Introduction

LinkSwitch™-TN2 ICs combine a high-voltage (725 V) power MOSFET switch with an ON/OFF controller onto a monolithic IC. Jitter is added to the switching frequency to achieve lower EMI and the IC is fully fault protected. Auto-restart limits device and circuit dissipation during overload and output short-circuit while over temperature protection disables the internal power MOSFET during thermal faults. The high thermal shutdown threshold is ideal for applications where the ambient temperature is high while the large hysteresis protects the PCB and surrounding components from high average temperatures.

The LinkSwitch-TN2 family is designed for any application where a non-isolated supply is required such as appliances (coffee machines, rice cookers, dishwashers, microwave ovens etc.), night lights, emergency exit signs and LED drivers. LinkSwitch-TN2 ICs can be configured in all common topologies to give a line or neutral referenced output and an inverted or non-inverted output voltage – ideal for applications using TRIACs for AC load control. Using a switching power supply rather than a passive dropper (capacitive or resistive) gives a number of advantages, some of which are listed below.

- Universal input – the same power supply/product can be used worldwide
- High power density – smaller size, typically no X class capacitance needed for most designs
- High efficiency – full load efficiencies >75% typical for 12 V output, 120 mA
- Excellent line and load regulation
- High efficiency at light load – ON/OFF control maintains high efficiency even at light load
- Extremely energy efficient – input power <30 mW at no-load
- Entirely manufacturable with SMD components
- Fully fault protected (overload, short-circuit and thermal faults)
- Scalable – LinkSwitch-TN2 family allows the same basic design to be used from <50 mA to 360 mA.

Figure 1. Basic Configuration Using LinkSwitch-TN2 in a Buck Converter, 1(a) and Buck-Boost Converter, 1(b).
Scope

This application note is for designing a non-isolated power supply using the LinkSwitch-TN2 family of devices. This document describes the design procedure for buck and buck-boost converters using the LinkSwitch-TN2 family of integrated off-line switchers. The objective of this document is to provide power supply engineers with guidelines in order to enable them to quickly build efficient and low-cost buck or buck-boost converter based power supplies using low-cost off-the-shelf inductors. Complete design equations are provided for the selection of the converter’s key components. Since the power MOSFET and controller are integrated into a single IC, the design process is greatly simplified, the circuit configuration has few parts and no transformer is required. Therefore a quick start section is provided that allows off-the-shelf components to be selected for typical output voltages and currents. To simplify the task this application note refers directly to the PIXls design spreadsheet that is part of the PI Expert™ design software suite. The basic configuration used in LinkSwitch-TN2 power supplies is shown in Figure 1, which also serves as the reference circuit for component identifications used in the description throughout this application note.

In addition to this application note, the reader may also find the LinkSwitch-TN2 Reference Design Kit (RDK) containing an engineering prototype board, engineering report and device samples useful as an example of a working power supply. Further details on downloading PI Expert, obtaining a RDK and updates to this document can be found at www.power.com.

Quick Start

Readers wanting to start immediately can use the following information to quickly select the components for a new design, using Figure 1 and Tables 1 and 2 as references.

1. For AC input designs select the input stage (Table 10).
2. Select the topology (Tables 1 and 2). If better than ±5% output regulation is required, then use optocoupler feedback with suitable reference such as TL431 or a precision Zener is required.
3. Select the LinkSwitch-TN2 device, L, R FB or VZ, RBIAS, CFB, RZ and the reverse recovery time for DFW (Table 4: Buck, Table 5: Buck-Boost).
4. Select freewheeling diode to meet tRR determined in Step 3 (Table 3).
5. For direct feedback designs, if the minimum load <3 mA then calculate RPL = VO/3 mA.
6. Select C0 as 100 μF, 1.25 × VO, low ESR type.
7. Construct prototype and verify design.

