capacitance eliminates the use of electrolytic capacitors thereby increasing the life span of the micro-
inverters. A two-stage micro inverter with a simple power stage and control scheme was proposed in [17] as shown in Figure 2.3. The topology consists of a DC-DC flyback converter as the first stage for voltage amplification and galvanic isolation. The second stage has a full bridge inverter that carries out the AC current injection to the grid. Though, it provided a robust operation, it suffers from high switching losses as all the power MOSFET’s are hard switched in this topology.

![Two Stage PV system with flyback converter to boost DC voltage](image)

**Figure 2.3: Two Stage PV system with flyback converter to boost DC voltage**

Since the size of the inverter in AC modules is a major criteria, it is important to achieve high efficiency with minimum components. In [18], a boost converter operating under Discontinuous Conduction Mode (DCM) is used to generate a high stable DC voltage. The boost converter uses Pulse Train control algorithm which provides a fast dynamic response and a robust operation [19]. A high frequency flyback inverter with peak current mode control is employed for high quality AC current generation. The flyback inverter configuration is simple and uses fewer semiconductor switches as shown in Figure 2.4. Although this configuration reduces the overall cost of the system, there is
a significant loss associated with this topology as soft switching is not achieved in most of the switches.

![Diagram of a Flyback Inverter with a DC-DC boost converter for voltage amplification](image)

**Figure 2.4: Flyback Inverter with a DC-DC boost converter for voltage amplification**

In order to increase the efficiency of the multi-stage micro-inverters, different schemes need to be employed to achieve soft switching in either DC-DC or DC-AC conversion stages or both stages of the micro-inverter. A two stage micro-inverter consisting of a high efficiency step-up DC-DC converter and a full bridge DC-AC inverter was implemented in [20]. An active clamping flyback converter with a voltage-doubler rectifier is used in the first stage which reduces the switching losses by eliminating the reverse-recovery current of the output rectifying diodes. Figure 2.5 shows the two stages of the micro-inverter.
Another approach that is being implemented in a number of multi-stage micro-inverters, to improve the efficiency, is the use of resonant circuits to achieve ZVS for the power MOSFET’s [21-23]. An example of the resonant converter for PV application is the topology in Figure 2.6 which uses a two-inductor boost converter operating under resonance condition to provide the soft switching of the power MOSFET’s [21]. The boost converter operates under variable frequency to secure an adjustable output voltage range while maintaining the resonant switching transitions.
Soft commutation of the switches with the help of Capacitive Idling technique has been introduced in [24]. This circuit topology is derived from SEPIC converter and a two-switch flyback inverter such that soft switching is obtained for the active switches as shown in Figure 2.7. The major disadvantage with the proposed topology is the increase in conduction losses in the boost MOSFET since it carries both the input current and the reflected output current. A comparison of some of the existing micro-inverters with a DC link has been presented in [13][25].

![Figure 2.7: Inverter based on capacitive idling technique](image)

Although the two stage configuration is a very straightforward and simple power conversion technique, it has some essential limitations. The main limitations of the two-stage converter are the size and the maximum efficiency achievable to interface the PV panel to the utility grid. Since there are two stages of power conversion, the efficiency is inherently limited and also power density of the converter is compromised by the two stages. The other setback of the two-stage configuration is that it is very difficult to
realize ZVS for the second stage of the converter. Therefore, usually a low switching frequency PWM inverter is used to minimize the inevitable switching losses of the second stage. Operating with low switching frequency requires a bulky and lossy filter to remove the high frequency component and inject a high quality current to the grid.

2.2.2 Single Stage Micro-inverters

In a single stage micro-inverter, all the tasks are performed by a single DC-AC inverter. Since there would be power oscillations at twice the grid frequency, the inverter would have to be designed to handle a peak power equal to twice of the nominal power. The size of the inverter can be reduced due to lower component count in single stage inverters as compared to the multi-stage inverters. A micro-inverter having dedicated converter for voltage amplification and inverter for sine wave grid current generation results in increased number of components and its overall size. With the recent trend of miniaturization, compact single stage micro inverters have become the point of interest [26-28]. Since the single stage inverters do not have the flexibility of reduction of the size of capacitor size at the dc link, power decoupling needs to be taken care by either a large electrolytic capacitor at the input terminals or by use of auxiliary circuits [29-32].

A flyback inverter is one of the common topologies considered for PV modules since it provides a simple circuit configuration to achieve direct conversion of the DC power to AC power. The flyback topology has proved to provide a reliable and cost effective topology with reduced number of semiconductor switches [33]. Another important feature of flyback topology is the provision of isolation due to the use of high
frequency transformer. These benefits of a flyback inverter make it a viable solution for the single stage micro inverters.

The flyback inverter can operate in Continuous Conduction Mode (CCM) or Discontinuous Conduction Mode (DCM) to process the power. The CCM operation is not a feasible solution as the inverter tends to act as a load independent voltage source due to the incomplete discharge of the magnetizing inductance of the transformer [34]. Another major issue with the CCM operation is the presence of right half plane zero in the output current to the duty cycle transfer function, which introduces a challenge in controlling the output current. Charge control was adopted to control the input current waveform instead of the output current in [35] to operate the inverter in CCM. However, the charge control scheme may cause sub-harmonic oscillations. It, also, may result in a constant change in the operating mode of the inverter between CCM and DCM, under certain operating conditions. As a result, the DCM has been more attractive solution in the past for flyback inverters even though the voltage and current stress are higher on the power switches.

In [36], the analysis of the conventional flyback inverter has been performed. Figure 2.8 shows the conventional single stage flyback inverter operating in DCM. The efficiency of the flyback inverter is fairly low due to the hard switching of the power MOSFETs in the conventional configuration.
In [37], a soft switching scheme is proposed for flyback inverters through actively switched snubber circuit in parallel to the primary switch. However, adding extra active and passive components increases the system complexity and offsets the advantage of the proposed approach. Figure 2.9 shows the additional components used to achieve ZVS.

Figure 2.8: Conventional grid connected flyback inverter

Figure 2.9: Flyback Inverter with actively switched snubber for ZVS
Active clamp circuits are also commonly used in flyback inverters in order to achieve soft-switching and clamp the voltage across the switch thereby reducing the voltage stress on power MOSFETs [38, 39]. Although, the switching losses and the voltage stress in the primary switch are mitigated with the active clamp approach, the additional switch used in the clamp circuit is usually hard-switched which results in additional switching losses especially at high switching frequencies. The above mentioned topologies use a high frequency center tapped transformer for the unfolding stage. Figure 2.10 shows a flyback topology which avoids the use of center tapped transformer thereby improving the magnetic design [40]. It uses two flyback converters, one for each half of the ac cycle, and bidirectional switches on the secondary side.

![Flyback Inverter with full bridge configuration on primary side](image)

Figure 2.10: Flyback Inverter with full bridge configuration on primary side

Resonant based flyback inverters are also the other approach to obtain soft switching for flyback inverters [41, 42]. In [42], a single-stage quasi-resonant flyback inverter with a Pulse Frequency Modulation (PFM) control scheme has been proposed as
shown in the Figure 2.11. The PFM uses a constant on-time and a variable frequency to control the energy transfer from the input to the output. The resonant capacitor in the secondary winding helps provide soft switching.

Figure 2.11: Single-stage Flyback with PFM control

Despite the benefits of achieving ZVS in DCM operation of flyback inverters, there is still a limit on the transferable power due to the inevitable dead-time required for the complete discharge of the transformer magnetizing energy in each switching cycle. In order to increase the limit of power transfer capability, the boundary conduction mode (BCM) has been proposed, which eliminates the dead-time [34, 43]. In [34] the features of a flyback inverter in constant frequency DCM and a variable frequency BCM have comprehensively been analyzed. The control scheme used for BCM operation is, however, more complicated than the conventional DCM operation due to the addition of secondary current sensor. Since higher power can be achieved in BCM operation, a dual switching strategy was proposed in [43], where the mode of operation changes between the BCM and DCM according to the input conditions. Even with an increase in the power
level, the switching losses contribute towards reduced efficiency levels in these references.

2.3 Concept of Proposed Topology

In order to overcome the aforementioned problems, a novel Zero Voltage Switching (ZVS) approach in a grid connected single-stage flyback inverter without using any additional auxiliary circuits. The flyback inverter is operated in modified BCM in order to maximize the transferable output power. The proposed modified BCM operation allows the charging of the transformer magnetizing inductance in the reverse direction of power flow to provide reactive current for soft-switching. The combination of a switch and diode in the conventional flyback inverters in the secondary is replaced by a bi-directional switch to permit the current to conduct from the grid-side to the primary-side. Thus, the presence of negative current during the turn ON of primary switch results in a soft-switched turn ON of the primary MOSFET. Further, the secondary MOSFET that replaces the diode is turned ON under ZVS thereby improving the overall efficiency of the inverter. The proposed concept requires an efficient control of the amount of reactive current to attain high quality current injection to the utility grid. Hence, a variable frequency approach for the switching of the MOSFETs based on the amount of reactive current is introduced in the proposed scheme. The working principle, detailed analysis and the development of the control scheme of the inverter form the gist of the thesis.
2.4 Summary

In this chapter the different configurations for interfacing of the PV panels to the utility grid were reviewed. It was observed that the benefits of AC module concept and their utility made it an ideal candidate for low power PV systems. Thus, a brief review of the features of existing topologies for the AC modules was conducted. A look into the features of these topology showed the various issues related to previously proposed inverters such as

- Lower efficiency due to hard-switching of power MOSFET’s
- Presence of auxiliary circuits to achieve soft-switching.
- Larger component count for two stage inverters
- Limitation in power transfer for DCM operation of flyback based topologies

The review especially concentrates on the flyback topologies because of its reliability and the simplicity due to reduced number switches. A brief idea of the proposed flyback inverter topology is presented in this chapter.
Chapter 3

Proposed Micro-Inverter: Main Circuit Topology

The review of the present scenario of PV generation for residential applications gives an insight into the drawbacks associated with the existing topologies. It provides an understanding of the factors that need to be considered while designing the micro-inverter. On the basis of the review, a single-stage inverter is proposed, which offers soft-switching for the power MOSFETs, leading to a very efficient and compact solution to interface the PV panel to the utility grid. This chapter concentrates on the detailed working of the main circuit topology.

3.1 Principle of Operation

Figure 3.1 shows the circuit diagram of the proposed topology based on a single stage flyback inverter. The topology consists of a center tapped flyback transformer, a primary switch, input decoupling capacitor, bidirectional switches on the secondary and the output filter. The proposed circuit is similar to a conventional flyback inverter with bi-directional switches in the transformer secondary providing the unfolding of the grid current. Proper switching of the bi-directional switches on the secondary helps in achieving soft commutation for most of the switches without the need of any additional circuitry.
Figure 3.1: Circuit configuration of the proposed flyback inverter

The primary switch $S_m$ is triggered to charge up the magnetizing inductance up to the required reference as is the case with a conventional flyback converter. The energy stored in the magnetizing winding is transferred to the grid by turning ON either the secondary switch $S_{acp1}$ during the positive or $S_{acn1}$ during the negative half cycle respectively. On completion of the power transfer to the grid, the magnetizing inductance of the transformer is allowed to charge in the opposite direction with the help of the bidirectional switches on the secondary of the transformer. Figure 3.2 shows the detailed modes of operation of the proposed circuit in one switching cycle. Since the flyback inverter operates under modified BCM, the switching frequency is variable. A switching cycle is divided into six modes of operation. Since the switching frequency of the inverter is significantly higher than the grid frequency, the quantities varying with respect to the grid frequency such as the grid voltage, grid current, duty cycle and reference current are considered constant during one switching cycle.
Figure 3.2: Operation waveforms of proposed inverter in a switching cycle
Mode I (Interval $0 < t < t_1$)

The primary switch $S_m$ is turned ON while all the other switches remain OFF. Figure 3.3 shows the components that are active during this interval. The input PV voltage is applied across the magnetizing inductance $L_m$ of the transformer primary. Hence, the primary current increases with a slope dependent on the voltage applied across the inductor. The primary current can be expressed as:

$$i_{sm}(t) = \frac{V_{dc}}{L_m} t - i_{sm_{pk_{n}}}$$

(3.1)

When the current reaches the peak value $i_{sm_{pk_{p}}}$ at $t = t_1$, primary switch is turned OFF. The peak value of the switch current is given by the expression:

$$i_{sm_{pk_{p}}} = \frac{V_{dc}}{L_m} d_1 T_s - i_{sm_{pk_{n}}}$$

(3.2)

where, $d_1$ is the duty cycle

$t_{on} = d_1 T_s$ is the duration for which the primary switch is ON.

The value of $i_{sm_{pk_{p}}}$ depends on the instantaneous value of the ac output power. The average value of the input current can be expressed by integrating the primary current over the switching cycle.

$$i_{avg} = \frac{L_m}{2V_{dc} T_s} \left( i_{sm_{pk_{p}}}^2 - i_{sm_{pk_{n}}}^2 \right)$$

(3.3)

One of the secondary bidirectional switches ($S_{acp1}$ or $S_{acn1}$) depending on the ac cycle) can be turned ON during this interval as the transformers are connected similar to a flyback operation.
Mode II (Interval $t_1 < t < t_2$)
Switch $S_m$ is turned OFF at the beginning of the interval while the secondary switch $S_{acp1}$ is ON as shown in Figure 3.4. The drain-source voltage across switch $S_m$ starts to rise slowly as the capacitor $C_{sn}$ across the switch starts charging. The polarity of the transformer is such that the secondary current does not flow even though the secondary switch $S_{acp1}$ remains ON. This mode comes to an end when the capacitor has charged to its maximum value of $V_{dc} + Nv_{grid}$. This transition from the ON state to OFF state of switch $S_m$ occurs within a very short interval of time.
Mode III (Interval $t_2 < t < t_3$)

During this interval, either the secondary switch $S_{acp1}$ or $S_{acn1}$ depending on the positive or negative half cycle of the grid voltage is ON. Hence, the energy stored in the magnetizing inductance of the transformer is released to the grid. Figure 3.5 highlights the active switch $S_{acp1}$ and the anti-parallel diode of switch $S_{acp2}$ during this interval.

Figure 3.5: Equivalent Circuit for Mode III (positive half cycle)
Mode IV (Interval $t_3 < t < t_4$)

At the beginning of this mode of operation, the anti-parallel diode of the switches $S_{acp2}$ or $S_{acn2}$ are conducting depending on the polarity of grid voltage. As the current in the magnetizing winding tends to zero, either $S_{acp2}$ or $S_{acn2}$ is switched ON with zero voltage depending on the sign of grid voltage. The secondary current $i_{sec}$ would continue to flow until it reaches zero. Since the bidirectional switches are ON, the current in the secondary side changes direction and starts charging the magnetizing inductance in the opposite direction. At the instant when $i_{sec}$ equals the reference value, the secondary switches are turned OFF. Figure 3.6 shows that both the switches $S_{acp1}$ and $S_{acp2}$ are ON during this interval allowing the secondary current to reverse its direction. Since the voltage observed across the magnetizing inductance of the transformer remains identical during the modes III and IV, expressions for current and voltage can be considered as the same for the two intervals.

![Figure 3.6: Equivalent Circuit for Mode IV (positive half cycle)]
Thus, in the interval \( t_1 < t < t_4 \), the absolute value of secondary current can be expressed as

\[
|i_{sec}| = \left| \frac{-v_{grid}(t - t_1) + \frac{i_{sm_{- pk_{- p}}}}{N}}{N^2 L_m} \right|
\]  

(3.4)

\( N \) denotes the turn’s ratio of the transformer \((N_s/N_p)\). At the instant \( t = t_4 \), the secondary current equals the \( i_{sm_{- pk_{- s}}}/N \). Substituting the value of secondary current in (3.4), the expression for the duty cycle for the interval when the secondary switches are ON can be obtained as

\[
t_4 - t_1 = (d_2 + d_3)T_S = \frac{(i_{sm_{- pk_{- p}}} + i_{sm_{- pk_{- n}}})NL_m}{v_{grid}}
\]  

(3.5)

where, \( d_2 \) and \( d_3 \) are the duty cycle for intervals III and IV respectively.

**Mode V (Interval \( t_4 < t < t_5 \))**

This mode begins when the bidirectional switches on the secondary are turned OFF as shown in Figure 3.7. Thus all the switches are in OFF state during the interval. As the magnetizing inductance \( L_m \) was charged in reverse direction in the previous interval, current \( i_{Lm} \) flows in the direction so as to transfer the energy stored in \( L_m \) to the input capacitor. The voltage across the switch \( S_m \) starts decreasing slowly as the capacitor across the switch starts discharging. This mode comes to an end with the complete discharge of the capacitor across the switch.

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Mode VI (Interval $t_5 < t < t_6$)

As the capacitor is completely discharged, the drain-source voltage of the primary switch $S_m$ nearly equals zero. It causes the anti-parallel diode of the switch to conduct as shown in Figure 3.8. Hence the magnetizing current $i_{Lm}$ continue to increase from its negative value. Since the drain source voltage has been forced to zero, the primary switch can be turned on with ZVS during this interval.

**Figure 3.7: Equivalent Circuit for Mode V**

**Figure 3.8: Equivalent Circuit for Mode VI**
It can be observed that *Modes II, V and VI* occur during the transition between the switching of the different switches in the circuit. As a result, the time intervals of these modes are quite small as compared to the *Modes I, III and IV*. Also, it is clear from the circuit operation that the duty cycle is basically split into two parts such that

\[
d_2 + d_3 = (1 - d_1)
\]  

(3.6)

Hence expression (3.5) can be written as

\[
(1 - d_1)T_s = \frac{(i_{sm \_ pk \_ p} + i_{sm \_ pk \_ n})NL_m}{v_{grid}}
\]  

(3.7)

The high frequency component present in the secondary current is filtered by an L-C output filter such that the average value of the secondary current during the switching interval \( T_s \) is obtained by integrating the \( |i_{sec}| \) over the period.

\[
|i_{sec\_avg}| = \frac{L_m}{2v_{grid}T_{s\_avg}}(i_{sm \_ pk \_ p}^2 - i_{sm \_ pk \_ n}^2)
\]  

(3.8)

The above expressions were derived for one switching cycle. In order to have a holistic view of the operation of the circuit, the waveforms over an AC cycle are shown in Figure 3.9.
The peak of the magnetizing current is controlled according to the sinusoidal reference $i_{\text{prim \_ref}}(\omega t)$. Similarly, the peak of the secondary current that charges the magnetizing winding in opposite direction is controlled according to the sinusoidal reference $i_{\text{sec \_ref}}(\omega t)$.