<table>
<thead>
<tr>
<th>Topology</th>
<th>Basic Circuit Schematic</th>
<th>Key Features</th>
</tr>
</thead>
<tbody>
<tr>
<td>High-Side</td>
<td></td>
<td>1. Output referenced to input.</td>
</tr>
<tr>
<td>Buck – Direct</td>
<td><img src="PI-8199-120516" alt="High-Side Buck Direct Feedback Circuit" /></td>
<td>2. Positive output (V0) with respect to -VIN.</td>
</tr>
<tr>
<td>Feedback</td>
<td></td>
<td>3. Step down : V0 &lt; VIN.</td>
</tr>
<tr>
<td></td>
<td></td>
<td>4. Low cost direct feedback (±5% typ.)</td>
</tr>
<tr>
<td></td>
<td></td>
<td>5. Requires an output load to maintain regulation (Note 2).</td>
</tr>
<tr>
<td>High-Side</td>
<td><img src="PI-8200-120516" alt="High-Side Buck-Boost Direct Feedback Circuit" /></td>
<td>1. Output referenced to input.</td>
</tr>
<tr>
<td>Buck-Boost – Direct Feedback</td>
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<td>2. Positive output (V0) with respect to -VIN.</td>
</tr>
<tr>
<td></td>
<td></td>
<td>3. Step down : V0 &lt; VIN or V0 &gt; VIN.</td>
</tr>
<tr>
<td></td>
<td></td>
<td>4. Low cost direct feedback (±5% typ.).</td>
</tr>
<tr>
<td></td>
<td></td>
<td>5. Fail-safe – output is not subjected to input voltage if the internal MOSFET fails.</td>
</tr>
<tr>
<td></td>
<td></td>
<td>6. Requires an output load to maintain regulation (Note 2).</td>
</tr>
</tbody>
</table>

Table 1. LinkSwitch-TN2 Circuit Configurations using Directly Sensed Feedback.

Notes:
1. Low Cost, directly sensed feedback typically achieves overall regulation tolerance of ±5% with 3 mA pre-load for 12 V design
2. To ensure output regulation, a pre-load may be required to maintain a minimum load current of 3 mA (buck and buck-boost only).
3. Boost topology (step up) is also possible but not shown.
### Topology

<table>
<thead>
<tr>
<th>Basic Circuit Schematic</th>
<th>Key Features</th>
</tr>
</thead>
<tbody>
<tr>
<td><strong>High-Side Buck – Optocoupler Feedback</strong>&lt;br&gt;<img src="PI-784a-011017" alt="Circuit Diagram" />&lt;br&gt;1. Output referenced to input&lt;br&gt;2. Positive output ($V_O$) with respect to $-V_{IN}$&lt;br&gt;3. Step down: $V_O &lt; V_{IN}$&lt;br&gt;4. Optocoupler feedback&lt;br&gt;  - Accuracy only limited by reference choice&lt;br&gt;  - Low cost non-safety rated optocoupler&lt;br&gt;  - No pre-load required&lt;br&gt;5. Minimum no-load consumption</td>
<td></td>
</tr>
<tr>
<td><strong>Low-Side Buck – Optocoupler Feedback</strong>&lt;br&gt;<img src="PI-784s-011017" alt="Circuit Diagram" />&lt;br&gt;1. Output referenced to input&lt;br&gt;2. Negative output ($V_O$) with respect to $+V_{IN}$&lt;br&gt;3. Step down: $V_O &lt; V_{IN}$&lt;br&gt;4. Optocoupler feedback&lt;br&gt;  - Accuracy only limited by reference choice&lt;br&gt;  - Low cost non-safety rated optocoupler&lt;br&gt;  - No pre-load required</td>
<td></td>
</tr>
<tr>
<td><strong>Low-Side Buck-Boost – Optocoupler Feedback</strong>&lt;br&gt;<img src="PI-784b-011017" alt="Circuit Diagram" />&lt;br&gt;1. Output referenced to input&lt;br&gt;2. Positive output ($V_O$) with respect to $+V_{IN}$&lt;br&gt;3. Step up/down: $V_O &gt; V_{IN}$ or $V_O &lt; V_{IN}$&lt;br&gt;4. Optocoupler feedback&lt;br&gt;  - Accuracy only limited by reference choice&lt;br&gt;  - Low cost non-safety rated optocoupler&lt;br&gt;  - No pre-load required&lt;br&gt;5. Fail-safe – output is not subjected to input voltage if the internal power MOSFET fails&lt;br&gt;6. Minimum no-load consumption</td>
<td></td>
</tr>
</tbody>
</table>

**Table 2.** LinkSwitch-TN2 Circuit Configurations using Optocoupler Feedback.<br>Note:<br>1. Regulation of optocoupler feedback only limited by accuracy of reference (Zener or IC).<br>2. Optocoupler does not need to be safety approved.<br>3. Reference bias current provides minimum load. The value of $R_z$ is determined by Zener test current or reference IC bias current, typically $470 \Omega$ to $2 \kOmega$, $1/8$ W, 5%.<br>4. Boost topology (step-up) is also possible but not shown.<br>5. Optocoupler feedback provides lowest no-load consumption.<br>

<table>
<thead>
<tr>
<th>Part Number</th>
<th>$V_{RRM}$ (V)</th>
<th>$I_{f}$ (A)</th>
<th>$t_{RR}$ (ns)</th>
<th>Package</th>
<th>Manufacturer</th>
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<td>50</td>
<td>Leaded</td>
<td>Vishay</td>
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<tr>
<td>UF4005</td>
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<td>1</td>
<td>75</td>
<td>Leaded</td>
<td>Vishay</td>
</tr>
<tr>
<td>BYV26C</td>
<td>600</td>
<td>1</td>
<td>30</td>
<td>Leaded</td>
<td>Vishay/Philips</td>
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<tr>
<td>FE1A</td>
<td>600</td>
<td>1</td>
<td>35</td>
<td>Leaded</td>
<td>Vishay</td>
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<tr>
<td>STTA10 6</td>
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<td>1</td>
<td>20</td>
<td>Leaded</td>
<td>ST Microelectronics</td>
</tr>
<tr>
<td>STTA10 6U</td>
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<td>SMD</td>
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<td>US1J</td>
<td>600</td>
<td>1</td>
<td>75</td>
<td>SMD</td>
<td>Vishay</td>
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</table>

**Table 3.** List of Ultrafast Diodes Suitable for use as the Freewheeling Diode.
## Table 4. Components Quick Select for Buck Converters.

*Select nearest standard or combination of standard values.*

<table>
<thead>
<tr>
<th>$V_{OUT}$</th>
<th>$I_{OUT(MAX)}$</th>
<th>Inductor</th>
<th>$\mu H$</th>
<th>$I_{RMS(MAX)}$</th>
<th>Tokin</th>
<th>Colicraft</th>
<th>LNK320X</th>
<th>Mode</th>
<th>$Diode \quad t_{PR}$</th>
<th>$R \times FB$</th>
<th>$V_z$</th>
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<tbody>
<tr>
<td>≤63</td>
<td>80</td>
<td>2000</td>
<td>122</td>
<td>2000</td>
<td>84</td>
<td>–</td>
<td>–</td>
<td>–</td>
<td>–</td>
<td>–</td>
<td>–</td>
</tr>
<tr>
<td>120</td>
<td>160</td>
<td>870</td>
<td>152</td>
<td>870</td>
<td>167</td>
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<td>–</td>
<td>–</td>
<td>–</td>
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<tr>
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<td>225</td>
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<td>327</td>
<td>SBC4-681-211</td>
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<td>–</td>
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<td>≤ 35 ns</td>
<td>3.48 k</td>
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<td>MDCM</td>
<td>≤ 75 ns</td>
<td>≤ 35 ns</td>
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<td>80</td>
<td>2000</td>
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<td>1500</td>
<td>167</td>
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<td>870</td>
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<td>870</td>
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<td>680</td>
<td>336</td>
<td>1200</td>
<td>364</td>
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<td>LNK3206</td>
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<td>≤ 75 ns</td>
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<tr>
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<td>80</td>
<td>3000</td>
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<td>223</td>
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<td>LNK3206</td>
<td>MDCM</td>
<td>≤ 75 ns</td>
</tr>
<tr>
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<td>3600</td>
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</tr>
<tr>
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<td>750</td>
<td>282</td>
<td>2000</td>
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<td>–</td>
<td>–</td>
<td>–</td>
<td>–</td>
<td>–</td>
</tr>
</tbody>
</table>

Other Standard Components

- $R_{bias}$: 2.49 kΩ, 1%, 1/8 W
- $C_{M}: 0.1 \mu F, 50 V$ Ceramic
- $C_{F}: 10 \mu F, 1.25 \times V_o$
- $D_{DF}^*: 1N4005GP$
- $R_z: 470 \Omega$ to 2 kΩ, 1/8 W, 5%