\begin{align}
  v_{\text{grid}} &= V_{\text{grid \_pk}} \sin \omega t \\
  i_{\text{grid}} &= I_{\text{grid \_pk}} \sin \omega t
\end{align}  

(3.9)  

(3.10)
\[ i_{sm_{pk_{p}}} (\omega t) = i_{prim_{ref}} (\omega t) = I_{prim_{ref peak}} \sin(\omega t) \quad (3.11) \]

\[ i_{sm_{pk_{n}}} (\omega t) = N_i_{sec_{ref}} (\omega t) = I_{sec_{ref peak}} \sin(\omega t) \quad (3.12) \]

The polarity of the average secondary current is determined by the switch \( S_{acp} \) (for positive half cycle) and \( S_{acn} \) (for negative half cycle) for injection of the AC current into the utility grid. The magnitude of the AC current injected is obtained from the equations (3.8) - (3.12) and is given as

\[ i_{grid} = \frac{L_m}{2V_{grid_{pk}} T_s} (I^2_{prim_{ref peak}} - I^2_{sec_{ref peak}}) \sin(\omega t) \quad (3.13) \]

### 3.2 Design of Circuit Parameters

The magnetizing current of the transformer for BCM operation is shown in Figure 3.10. The ON time of the primary switch is variable while the OFF time remains constant during a switching cycle. The switching period for the flyback inverter can be expressed as

\[ T_s (\omega t) = t_{on} (\omega t) + t_{off} (\omega t) \quad (3.14) \]

As the peak of the magnetizing current needs to follow a sinusoidal waveform, the ON time of the switch needs to be controlled according to (3.15).

\[ t_{on} (\omega t) = t_{on_{pk}} \sin \omega t \quad (3.15) \]
Where, \( t_{on\_pk} \) is maximum duration for which the primary switch is turned ON in the AC cycle. It occurs at \( \omega t = \pi/2 \) when the maximum power needs to be transferred to the output. The ON time and OFF time can be expressed in terms of the circuit parameters with the help of equations (3.2) and (3.7).

\[
t_{on}(\omega t) = \left( i_{sm\_pk\_p}(\omega t) + i_{sm\_pk\_n}(\omega t) \right) \frac{L_m}{V_{dc}} \tag{3.16}
\]

\[
t_{off} = \left( i_{sm\_pk\_p}(\omega t) + i_{sm\_pk\_n}(\omega t) \right) \frac{N L_m}{V_{grid}(\omega t)} \tag{3.17}
\]

It can be seen from the equations (3.15), (3.16) and (3.17) that the OFF time is a constant and can be expressed as

\[
t_{off} = \frac{NV_{dc}}{V_{grid\_pk}} t_{on\_pk} \tag{3.18}
\]
The average switching frequency over the ac cycle is obtained to facilitate the designing of the circuit parameters. It is defined as

\[
f_{s_{\text{avg}}} = \frac{1}{T_{s_{\text{avg}}}} = \frac{1}{\frac{\pi}{0} T_s(\theta) d\theta}
\]

\[
= \frac{1}{t_{\text{on}_{\text{pk}}}} \left( \frac{2}{\pi} + \frac{NV_{dc}}{V_{\text{grid}_{\text{pk}}}} \right)
\]

where, \( \theta = \omega t \)

### 3.2.1 Design of Transformers Turn’s Ratio

The selection of the transformer turn’s ratio is the most important aspect in the magnetic design of the transformer. The selection of the turn’s ratio depends on the maximum duty cycle of the switch. The duty cycle can be expressed in terms of the turns ratio, input and output voltage by comparing equations (3.2) and (3.7).

\[
d_i(\theta) = \frac{v_{\text{grid}}(\theta)}{v_{\text{grid}}(\theta) + NV_{dc}}
\]

The plot of instantaneous value of duty cycle over the AC cycle for different values of transformer turns ratio is shown in Figure 3.11. It can be observed from the plot that the peak value of duty cycle over the AC cycle is reduced with the increase in turns ratio. The maximum duty cycle is reached at the instant when maximum power has to be delivered to the grid. The maximum duty cycle needed to transfer maximum power to output is increased to around 70-80% for lower turn’s ratio. The variation of maximum
duty cycle with the variation in turns ratio for different input voltages is shown in Figure 3.12. Hence, a lower turn’s ratio is not a feasible choice as the higher peak of duty cycle could result in an incomplete discharge of the magnetizing inductance for a particular range of operating frequency. In order to design an efficient inverter, a turn’s ratio of 1:5 was chosen for the transformer.

![Figure 3.11: Variation of duty cycle with turns ratio (N) for V\textsubscript{dc} = 45V](image)

Figure 3.11: Variation of duty cycle with turns ratio (N) for $V_{dc} = 45V$
3.2.2 Design of Magnetic Inductance for Flyback Transformer

For a flyback topology, the value of magnetic inductance decides the amount of power that can be transferred to the secondary for a particular operating frequency. In order to optimum value of magnetic inductance of the transformer for maximum efficiency, the relationship between the input power and the power transferred to the grid with respect to the circuit parameters needs to be investigated. Assuming 100% efficiency and negligible leakage inductances of the transformer, the averaged dc input power has to be equal to the averaged ac power.
\[ P_{dc} = P_{ac} \]  \hspace{1cm} (3.21)

\[ V_{dc} I_{dc} = \frac{1}{2} V_{grid\_pk} I_{grid\_pk} \]  \hspace{1cm} (3.22)

The input DC current can be obtained by integrating average primary current over the ac cycle.

\[ I_{dc} = \frac{1}{\pi} \int_0^\pi i_{avg}(\theta) d\theta \]  \hspace{1cm} (3.23)

On integration of the average primary current \( i_{avg} \), derived for the on time of the switch, the DC current takes the form

\[ I_{dc} = \frac{I_{prim\_ref\_peak} - I_{sec\_ref\_peak}}{2} \left( \frac{NV_{dc}}{V_{grid\_pk}} \right) \]  \hspace{1cm} (3.24)

where,

\[ F\left( \frac{NV_{dc}}{V_{grid\_pk}} \right) = \frac{1}{\pi} \int_0^\pi \frac{\sin^2 \theta}{\sin \theta + \frac{NV_{dc}}{V_{grid\_pk}}} d\theta \]  \hspace{1cm} (3.25)

Equation (3.25) can be analytically solved by using the formula

\[ \frac{1}{\pi} \int_0^\pi \frac{\sin^2 \theta}{\sin \theta + \frac{NV_{dc}}{V_{grid\_pk}}} d\theta = \left[ \frac{2}{\pi} - \left( \frac{NV_{dc}}{V_{grid\_pk}} \right) + \left( \frac{NV_{dc}}{V_{grid\_pk}} \right)^2 \right] S\left( \frac{NV_{dc}}{V_{grid\_pk}} \right) \]  \hspace{1cm} (3.26)

where,
\[ S \left( \frac{NV_{dc}}{V_{grid\_pk}} \right) = \frac{1}{\pi} \int_{0}^{\pi} \frac{d\theta}{\sin \theta + \frac{NV_{dc}}{V_{grid\_pk}}} \]

\[ = \frac{2}{\pi} \tan^{-1} \left( \sqrt{\frac{NV_{dc}}{V_{grid\_pk}}} - 1 \right) \quad (3.27) \]

Substituting the expression for \( I_{dc} \) in (3.22) gives

\[ (I_{prim\_ref\_peak} - I_{sec\_ref\_peak}) = \frac{V_{grid\_pk}I_{grid\_pk}}{V_{dc}F \left( \frac{NV_{dc}}{V_{grid\_pk}} \right)} \quad (3.28) \]

Hence, the design equation for the magnetizing inductance of the transformer can be determined from equations (3.15), (3.16) and (3.28)

\[ L_m = \frac{V_{dc}t_{on\_pk}}{2I_{sec\_ref\_peak} + \frac{V_{grid\_pk}I_{grid\_pk}}{V_{dc}F \left( \frac{NV_{dc}}{V_{grid\_pk}} \right)}} \quad (3.29) \]

The secondary reference current \( i_{sec\_ref}(\omega t) \) is a small quantity independent of the input power as its purpose is to provide the necessary current at the primary to achieve ZVS for the primary switch. Since the circuit operates under a variable switching frequency, the expression for instantaneous values of frequency can be obtained by substituting \( t_{on\_pk} \) in the above expression.
\[
L_m = \frac{V_{dc}}{2I_{sec\_ref\_peak} + \frac{V_{grid\_pk}I_{grid\_pk}}{V_{dc}F\left(\frac{NV_{dc}}{V_{grid\_pk}}\right)}} \times \frac{1}{f_{s\_avg}\left(\frac{2}{\pi} + \frac{NV_{dc}}{V_{grid\_pk}}\right)}
\]

(3.30)

The plot of the variation of magnetizing inductance with the average operating frequency provides a good basis for determination of the value of \(L_m\) for an efficient design.

**Figure 3.13: Variation of Magnetic Inductance with Average Frequency**

Figure 3.13 shows the plot of \(L_m\) versus \(f_{s\_avg}\) for different turns ratio for a 250W proposed flyback inverter. The power transfer to output is dependent on the amount of energy that can be stored in the magnetizing inductance when the primary switch is ON. This energy storage is limited by the peak of primary switch current, the value of
magnetizing inductance and the switching frequency. A higher peak current results in a complex design of the transformer windings. The choice of average operating frequency and the corresponding value of $L_m$ for a particular turns ratio can be obtained from the plot in order to keep the peak current within reasonable limits. A magnetizing inductance of $23\mu\text{H}$ is used for the transformer.