Table 4. Components Quick Select for Buck Converters. *Select nearest standard or combination of standard values.*
<table>
<thead>
<tr>
<th>$V_{OUT}$</th>
<th>$I_{OUT(MAX)}$</th>
<th>Inductor</th>
<th>LNK320X</th>
<th>Mode</th>
<th>Diode $t_{pr}$</th>
<th>$R \times FB$</th>
<th>$V_z$</th>
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<tr>
<td>5</td>
<td>≤65 80</td>
<td>1200 70&lt;br&gt;1500 80</td>
<td>SBC3-152-251&lt;br&gt;SBC2-681-211</td>
<td>RFB0807-122&lt;br&gt;RFB0807-152</td>
<td>MDCM&lt;br&gt;CCM</td>
<td>≤ 75 ns&lt;br&gt;≤ 35 ns</td>
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<td>RFB0807-681&lt;br&gt;RFB0807-681</td>
<td>MDCM&lt;br&gt;CCM</td>
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<td></td>
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<tr>
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<td>≤ 75 ns&lt;br&gt;≤ 35 ns</td>
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<td>≤ 75 ns&lt;br&gt;≤ 35 ns</td>
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<td>SBC4-152-251&lt;br&gt;SBC4-222-211</td>
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<td>MDCM&lt;br&gt;CCM&lt;br&gt;CCM</td>
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<tr>
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</tr>
<tr>
<td>220 360</td>
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</tr>
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<td>680 180&lt;br&gt;1500 220&lt;br&gt;1800 210</td>
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<td>RFB0807-332&lt;br&gt;RFB0807-682</td>
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<td>50 120 160</td>
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<td>RFB0807-681&lt;br&gt;RFB0810-222&lt;br&gt;RFB0810-332</td>
<td>MDCM&lt;br&gt;CCM&lt;br&gt;CCM</td>
<td>≤ 75 ns&lt;br&gt;≤ 75 ns&lt;br&gt;≤ 35 ns</td>
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<td>175 225</td>
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<td>RFB1010-182&lt;br&gt;RFB1010-182</td>
<td>MDCM&lt;br&gt;CCM</td>
<td>≤ 75 ns&lt;br&gt;≤ 35 ns</td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

Other Standard Components
- $R_{bias}$: 2.49 kΩ, 1%, 1/8 W
- $C_{BP}$: 0.1 μF, 50 V Ceramic
- $C_{FB}$: 10 μF, 1.25 × $V_o$
- $D_{FB}$: 1N4005GP
- $R_Z$: 470 Ω to 2 kΩ, 1/8 W, 5%

Table 5. Components Quick Select for Buck-Boost Converters. *Select nearest standard or combination of standard values. The inductor values indicated in the table above are conservative estimates. In some designs it may be possible to reduce the inductance value further based on evaluation results.
LinkSwitch-TN2 Circuit Design

LinkSwitch-TN2 Operation

The basic circuit configuration for a buck converter using a LinkSwitch-TN2 IC is shown in Figure 1(a). To regulate the output, an ON/OFF control scheme is used as illustrated in Table 6. As the decision to switch is made on a cycle-by-cycle basis, the resultant power supply has extremely good transient response and removes the need for control loop compensation components. If no feedback is received for 50 ms, then the supply enters auto-restart mode of operation and switching is inhibited for a period of time to limit power dissipation.

Reference Schematic And Key

<table>
<thead>
<tr>
<th>Reference Schematic And Key</th>
<th>I_d</th>
</tr>
</thead>
<tbody>
<tr>
<td>Is I_FB &gt;49 mA?</td>
<td>No</td>
</tr>
</tbody>
</table>

At the beginning of each cycle, the FEEDBACK (FB) pin is sampled.
- If I_FB < 49 µA then next cycle occurs
- If I_FB > 49 µA then next switching cycle is skipped

Normal Operation

High load – few cycles skipped
Low load – many cycles skipped

Auto-Restart

If no feedback (I_FB < 49 µA) for > t_AR(ON) (50 ms), then output switching is disabled equal to the auto-restart off-time. The first time a fault is asserted the off-time is 150 ms (t_AR(OFF) first off period). If the fault condition persists, subsequent off-times are 1500 ms long (t_AR(OFF) subsequent periods).

Table 6. LinkSwitch-TN2 Operation.
To allow direct sensing of the output voltage without the need for a reference (Zener diode or reference IC), the FEEDBACK pin voltage is tightly tolerated over the entire operating temperature range. For example, this allows a 12 V design with an overall output tolerance of ±5%. For higher performance, an optocoupler can be used with a reference as shown in Table 2. Since the optocoupler just provides level shifting, it does not need to be safety rated or approved. The use of an optocoupler also allows flexibility in the location of the device, for example it allows a buck converter configuration with the LinkSwitch-TN2 IC in the low-side return rail, reducing EMI as the SOURCE pins and connected components are no longer part of the switching node.