3.2.3 Flyback Transformer Design

Since the transformer is the most important factor that determines the performance of a flyback inverter, it is imperative to design a transformer with minimal losses. In a flyback transformer, the current flows only in the primary side while the core is charged and in the secondary side only when the core is discharged. The design of the transformer depends on the mode of operation of the inverter. In the case of either DCM or BCM operation, high peak-to-peak ripple inductor current at the primary results in large flux swings. A large variation in flux density leads to high core losses thereby reducing the efficiency of the transformer. This section explains the procedure followed for the selection of the core and the various parameters of the flyback transformer.

There are certain design constraints that need to be considered for building an efficient transformer [44]. The major four are:

- **Maximum Flux Density**: The peak current through the winding ($I_{\text{max}}$) decides the maximum flux density ($B_{\text{max}}$) at which the transformer can operate. The core should be selected such that the $B_{\text{max}}$ is less than the worst case saturation flux density ($B_{\text{sat}}$) of the core material. The relation between the flux density and the
peak current forms the first design constraint. The constraint is based on the assumptions that the reluctance of the air gap is much larger than the reluctance of the core and the leakage inductances are negligible.

\[ N_p I_{\text{max}} = B_{\text{max}} \frac{l_g}{\mu_0} \]  

(3.31)

where, \( N_p \) is the number of turns in primary

\( l_g \) is the air gap length

The number of turns and the air gap length should be chosen such that the maximum flux density is less than \( B_{\text{sat}} \) for the known value of \( I_{\text{max}} \).

- **Inductance:** The value of magnetizing inductance required for the rated power that needs to be obtained, forms the second constraint.

\[ L_m = N_p^2 \frac{\mu_0 A_c}{l_g} \]  

(3.32)

where, \( A_c \) is the cross sectional area of the core

- **Winding Area:** The area of the core available for the windings forms the third important constraint. The constraint can be expressed as

\[ K_u W_A \geq N_p A_W \]  

(3.33)

\( K_u \), window utilization factor, is the fraction of the core window area that is filled with copper.

\( W_A \) is the window area of the core

\( A_W \) is the cross sectional area of the conductor
The constraint thus states that the area available for winding the conductors on the core should be greater than the area occupied by the total number of turns of windings.

- **Winding Resistance:** The relation between winding resistance, the mean-length per turn (MLT) and the area of the conductor forms the fourth major design constraint.

\[
R = \frac{\rho N_p (MLT)}{A_w}
\]  

(3.34)

\( \rho \) is resistivity of the conductor material

The total copper loss in the windings is given by:

\[
P_{Cu} = \frac{\rho (MLT) N_p^2 I_{tot}^2}{W_A K_u}
\]  

(3.35)

\( I_{tot} \) is the sum of the rms winding currents referred to the primary

The elimination of the unknown quantities from(3.35) and rearranging the equation gives a relationship between the terms dependent on core geometry and the specifications. The terms dependent on the core material can be defined as the constant \( K_g \).

\[
K_g \geq \frac{\rho L_m^2 I_{tot}^2 I_{max}^2}{B_{max}^2 K_u P_{Cu}}
\]  

(3.36)

As the value of \( K_g \) can be determined from the datasheet of the core manufacturer, the above equation governs the selection of the core based on the specifications of the
inverter circuit. Once the core is selected, various transformer parameters are obtained as follows:

- The maximum flux density that can be handled by the core without getting saturated is obtained from the manufacturer’s datasheet. In order to calculate the parameters, an initial value of $B_{\text{max}}$ lower than the value in the datasheet is assumed.

- Based on the specifications, the number of turns required for the primary windings is calculated. The fringing flux at near the air gap is not considered in this case.

$$N_p = \frac{L_m I_{\text{max}}}{B_{\text{max}} A_c} \times 10^4 \text{ turns}$$  \hspace{1cm} (3.37)

- The length of air gap that needs to be provided is determined from expression (3.38).

$$l_g = \frac{0.4 \pi N_p^2 A_c}{L_m} \times 10^{-8} \text{ cm}$$  \hspace{1cm} (3.38)

- The fringing flux decreases the total reluctance of the magnetic path and hence increases the inductance by a factor $F$ as compared to the value obtained from the inductance equation.

$$F = 1 + \left( \frac{l_g}{\sqrt{A_c}} \right) \ln \left( \frac{2G}{l_g} \right)$$  \hspace{1cm} (3.39)

Where value of $G$ is dependent on the shape and size of the core
The effect of the fringing flux is to reduce the number of turns required for
the windings such that the new number of turns is given by

\[ N_p^F = \sqrt{\frac{l_g L_m}{0.4\pi A_F}} \times 10^4 \text{ turns} \]  \hspace{1cm} (3.40)

- As the number of turns decreases due to fringing effect, the value of maximum
  flux density would increase as compared to initially assumed value. The new
  value of maximum flux density is given by

\[ B_{max}^F = F \left( \frac{0.4\pi N_p^F I_{max}}{l_g} \right) \times 10^{-4} \text{ Tesla} \]  \hspace{1cm} (3.41)

The maximum flux density taking fringing effect into consideration should be
lower than the \( B_{sat} \) of the core.

### Table 3.1: Circuit Parameter Values for Proposed Topology

<table>
<thead>
<tr>
<th>Circuit Parameters</th>
<th>Symbols</th>
<th>Values</th>
</tr>
</thead>
<tbody>
<tr>
<td>Output Power</td>
<td>( P_{ac} )</td>
<td>250W</td>
</tr>
<tr>
<td>DC Input Voltage</td>
<td>( V_{dc} )</td>
<td>35V-75V</td>
</tr>
<tr>
<td>Input Capacitor</td>
<td>( C_{dc} )</td>
<td>5mF</td>
</tr>
<tr>
<td>Grid Voltage</td>
<td>( V_{ac} )</td>
<td>220V</td>
</tr>
<tr>
<td><strong>Centre Tapped Flyback Transformer</strong></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Turns Ratio</td>
<td>( N_p:N_s )</td>
<td>1:5</td>
</tr>
<tr>
<td>Core Type</td>
<td></td>
<td>EE55</td>
</tr>
<tr>
<td>Magnetizing Inductance</td>
<td>( L_m )</td>
<td>23( \mu )H</td>
</tr>
<tr>
<td>Number of Turns in Primary</td>
<td>( N_p )</td>
<td>11</td>
</tr>
<tr>
<td>Peak magnetic flux density</td>
<td>( B_{max} )</td>
<td>0.23T</td>
</tr>
<tr>
<td>Air gap length</td>
<td>( l_g )</td>
<td>0.395cms</td>
</tr>
</tbody>
</table>
Using the proposed design scheme, a 250W grid connected single stage flyback inverter is designed for PV applications. The circuit parameters of the proposed flyback inverter are summarized in Table I.

3.3 Simulation Results

A 250W flyback inverter was designed and simulated in PSIM software with the component values obtained from the design procedure mentioned in previous sections and as listed in Table 3.1. The switch current $i_{sm}$, drain source voltage $v_{ds}$ and the gating for primary switch in Figure 3.14 illustrates the zero voltage switching for the primary switch. It can be seen that during the turn ON, the gating voltage for primary switch is applied only after $v_{ds}$ has become zero. The negative current seen in the Figure 3.14 is due to the reverse charging of magnetizing inductance of the transformer which aids the discharge of MOSFET’s output capacitance. A simple capacitive snubber is used in order to reduce the voltage spikes observed across the primary switch at turn OFF in the simulations.
Figure 3.14: Simulation waveforms of ZVS for primary switch

The bidirectional switches in secondary are also soft-switched at turn ON as shown in Figure 3.15. It can be observed that the voltage across the secondary switch is zero during their turn ON. The current $i_{sec}$ is shown inverted since it is negative with respect to the additional switch during the energy transfer to the output. Proper selection of the snubber capacitance helps in achieving soft switching for most of the switching cycles within the ac cycle.

Figure 3.15: Simulation Waveforms of ZVS for secondary switch
The variation of the primary switch voltage and the switch current over the AC cycle for an input DC voltage of 45V is shown in Figure 3.16. It can be observed from the simulation waveforms that the negative portion of the primary current variation is sinusoidal.

![Figure 3.16: Simulation waveform for primary switch voltage and current over ac cycle](image)

A pair of secondary MOSFET’s switch at line frequency while the other pair switches at the high frequency. The gating of the secondary MOSFETs along with the switch current for the positive half cycle is shown in Figure 3.17.

![Figure 3.17: Secondary switch gating and current waveforms in ac cycle](image)
The closed loop output current and the grid voltage waveforms are shown in Figure 3.18. It can be observed that the grid voltage and the output current are in phase.

![Figure 3.18: Output Current and Grid Voltage waveforms](image)

**3.4 Experimental Results**

On the basis of the foregoing design procedure, a 250W prototype is implemented to verify the performance of the proposed inverter. In order to implement the controller, the floating point TMS320F28335 eZdsp was employed. The grid voltage and the output ac current were sensed and conditioned before inputting to the 12-bit ADC of the processor. The synchronization of the grid was carried out digitally by extracting the phase from the grid voltage with the help of grid synchronization algorithms. The phase extracted from the algorithm aids in determining the reference signals required for the charging of the magnetizing winding in the reverse direction. Since the knowledge of transformer secondary current is required to generate the gating pulse for the bidirectional
switches, it is sensed with the help of a current transformer. The secondary current provides the information required to turn on the bidirectional switches. The ePWM modules in the DSP, generate the gating pulses for the various switches in accordance with the controller output.