Selecting the Topology
If possible, use the buck topology. The buck topology maximizes the available output power from a given LinkSwitch-TN2 IC and inductor value. Also, the voltage stress on the power switch and freewheeling diode and the average current through the output inductor are slightly lower in the buck topology as compared to the buck-boost topology.

Selecting the Operating Mode — MDCM and CCM Operation
At the start of a design, select between mostly discontinuous conduction mode (MDCM) and continuous conduction mode (CCM) as this decides the selection of the LinkSwitch-TN2 device, freewheeling diode and inductor. For maximum output current select CCM, for all other cases MDCM is recommended. Overall, select the operating mode and components to give the lowest overall solution cost. Table 7 summarizes the trade-offs between the two operating modes.

Additional differences between CCM and MDCM include better transient response for DCM and lower output ripple (for same capacitor ESR) for CCM. However these differences, at the low output currents of LinkSwitch-TN2 applications, are normally not significant.

The conduction mode CCM or MDCM of a buck or buck-boost converter primarily depends on input voltage, output voltage, output current and device current limit. The input voltage, output voltage and output current are fixed design parameter; therefore the LinkSwitch-TN2 current limit is the only design parameter that sets the conduction mode.

The phrase “mostly discontinuous” is used as with On/Off control, since a few switching cycles may exhibit continuous inductor current, the majority of the switching cycles will be in the discontinuous conduction mode. A design can be made fully discontinuous but that will limit the available output current, making the design less cost effective.

<table>
<thead>
<tr>
<th>Operating Mode</th>
<th>MDCM</th>
<th>CCM</th>
</tr>
</thead>
<tbody>
<tr>
<td><strong>Operating Description</strong></td>
<td>Inductor current falls to zero during $t_{off}$ boarder-line between MDCM and CCM when $t_{idle} = 0$.</td>
<td>Current flows continuously in the inductor for the entire duration of a switching cycle.</td>
</tr>
<tr>
<td><strong>Inductor</strong></td>
<td>Lower Cost</td>
<td>Higher Cost</td>
</tr>
<tr>
<td></td>
<td>Lower value, smaller size.</td>
<td>Higher value, larger size.</td>
</tr>
<tr>
<td><strong>Freewheeling Diode</strong></td>
<td>Lower Cost</td>
<td>Higher Cost</td>
</tr>
<tr>
<td></td>
<td>75 ns ultrafast reverse recovery type ($\leq 35$ ns for ambient $&gt; 70$ °C).</td>
<td>35 ns ultrafast recovery type required.</td>
</tr>
<tr>
<td><strong>LinkSwitch-TN2</strong></td>
<td>Potentially Higher IC Cost</td>
<td>Potentially Lowest IC Cost</td>
</tr>
<tr>
<td></td>
<td>May require larger device to deliver required output current – depends on required output current.</td>
<td>May allow smaller device to deliver required output current, depends on required output current.</td>
</tr>
<tr>
<td><strong>Efficiency</strong></td>
<td>Higher Efficiency</td>
<td>Lower Efficiency</td>
</tr>
<tr>
<td></td>
<td>Lower switching losses.</td>
<td>Higher switching losses.</td>
</tr>
<tr>
<td><strong>Overall</strong></td>
<td></td>
<td>Typically Higher Cost</td>
</tr>
</tbody>
</table>

Table 7. Comparison of Mostly Discontinuous Conduction (MDCM) and Continuous Conduction (CCM) Modes of Operation.
Step-by-Step Design Procedure

Step 1 – Enter Application Variables $V_{AC_{MIN}}$, $V_{AC_{MAX}}$, $f_L$, $V_{OUT}$, $I_{OUT}$, $\eta$, $C_{IN}$

Input Voltage

Determine the input voltage range from Table 8.

<table>
<thead>
<tr>
<th>Nominal Input Voltage (VAC)</th>
<th>$V_{AC_{MIN}}$</th>
<th>$V_{AC_{MAX}}$</th>
</tr>
</thead>
<tbody>
<tr>
<td>100/115</td>
<td>85</td>
<td>132</td>
</tr>
<tr>
<td>230</td>
<td>195</td>
<td>265</td>
</tr>
<tr>
<td>Universal</td>
<td>85</td>
<td>265</td>
</tr>
</tbody>
</table>

Table 8. Standard Worldwide Input Line Voltage Ranges.