![Diagram of ZVS for primary switch]

**Figure 3.19: ZVS for primary switch**

The experimental waveform exhibits soft-switching for both the primary and the additional secondary switch. Figure 3.19 shows the ZVS of primary switch as the drain-source voltage is forced to zero before the primary switch is gated. Figure 3.20 shows the ZVS of secondary switch where the secondary current is inverted with respect to the bidirectional switches. It can be seen that the drain-source voltage of the secondary MOSFET is zero, when the gating pulse is applied. As a result, the secondary MOSFET turns ON under ZVS.
The presence of negative current for the primary switch over the line cycle is shown in Figure 3.21. The envelope of the negative current follows a sinusoidal waveform as the negative peak of the primary current is controlled to optimize the reactive current.
The output current of the inverter and the grid voltage is shown in Figure 3.22. The experimental waveforms confirm with the theoretical and the simulation waveforms. Since the grid voltage and output current are in phase, the reactive power transfer to the grid is negligible.

![Waveform diagram](image)

**Figure 3.22: Experimental Waveforms of Grid Voltage and Output Current of Inverter**

The experiment was performed for different load conditions to obtain the efficiency curve shown in Figure 3.23. The efficiency of the overall system showed an anomalous behavior as compared to the theoretical calculations. The difference in efficiency in the experimental results was due to the parasitic effects of the interconnecting wires since no single printed circuit board was built for the overall system. Furthermore, optimization of the power MOSFETs and other circuit components could have accounted for the observed deviation in the system efficiency.
3.5 Summary

In this chapter, the working of the single stage grid-connected flyback inverter proposed for PV systems is explained. The detailed analysis of the individual modes of operation of the inverter is presented along with the theoretical waveforms. The effect of presence of reactive current in the proposed inverter is explored with the help of the steady-state expressions. The variation of the circuit parameters with respect to the critical quantities such as input voltage, output power etc. are studied to obtain optimum values for the circuit parameters. Since transformer is the crucial parameter in the design of any flyback inverter, a detailed explanation of selection of the core parameters using the $K_g$ method for the transformer design are provided. The proposed inverter is
implemented based on the design equations explained in this chapter. Simulation and the experimental results are provided to confirm the feasibility of the proposed inverter topology.
Chapter 4

Inverter Control and Grid Synchronization

4.1 Introduction

The main objective of the PV inverter is to supply active power to the utility grid. In order to inject a pure sinusoidal current to the grid, the inverter system should have a feedback loop that causes the output current to follow a sinusoidal reference current. The feedback system should be designed such that the grid current is accurately regulated and remains insensitive to disturbances in the input or the grid voltage. Further, the closed loop system should be stable for the entire operating range.

A dynamic model of the inverter system is required to design the feedback control for the system. Averaged switch modeling approach is a common technique used to obtain the large signal model of a converter. The idea in this approach is to take the average of the converter waveform to effectively predict the low-frequency behavior of the system while neglecting the high-frequency switching harmonics [44-47]. The switch network is replaced by basic circuit elements such as dependent or independent voltage or current source, loss-free resistor or power source based on the averaged terminal quantities [44, 48, 49]. The resulting equivalent circuit provides the time invariant averaged switch model of the converter. The small signal model for the inverter is derived by linearization of the large-signal averaged equivalent circuits. Generally, the
control-to-output transfer function obtained by solving the small signal equivalent circuit of conventional flyback topology has two poles. Moreover, the flyback topology is a non-minimum phase system due to the presence of right half plane zero in the control-to-output transfer function caused by the supply of energy during off-period of primary switch [44]. The inductor dynamics at low frequencies are insignificant since the pole due to magnetizing inductance of the transformer usually occurs at a frequency closer to the switching frequency. Thus the effect of magnetizing inductance is usually neglected to obtain a single pole transfer function for flyback converters.

The conventional averaging methods for dynamical characterization of the power converters were for fixed-frequency operation. In case of BCM mode of operation of flyback converters, it would not be possible to use conventional averaging methods due to variable frequency switching of the power MOSFETs [50]. In [46], the small signal modeling of a variable frequency flyback topology operating in BCM is derived by using a modified circuit averaging technique. Though, this model predicted the presence of right half plane zero, it only gave a first order control-to-output transfer function. Therefore, the predicted phase behavior of this model was found to be inaccurate at high frequencies.

In this chapter, the effect of the variable frequency operation on the dynamics of the proposed flyback topology is studied. The small signal modeling is based on the modified state-space averaging technique introduced in [51] where the effect of varying on-time constraints derived from the inductor current waveform is considered. The
presence of the negative current for the primary switch in the modified BCM operation complicates the modeling of the topology. Hence, the effect of the negative current is taken into consideration while finding the small signal model of the proposed topology. The effect of magnetizing inductance is considered while obtaining the transfer functions as it provides a better understanding of the dynamics of the whole system. Since the output current needs to follow the sinusoidal reference current, the effect of the output filter is observed while designing the feedback control. The control-to-output transfer function determined by taking into account the above mentioned criteria helps design a suitable compensator for a robust and stable control system. A step by step modeling of the proposed inverter is provided in the subsequent sections of this chapter. The grid synchronization scheme used for generating the sinusoidal current reference in phase with the grid voltage is presented in the latter half of the chapter.

4.2 Modeling of the Proposed Flyback Inverter

4.2.1 Modeling Principles for Direct On-Time Control

During a particular switching cycle, the grid side AC quantities are constant as the switching frequency is significantly higher than grid frequency. Hence, for a switching cycle, the circuit operation of a flyback inverter is equivalent to a DC-DC flyback converter. Since the state-space averaging is based on the averaging of the circuit waveforms over a switching cycle, the steps followed for modeling of the proposed flyback inverter is similar to a DC-DC flyback converter operating under BCM. The
A major change in the modeling process is the effect of the negative current in the magnetizing inductance on the switch current and voltage waveforms. Since a flyback converter is a transformer-isolated version of a buck-boost converter, modeling of a buck-boost converter with all the primary-side parameters of a flyback converter reflected to the secondary would yield the same results [44]. Figure 4.1 shows the equivalent flyback converter where the input parameters being the reflected quantities. This circuit represents the operation of the proposed inverter for a particular switching cycle.

\begin{figure}[h]
\centering
\includegraphics[width=\textwidth]{figure41.png}
\caption{Equivalent flyback converter with parameters reflected to secondary}
\end{figure}

\begin{align*}
  v_{in} &= NV_{dc} \\
  L &= N^2 L_m
\end{align*}

(4.1) \hspace{1cm} (4.2)

The solution of the system dynamics is obtained by using time-varying averaged values of inductor current \( \langle i_L \rangle \) and capacitor voltage \( \langle v_C \rangle \) as the state variables, the input
voltage \(v_{\text{in}}\) and output current sink \(i_O\) as input variables, and the output voltage \(v_O\) and input current \(i_{\text{in}}\) as the output variables. In case of a fixed frequency operation, the duty cycle is usually considered as the fundamental control variable. A constant time period results in equivalent dynamic properties of both duty cycle and the on-time of the primary switch. However, under variable frequency operation, on-time \(t_{\text{on}}\) has to be used as the control variable because of the different dynamical properties of duty ratio and on-time [51].

A peak current mode (PCM) control is implemented for the proposed inverter scheme. The control current for the PCM is the peak of inductor current \(i_{pk}\). In order to obtain the state-space equations for PCM, the relation between the on-time and peak inductor current is calculated. The inductor current waveform, shown in Figure 4.2, gives the relation between on-time and cycle time constraints for the variable frequency operation. In each switching cycle, the inductor current rises from its negative peak \(i_{npk}\) until it reaches a peak current equal to \(i_{pk}\).

\[
t_s = t_{\text{on}} + t_{\text{off}}
\]

\[
\langle i_L \rangle = \frac{\langle i_{pk} + i_{npk} \rangle}{2}
\]

\[
t_s = t_{\text{on}} + \frac{2L\langle i_L \rangle}{v_C}
\]

\[
i_{pk} = \langle i_L \rangle + \frac{m_1 t_{\text{on}}}{2}
\]
The relation between the on-time and the inductor peak currents needed for PCM can be obtained from the above expressions.

\[ t_{on} = \frac{2L(i_{pk} - \langle i_L \rangle)}{v_{in}} \]  
\[ t_{on} = \frac{2L(\langle i_L \rangle - i_{npk})}{v_{in}} \]

Figure 4.2: Magnetizing Inductor current waveform in modified BCM

In this approach of modeling, the time-varying averaged state variables and output variables are represented in terms of the circuit parameters. The derivative of the averaged inductor current is given by
\[
\frac{d\langle i_L \rangle}{dt} = \frac{t_{on}}{t_s} m_1 - \frac{t_{off}}{t_s} m_2
\]  
\hspace{2cm} (4.10)

Substituting the value of the slopes \( m_1 \) and \( m_2 \) for the corresponding sub-intervals, expression (4.10) can be expressed in terms of the state and output variables.

\[
\frac{d\langle i_L \rangle}{dt} = \frac{t_{on}}{t_s} \left( v_{in} + v_C \right) - \frac{v_C}{t_s L} \frac{v_C}{L}
\]  
\hspace{2cm} (4.11)

The derivative of the capacitor voltage is approximated based on the average charge delivered to the capacitor and removed from it in each switching cycle.

\[
\frac{dv_C}{dt} = \left( 1 - \frac{t_{on}}{t_s} \right) \frac{\langle i_L \rangle}{C_f} - \frac{v_C}{R_{ac} C_f} - \frac{i_o}{C_f}
\]  
\hspace{2cm} (4.12)

The average output voltage equals the capacitor voltage as can be seen from the Figure 4.1.

\[ v_o = v_C \]  
\hspace{2cm} (4.13)

The averaged input current for the converter is the average on-time inductor current.

\[ \langle i_{in} \rangle = \frac{t_{on}}{t_s} \langle i_L \rangle \]  
\hspace{2cm} (4.14)

The equations (4.11) - (4.14) form the non-linear state and output equations. In order to find the small signal model, these equations are linearized by using the partial derivatives [44]. To find the partial derivatives for the time-varying averaged quantities, each equation is differentiated with respect to the variables individually, keeping the other
variables constant. The small signal expression using partial derivative of a function \( f(x,y,z) \) is defined as:

\[
\hat{f}(x,y,z) = \left. \frac{\partial f(x,Y,Z)}{\partial x} \right|_{x=x} + \left. \frac{\partial f(X,y,Z)}{\partial y} \right|_{y=y} + \left. \frac{\partial f(X,y,z)}{\partial z} \right|_{z=z}
\]

(4.15)

Hence, the small signal expressions for the different time-varying averaged quantities using partial derivatives are given by

\[
\frac{d\hat{i}_L}{dt} = -\frac{D}{L} \hat{v}_C + \frac{D}{L} \hat{v}_{in} + \left( \hat{V}_{in} + \hat{V}_O \right) t_{on} - \frac{D}{L} \hat{V}_{in} \frac{T_S}{T_S}
\]

(4.16)

\[
\frac{d\hat{v}_C}{dt} = \frac{D}{L} \frac{\hat{i}_L}{T_C} t_{on} + \frac{1}{C_f} \hat{i}_O - \frac{D}{L} \frac{T_C}{T_S} \frac{I_L}{T_S}
\]

(4.17)

\[
\hat{v}_C = \hat{v}_O
\]

(4.18)

\[
\hat{v}_{in} = D \hat{v}_L \frac{T_C}{T_S} - \frac{DI_L}{T_S} \frac{T_C}{T_S}
\]

(4.19)

The parameters in uppercase are the steady state values of the state variables \((I_L, V_C)\), output voltage \((V_O)\), input voltage \((V_{in})\) and the duty ratio \((D)\). The steady state value for the switching period \((T_S)\) is given by

\[
T_S = \frac{2LI_L}{DV_{in}} = \frac{2LI_L}{DV_O}
\]

(4.20)

Since the proposed inverter is operating under variable frequency, the small signal variations of the switching period and the on-time also needs to be considered. The on-
time constraint can be considered in terms of both the positive and the negative peak inductor current.

\[
\hat{t}_{on} = \frac{2L}{V_{in}} \left( i_{pk} - \frac{2LI_{L}}{V_{C}} - \frac{T_{on} V_{in}}{2L} \right) 
\] (4.22)

\[
\hat{t}_{on} = \frac{2L}{V_{in}} \left( i_{npk} - \frac{T_{on} V_{in}}{2L} \right) 
\] (4.23)

The linearized forms of the on-time in (4.22) and (4.23) are with respect to the positive and negative peak inductor currents in case of the modified BCM operation. Since both positive and negative peak currents can independently determine the on-time, it is possible to develop control-to-output transfer functions for both cases individually. Using these constraints, the dynamics of a peak current mode controlled converter for both positive and negative peak currents can be obtained as explained in the subsequent sections.

4.2.2 Transfer Function Derivation

The state space representation of the state variables in terms of the positive peak current is determined by substituting the expressions (4.21) and (4.22) in the set of linear differential equations for the state and output variables. The new set of the expressions are given by
\[
\frac{d\langle i_L^\wedge \rangle}{dt} = -\frac{2DV_{in}}{LI_L}\langle i_L^\wedge \rangle + \frac{DV_{in}}{LI_L}i_{pk} \tag{4.24}
\]

\[
\frac{dv_C^\wedge}{dt} = \frac{D(1+2D)}{C_f}\langle i_L^\wedge \rangle - \left(\frac{1}{R_{ac}C_f} + \frac{DD'I_L}{V_oC_f}\right)v_C^\wedge - \frac{DD'I_L^\wedge}{V_inC_f}v_{in} - \frac{1}{C_f}i_o^\wedge - \frac{DD'i_{pk}^\wedge}{C_f} \tag{4.25}
\]

\[
v_o^\wedge = v_C \tag{4.26}
\]

\[
\langle i_{in}^\wedge \rangle = D(2D-1)\langle i_L^\wedge \rangle + \frac{DD'I_L}{V_o}v_C^\wedge - \frac{DD'I_L^\wedge}{V_in}v_{in} + DD'i_{pk}^\wedge \tag{4.27}
\]

The set of small signal state space equations can be represented using Matrix techniques for open loop. In order to obtain the transfer functions, the open loop system equations can be written in s-domain using the Laplace transforms.

\[
sX(s) = A_{1}X(s) + B_{1}U(s) + E_{1}I_{pk}(s) \tag{4.28}
\]

\[
Y(s) = C_{1}X(s) + D_{1}U(s) + P_{1}I_{pk}(s) \tag{4.29}
\]

The individual terms of the matrices, obtained from the small signal state space equations (4.24) - (4.27) are

\[
X(s) = \begin{pmatrix} I_L(s) \\ V_c(s) \end{pmatrix} \quad Y(s) = \begin{pmatrix} I_{in}(s) \\ V_o(s) \end{pmatrix} \quad U(s) = \begin{pmatrix} V_{in}(s) \\ I_o(s) \end{pmatrix} 
\]

\[
A_1 = \begin{pmatrix} -\frac{2DV_{in}}{LI_L} & 0 \\ \frac{D(1+2D)}{C_f} & -\left(\frac{1}{R_{ac}C_f} + \frac{DD'I_L}{V_oC_f}\right) \end{pmatrix} \quad B_1 = \begin{pmatrix} 0 & 0 \\ \frac{DD'I_L}{V_in} & -\frac{1}{C_f} \end{pmatrix}
\]
The solution of the two simultaneous matrix equations solved using simple matrix calculus is given by

\[ Y(s) = \left[ C_1 (sI - A_1)^{-1} B_1 + D_1 \right] U(s) + \left[ C_1 (sI - A_1)^{-1} E_1 + P_1 \right] I_{pk}(s) \quad (4.30) \]

The second term relates the output variable to the control variable. Hence, the control-to-output transfer function can be determined by calculation of the second matrix.

\[ T_p(s) = C_1 (sI - A_1)^{-1} E_1 + P_1 \quad (4.31) \]

Since the input variables are set to zero while calculating the control-to-output transfer function, first term in equation (4.30) becomes zero.

\[
\begin{bmatrix}
  \langle I_{in}(s) \rangle \\
  V_o(s)
\end{bmatrix} =
\begin{bmatrix}
  G_{p1} \\
  G_{p2}
\end{bmatrix} I_{pk}(s)
\quad (4.32)
\]

It can be observed that the term \( G_{p2} \) gives the ratio of output voltage and the positive peak of inductor current. The calculation of the above matrices yields the required transfer function for the converter.
It is evident from the working principle of the proposed inverter that the current on the secondary side of the transformer is allowed to become negative with the help of the bidirectional switches. Since the negative current on the secondary side is extracted from the grid, it is essential to control the amount of reactive current from the grid. With the change in the atmospheric conditions the amount of current injected to the utility grid varies. It is important that the magnitude of the negative peak current also vary relative to the variation in input power. The variation in the negative peak inductor current will act as a disturbance for the control-to-output transfer function as the average value of inductor current is dependent on negative peak of inductor current. Hence, the knowledge of the transfer function for negative peak of inductor current would provide a better understanding of the system.

The derivation procedure for the transfer function is similar to the method shown for the positive peak current. The set of small signal state space equations based on the constraint for negative peak of inductor current in equation (4.23) are given by

\[
\frac{d}{dt} \hat{i}_{L} = -\frac{DV_{in}}{LI_{L}} \hat{i}_{npk}
\]  

(4.34)
These small signal state space equations can be written in matrix form like the previous case.

\[
\begin{align*}
A_2 &= \begin{bmatrix} 0 & 0 \\ \frac{D}{C_f} & -\left(\frac{1}{R_{ac}C_f} + \frac{DDI_L}{V_oC_f}\right) \end{bmatrix}, & B_2 &= \begin{bmatrix} 0 & 0 \\ \frac{DDI_L}{V_{in}} & -\frac{1}{C_f} \end{bmatrix} \\
C_2 &= \begin{bmatrix} D & \frac{DDI_L}{V_o} \\ 0 & 1 \end{bmatrix}, & D_2 &= \begin{bmatrix} -\frac{DDI_L}{V_{in}} & 0 \\ 0 & 0 \end{bmatrix} \\
E_2 &= \begin{bmatrix} -\frac{DV_{in}}{LI_L} \\ \frac{DD}{C_f} \end{bmatrix}, & P_2 &= \begin{bmatrix} -DD \\ 0 \end{bmatrix}
\end{align*}
\]

The matrix equations are solved in order to find the transfer function with respect to the negative peak of inductor current. The input variables are set to zero to obtain \(G_{n2}\).

\[
\begin{bmatrix} \langle i_{in}(s) \rangle \\ V_o(s) \end{bmatrix} = \begin{bmatrix} G_{n1} \end{bmatrix} \begin{bmatrix} I_{npk}(s) \end{bmatrix}
\]
\[
G_{n2}(s) = \frac{V_o(s)}{I_{npk}(s)} = \frac{DD'}{C_f} \times \frac{s - \frac{V_m}{LI_L}}{s + \frac{DD' I_L}{V_o C_f} + \frac{1}{R_{dc} C_f}}
\] (4.39)

A look into the control-to-output transfer functions (4.33) and (4.39) gives an insight into the behavior of the system. Since, the effect of high frequency pole is considered, this method exhibit a second-order system behavior. The fact that the flyback converters have a non-minimum phase is confirmed by the presence of the right half plane zero at \(\omega_Z = \frac{V_L}{LI_L}\). The effect of the RHP and the high frequency pole is to result in a sharp decrease in phase of the system. The knowledge of the control-to-output transfer function provides the foundation for the controller design explained in next section.