Line Frequency, $f_L$

50 Hz for universal or single 100 VAC, 60 Hz for single 115 VAC input. 50 Hz for single 230 VAC input. These values represent typical line frequencies rather than minimum. For most applications this gives adequate overall design margin. For absolute worst case or based on the product specification reduce these numbers by 6% (47 Hz or 56 Hz). For half-wave rectification use $f_L/2$.

Nominal Output Voltage, $V_{OUT}$ (V)

Enter the nominal output voltage of the main output during the continuous load condition. Generally the main output is the output from which feedback is derived.

Output Power, $I_{OUT}$ (A)

Enter the output current of the power supply. If the power supply is a multiple output power supply, enter the sum total equivalent current of all the outputs, sum total power divide the main output voltage.

In multiple output designs the output power of the main output (typically the output from which feedback is taken) should be increased such that the maximum continuous output power as applicable matches the sum of the output power from all the outputs in the design. The individual output voltages and currents should then be entered at the bottom of the spreadsheet.

Power Supply Efficiency, $\eta$

Enter the estimated efficiency of the complete power supply measured at the output terminals under peak load conditions and worst-case line (generally lowest input voltage). Start with a value of 0.7 for a 12 V output, 0.55 for a 5 V output if no better reference data available, typical for a design where the majority of the output power is drawn from an output voltage of 12 V or greater. Once a prototype has been constructed then the measured efficiency should be entered.

Total Input Capacitance, $C_{IN}$ ($\mu$F)

Enter total input capacitance using Table 9 for guidance.

The capacitance is used to calculate the minimum and maximum DC voltage across the bulk capacitor and should be selected to keep the minimum DC input voltage, $V_{MIN}$ > 70 V.

<table>
<thead>
<tr>
<th>AC Input Voltage (VAC)</th>
<th>Total Input Capacitance per Watt Output Power $\mu$F/W</th>
</tr>
</thead>
<tbody>
<tr>
<td>Full-Wave Rectification</td>
<td>Half-Wave Rectification</td>
</tr>
<tr>
<td>100/115</td>
<td>2</td>
</tr>
<tr>
<td>230</td>
<td>1</td>
</tr>
<tr>
<td>85-265</td>
<td>2</td>
</tr>
</tbody>
</table>

Table 9. Suggested Total Input Capacitance Values for Different Input Voltage Ranges.
Step 2 – Determine AC Input Stage

The input stage comprises fusible resistor(s), input rectification diodes and line filter network. Flameproof fusible resistors are recommended to be chosen and depending on the differential line input surge requirements, a wire-wound type may be required. The fusible resistor(s) provides fuse safety, inrush current limiting and differential mode noise attenuation. The EMI performance of half-wave rectified designs is improved by adding a second diode in the lower return rail. This provides EMI gating (EMI currents only flow when the diode is conducting) and also doubles differential surge withstand as the surge voltage is shared across two diodes. Table 10 shows the recommended input stage based on output power while Table 9 shows how to adjust the input capacitance for other input voltage ranges.

Step 3 – Determine Minimum and Maximum DC Input Voltages \( V_{\text{MIN}} \) and \( V_{\text{MAX}} \) Based on AC Input Voltage

Calculate \( V_{\text{MAX}} \) as:

\[
V_{\text{MAX}} = \sqrt{2} \times V_{\text{ACMAX}}
\]

Assuming that the value of input fusible resistor is small, the voltage drop across it can be ignored. Derive minimum input voltage \( V_{\text{MIN}} \):

Half bridge rectifier:

\[
V_{\text{MIN}} = \sqrt{(2 \times V_{\text{ACMIN}}^2) - \frac{2 \times P_o}{\eta} \left( \frac{1}{f_L} - t_c \right)}
\]

Full bridge rectifier:

\[
V_{\text{MIN}} = \sqrt{(2 \times V_{\text{ACMIN}}^2) - \frac{2 \times P_o}{\eta} \left( \frac{1}{2 \times f_L} - t_c \right)}
\]

Step 4 – Select LinkSwitch-TN2 Device Based on Output Current and Current Limit

Decide on the operating mode – refer to Table 7.