### 4.3 Controller Design of the Proposed Inverter

Figure 4.3 shows the block diagram for the control of the output current of the inverter. It is obvious from the block diagram that a current mode control scheme is used to regulate the output current from the inverter. The grid voltage is sensed in the outer loop to generate the sinusoidal current reference for the inner current control loop. Since, the disturbances in grid voltage may result in amplitude, phase or frequency variation, PLL block in the outer loop is required to synchronize the inverter output current with the grid voltage. The PLL block generates the exact phase needed for the generation of proper sinusoidal grid current reference. The MPPT block extracts the maximum power
from the PV panel. Hence, the magnitude of the grid current reference is dependent on
the output of the MPPT block. When the atmospheric conditions are favorable, the grid
current reference would be higher due to increased power available from the PV panel.

![Inverter control block diagram](https://example.com/fig4_3)

**Figure 4.3: Inverter control block diagram**

In order to design the controller for a system, the information of the transfer
function of the uncompensated system is required. The block diagram for the inner
current control loop is shown in Figure 4.4. Equations (4.33) and (4.39) gave the control-to-output transfer functions only for the plant. The inverter is connected to the grid using
an inductive filter as shown.

The equation governing the dynamic of an inductive filter is given by

\[
L_f \frac{di_{\text{grid}}}{dt} = v_o - v_{\text{grid}}
\]  

(4.40)

where, \(v_o\) is the inverter output voltage and \(v_{\text{grid}}\) is the grid voltage.
Figure 4.4: Block diagram for the inner current control loop

The transfer function of the inverter output voltage to the grid current for the inductive filter is

\[ G_f(s) = \frac{I_{\text{grid}}(s)}{V_0(s)} = \frac{1}{sL_f} \]  

(4.41)

Hence, the loop transfer functions for the open loop or uncompensated system for the positive and negative peaks of inductor currents are given by equations (4.42) and (4.43) respectively.

\[
G_{p_{-\text{uncomp}}}(s) = \frac{I_{\text{grid}}(s)}{I_{pk}(s)} = k_{\text{sense}}DD' \frac{s - \frac{V_{in}}{L I_L}}{s + \frac{2DV_{in}}{L I_L}} \left( s + \frac{DD'I_L}{V_o C_f} + \frac{1}{R_{ac} C_f} \right) 
\]

(4.42)

\[
G_{n_{-\text{uncomp}}}(s) = \frac{I_{\text{grid}}(s)}{I_{npk}(s)} = k_{\text{sense}}DD' \frac{s - \frac{V_{in}}{L I_L}}{s^2 + \frac{DD'I_L}{V_o C_f} + \frac{1}{R_{ac} C_f}} 
\]

(4.43)

\(k_{\text{sense}}\) is the sensor gain.
The presence of the RHP zero and the high frequency poles in the control-to-output transfer function do not adversely affect the magnitude curve of the system but exhibits a sharp fall in the phase of the uncompensated system at higher frequencies as shown in Figure 4.5. This sharp fall in the phase causes limitations in the achievable bandwidth of the closed loop system. It can be observed that the phase of the uncompensated system drops to around -225 degrees around 2 kHz and continues to fall rapidly with the increase in frequency. Hence, the compensator should provide a large phase boost in order to achieve a reasonable phase margin at the gain cross-over frequency. Larger the bandwidth required, a higher phase boost would be needed.

Since a large phase boost is required for a stable closed loop system, a Type-III compensator is used for peak current mode control of the compensation of the proposed inverter. A Type-III compensator can provide a maximum phase boost of 180 degrees due to the presence of two zeroes in its transfer function. The compensator transfer function for the peak current mode control is given by

\[ H(s) = 0.1488 \frac{(0.00167s + 1)(0.001s + 1)}{s(0.00001s + 1)(0.00000667s + 1)} \]  \hspace{1cm} (4.44)

Bode plot for the Type III compensator is also shown in Figure 4.5. Bode plot of the compensated system for the peak current mode control for the positive peak of the inductor current is shown in Figure 4.6. A phase margin of 49 degrees is achieved with a reasonable bandwidth for the closed loop system.
Figure 4.5: Open Loop and Compensator Bode Diagrams for Positive Peak

Figure 4.6: Bode Diagram of Compensated System for Positive Peak
Bode plot of the uncompensated transfer function for the control of the negative peak of the inductor current is shown in Figure 4.7. The open loop transfer function of negative peak of inductor current to the grid current has pair of poles at origin and a pole at higher frequency. The phase drops rapidly at higher frequencies similar to case of positive peak of inductor current. Hence, a large phase boost is required to achieve sufficient phase margin at the gain crossover frequency. A second order compensator with two zeroes a decade below the required gain crossover frequency is used in order to provide the phase boost. High frequency poles would be needed to attenuate the gain at higher frequencies. Bode diagram of the compensator is shown in Figure 4.7.

\[
H(s) = 96.125 \frac{(0.001s + 1)(0.0002s + 1)}{(0.00002s + 1)(0.0000125s + 1)}
\]  

(4.45)

Figure 4.7 : Open Loop and Compensator Bode Diagrams for Negative Peak
Bode plot of the compensated system for the peak current mode control for the negative peak of the inductor current is shown in Figure 4.8. A phase margin of 45 degrees is achieved with a reasonable bandwidth for the closed loop system. Hence, the compensator designed for both the cases is stable due to the sufficient phase margin at the cross-over frequency. The bandwidth of the closed loop system is high enough to provide a fast noise filtering and disturbance rejection.

4.4 Experimental Frequency Response Results

The frequency response of the proposed scheme was observed using the Venable 350 Model Frequency Response Analyzer. Bode plot for the uncompensated system is
shown is Figure 4.9. The presence of right half plane zero and poles results in the drop in phase of the uncompensated system to -360°.

![Bode plot of the type III compensator used for the control of the proposed topology](image)

**Figure 4.9: Frequency Response of control-to-output of Open loop system**

Since a large phase boost is required, the type III compensator was used to achieve sufficient phase margin. The bode plot of the type III compensator used for the control of the proposed topology is shown in Figure 4.10. Bode plot for the compensated system for the control of the positive peak of primary current is shown in Figure 4.11. It is observed that the plots obtained from the experimental setup matches the bode plots obtained by plotting the frequency response from the transfer functions.
Figure 4.10: Frequency Response of Type III Compensator

Figure 4.11: Frequency Response of Compensated System
4.5 Grid Synchronization

The presence of the disturbances and harmonics in the grid voltage makes it unfeasible to use the grid voltage directly as the synchronization signal. Since these disturbances in the grid should not be reflected on the reference signals for the proper control of output current, different grid synchronization schemes need to be employed for grid-connected inverter systems [12]. Phase Locked Loop (PLL) is a nonlinear filter structure which can extract the phase information from the grid voltage for the grid synchronization. The task of the PLL is to provide unity power factor operation by synchronizing the inverter output current and the grid voltage. The PLL structure is responsible for monitoring and detecting the grid voltage parameters such as phase and amplitude. The output of the PLL block is used to generate a clean sinusoidal reference, for output current control loop, independent of the distortions in the utility voltage. [52, 53].

A simple and accurate grid synchronization technique based on the Adaptive Notch Filtering (ANF) for single phase systems introduced in [54, 55] is implemented for the proposed flyback inverter in this thesis. The block diagram for the digital form of the ANF technique using the Bilinear Transformation is shown in Figure 4.12. The input to the ANF module is the distorted sinusoidal grid voltage. It can be observed from the block diagram that \( v_\alpha \) and \( v_\beta \) are the fundamental component and its 90° phase shifted signals of the grid voltage respectively. The ANF scheme is composed of few multipliers, adders and integrators. The output of one of the integrators of ANF gives the angular
frequency \( \omega \) of the grid voltage. The initial condition for this integrator is set to the nominal grid frequency. Since the amplitude of the grid voltage can be calculated from \( v_\alpha \) and \( v_\beta \), the sine and cosine functions of the phase angle are easily obtained. The control parameters for the ANF scheme are \( \xi \) and \( \gamma \). The capability of the algorithm to track the variations in grid voltage is determined by the convergence factor \( \xi \) while the noise sensitivity of the filter is controlled by the parameter \( \gamma \).

![Figure 4.12: Digital Implementation of the ANF technique](image)

The difference equations for implementation of the ANF algorithm are obtained from the block diagram in Figure 4.12.
\[ x_1(n) = 2\zeta e(n)\omega(n) \]  \hspace{1cm} (4.46)

\[ x_2(n) = x_1(n) - \omega(n)v_\alpha(n) \]  \hspace{1cm} (4.47)

\[ v_\alpha(n) = v_\alpha(n-1) + \frac{T}{2}(x_2(n) + x_2(n-1)) \]  \hspace{1cm} (4.48)

\[ v_\beta(n) = \omega(n)\left(v_\beta(n-1) + \frac{T}{2}(v_\alpha(n) + v_\alpha(n-1))\right) \]  \hspace{1cm} (4.49)

\[ x_3(n) = -\gamma e(n)v_\beta(n) + \omega_{ref} \]  \hspace{1cm} (4.50)

\[ \omega(n) = \omega(n-1) + \frac{T}{2}(x_3(n) + x_3(n-1)) \]  \hspace{1cm} (4.51)

\( T \) is the sampling frequency which is chosen as 20 kHz. The simulations waveforms for the ANF are shown. Figure 4.13 and Figure 4.14 shows the response of the ANF method when there is a step variation in the frequency of the grid voltage from 60Hz to 65Hz and 65Hz to 60 Hz respectively. It can be observed that the ANF method tracks the variations in the frequency accurately.

![Figure 4.13: Response of ANF where frequency of grid voltage jumps from 60Hz to 65Hz](image-url)
Figure 4.14: Response of ANF where frequency of grid voltage jumps from 65Hz to 60Hz

The ANF method of grid synchronization was implemented using the TMSF28335 eZdsp to generate the sinusoidal reference signal for the grid current control loop.

4.6 Maximum Power Point Tracking for Proposed Inverter

It is evident from the PV cell model that the I-V characteristic of the panel is dependent on the irradiance level and the panel temperature [56]. The panel temperature is governed by the atmospheric conditions, efficiency of the heat exchange mechanism within the cells, ambient temperature etc.

Figure 4.15 : Typical PV characteristics for varying irradiance levels
A typical I-V and P-V characteristic of a PV panel for varying irradiation is shown in Figure 4.15. The points labeled as MPP on the plot are the operating point of the panel, where it generates the maximum power output for a particular irradiance level. The unpredictable nature of the atmospheric conditions make it essential to track the maximum power point continuously to extract maximum power from the PV panel for the particular set of operating conditions.

Perturb and Observe method is a widely used method to achieve MPPT because of the ease and the low cost associated with the implementation of the algorithm [57]. The idea of the algorithm is to track the output power of the panel based on a small perturbation in the PV voltage. If the output power of the PV panel increases when the voltage is perturbed from its operating point due to the changes in atmospheric conditions, then we need to continue to increase the PV voltage to reach the MPP. Under this circumstance, the operating point is on the left of the MPP point in Figure 4.15. Hence, the perturbation is PV voltage, $\Delta V_{pv}$, would be positive until it reaches the MPP point. While a decrease in output power on increase of the PV voltage, signifies a movement away from the MPP. As a result the direction of voltage perturbation must be reversed to move closer to the MPP. Since there is a continuous perturbation taking place in this algorithm, there may be oscillations about the MPP at steady state. In order to keep the oscillations to minimum, a smaller perturbation step size can be used.
The block diagram for the Perturb and Observe method is shown in Figure 4.16.

The PV voltage and input voltages are differentiated to obtain the variation in output power with respect to the variation in the PV voltage.

\[
\frac{dP}{dV_{pv}} = I_{pv} + V_{pv} \frac{dI_{pv}}{dV_{pv}}
\]  

Figure 4.16: Block Diagram for Perturb and Observe Algorithm

Depending on the sign of the above differential, the reference PV voltage is adjusted such that the maximum power is extracted from the PV panel. A simple Type II compensator is used in order to generate the magnitude of the grid current reference for the inner current loop. The transfer function for the controller is given by

\[
H(s) = 0.667 \frac{(0.2s+1)}{s(0.011s+1)}
\]  

The output of the controller is multiplied with the sinusoidal signal that is synchronized with the grid voltage to obtain the proper grid current reference.
4.7 Summary

This chapter dealt with the modeling and control of the proposed inverter for the implementation of the PV system. The chapter began with an introduction to the techniques usually employed for the modeling of the inverters. A detailed analysis for the modeling based on the variable frequency operation of the inverter was presented. With the help of the small signal state space equations for the dynamics of the system, the control-to-output transfer function was found. The bode diagram of the open loop transfer function of the system provided an insight of the uncompensated system. The necessary phase boost for the uncompensated system was provided with the help of Type III Compensator in order to achieve a stable closed loop system.

Since the grid synchronization is unavoidable for grid connected inverters, ANF method is implemented for generation of clean sinusoidal current reference. The digital implementation of the ANF method with the waveforms is presented. Tracking of the Maximum Power Point is achieved using the simple Perturb and Observe algorithm. Hence, this chapter gives the analysis required for the implementation of complete PV system.
Chapter 5

Conclusions

5.1 Summary

With the increase in energy prices and the adverse environmental effects of the non-renewable energy sources, the interest in improving power generation from solar energy has grown over the past few decades. In this research, first the role of power electronics technology in improving the solar power generation in the recent times was observed. Then the challenges associated with the design of the power converters to achieve efficient power conversion for PV modules were taken into consideration while designing the inverter. In addition, an extensive background study on the various existing inverter topologies for PV module was conducted. It was observed that most of the existing single-stage flyback based topologies for PV modules used auxiliary circuits to achieve soft-switching in order to minimize the switching losses in the power MOSFETs.

Thus the main purpose of this thesis was to propose a novel scheme to achieve ZVS for a single stage grid-connected flyback inverter without the addition of any auxiliary circuits. The soft-switching of the primary switch was made possible by allowing a current flow from the grid-side to the transformer which provides the negative current for soft-switching of primary switch. This concept was implemented by the inclusion of bidirectional switches on the secondary side of the transformer. The
bidirectional switches, not only helped to provide the negative current to achieve ZVS for primary switch, but were also turned ON under ZVS. Hence, the soft-switching of the additional secondary MOSFETs meant that the overall efficiency of the inverter was not affected by inclusion of the extra secondary switches. A detailed analysis and design of the circuit parameters were presented to achieve an efficient flyback inverter with minimal possible current and voltage stresses on the switches.

The control system for the proposed inverter was designed with the help of modified state space averaging technique and the corresponding small signal model of the inverter. Grid synchronization of the inverters was found to be an important criterion while designing the complete PV module system. Thus, a grid synchronization technique based on ANF algorithm was implemented to keep the inverter system tolerant to the disturbances in the grid voltages. The digital control scheme for the inverter and the ANF algorithm was executed on a TMS320F28335 processor. The simulation and experimental results are included in the thesis which verifies the concept for achieving soft-switching in the proposed inverter topology.

5.2 Contributions

The main contributions of this thesis are summarized below:

1. This thesis proposed a novel scheme of achieving ZVS in a grid-connected single-stage flyback inverter. Negative current required to turn ON the primary switch under ZVS, is obtained by allowing the flow of reactive current from the grid-side to the transformer for a short interval of time. In order to allow the negative
current flow, bi-directional switches are included on the secondary side of the transformer. The combination of a power MOSFET and diode in the conventional flyback inverters on the grid-side is replaced by a bi-directional switch to permit the current to conduct from the grid-side to the primary-side. Further, these grid-side MOSFET that replaced the diode is also turned ON under ZVS thereby minimizing the switching losses on the secondary side of the transformer as well.

2. The proposed flyback inverter operates in modified BCM in order to maximize the transferable output power. The modified BCM operation allows the charging of the transformer magnetizing inductance in the reverse direction of power flow to provide reactive current for soft-switching. A variable frequency scheme is implemented over the line cycle for the flyback inverter in order to optimize the amount of reactive current drawn from the grid-side to achieve ZVS for the entire operating range.

3. The presence of negative inductor current introduces some subtleties in the modeling of the proposed scheme. Since there is negative current in the primary MOSFETs, the effect of the negative peak of the inductor current should be considered. A modified state space averaging technique is presented to obtain the accurate small signal model of the system. This precise model can be used to design the controller for the closed-loop system.

4. A peak current mode control is proposed for the control of the grid current, which is able to effectively control the positive peak of inductor current as well as the
negative peak. This controller optimizes the amount of reactive current drawn from grid to ensure ZVS. The compensators designed for both the control loops satisfy the stability criteria. Thus, a stable and robust closed loop system for a wide operating range is presented in the thesis.

5.3 Future Scope

There can be further improvements in the proposed inverter scheme to attain better results. These are summarized below:

1. A simple first order inductive filter is used at the output for the proposed topology. Usually the size of inductor becomes large in order to achieve a reasonable attenuation of the inverter switching ripples. Hence, a third order LCL filter could be used to improve the quality of the output current and reduce the size of the inductors.

2. While developing the small signal model of the inverter, the effect of parasitics were considered to be negligible. A more accurate model can be obtained by taking into account the effect of parasitics such as the leakage inductance, equivalent series resistance (ESR) of the capacitors, etc.

3. The modeling of the inverter was obtained by linearizing the circuit variables. Such models are generally approximate models of the topology. The linear control theory is applied to these linearized models to design the compensator. In order achieve a better controller for the proposed inverter, non-linear control techniques for compensator design can be explored in the future.
References