For MDCM operation, the output current \( I_o \) should be less than or equal to half the value of the minimum current limit of the chosen device from the data sheet.

\[
I_{\text{LIMIT,MIN}} > 2 \times I_o
\]

For CCM operation, the device should be chosen such that the output current \( I_o \) is more than 50%, but less than 80% of the minimum current limit \( I_{\text{LIMIT,MIN}} \):

\[
0.5 \times I_{\text{LIMIT,MIN}} < I_o < 0.8 \times I_{\text{LIMIT,MIN}}
\]

Please see the data sheet for LinkSwitch-TN2 current limit values.

A typical LinkSwitch-TN2 part can be programmed to operate in one of the two current limits. The "RED" or reduced current limit enables operation at a reduced current limit and is recommended when the part is to be used at a current level considerably lower than the rated output current. A "STD" or standard current limit will be selected in most applications to optimize on BP capacitor cost.

Use of a 0.1 \( \mu F \) capacitor results in the standard current limit value. Use of a 1 \( \mu F \) capacitor results in the current limit being reduced, allowing design with lowest cost surface mount buck chokes.
**Step 5 – Select the Output Inductor**

Choose any standard off-the-shelf inductor that meets the design requirements. As shown in the figure below, a “drum” or “dog bone” “I” core inductor is recommended with a single ferrite element due to its low cost and very low audible noise properties. However, the inductor should be selected as varnished type in order to get low audible noise.

Tables 4 and 5 provide inductor values and RMS current ratings for common output voltages and currents based on the calculations in the design spreadsheet. Select the next nearest highest voltage and/or current above the required output specification. Alternatively, the PIXi spreadsheet tool in the PI Expert software design suite or Appendix B can be used to calculate the exact inductor value (Eq. C13) and RMS current rating (Eq. C29). It is recommended that the value of inductor chosen should be closer to \( L_{\text{typ}} \) rather than \( 1.5 \times L_{\text{typ}} \) due to lower DC resistance and higher RMS rating. The lower limit of 680 \( \mu \)H limits the maximum \( \frac{dI}{dt} \) to prevent very high peak current values. Tables 4 and 5 provide reference part numbers for standard inductors from two suppliers.

\[
680 \mu H < L < 1.5 \times L_{\text{typ}}
\]

For LinkSwitch-TN2 designs, the mode of operation is not dependent on the inductor value. The mode of operation is a function of load current and current limit of the chosen device. The inductor value merely sets the average switching frequency. Figure 4 shows a typical standard inductor manufacturer’s data sheet. The value of off-the-shelf “drum core / dog bone / I core” inductors will drop up to 20% in value as the current increases. The constant \( K_{\text{TOL}} \) in equation (C14) and the design spreadsheet adjusts for both this drop and the initial inductance value tolerance. For example if a 680 \( \mu \)H, 360 mA inductor is required, referring to Figure 4, the tolerance is 10% and an estimated 9.5% for the reduction in inductance at the operating current (approximately \( 0.36/0.38 \times 10 \)). Therefore the value of \( K_{\text{TOL}} = 0.195 \) (19.5%). If no data is available, assume a \( K_{\text{TOL}} \) of 0.15 (15%).

Not all the energy stored in the inductor is delivered to the load, due to losses after the LinkSwitch-TN2 device, the inductor (resistance of winding and core losses), the freewheeling diode, feedback circuit, output capacitor loss and preload. This will limit the maximum power delivering capability and thus reduce the maximum output current. The minimum inductance must compensate for these losses in order to deliver specified full load power. To compensate for this, a loss factor \( K_{\text{LOSS}} \) is used. This has a recommended value of between 50% and 66% of the total supply losses as given by Equation 6. For example, a design with an overall efficiency \( (\eta) \) of 0.75 would have a \( K_{\text{LOSS}} \) value of between 0.875 and 0.833.

\[
K_{\text{LOSS}} = 1 - \left( \frac{1 - \eta}{2} \right) \cdot \left( \frac{2(1 - \eta)}{3} \right)
\]
Step 6 – Select Freewheeling Diode

For MDCM operation at $t_{aeb} \leq 70 \, ^\circ C$, select an ultrafast diode with $t_{rr} \leq 75 \, ns$. At $t_{aeb} > 70 \, ^\circ C$, $t_{rr} \leq 35 \, ns$. For CCM operation, select an ultrafast diode with $t_{rr} \leq 35 \, ns$. Allowing 25% design margin for the freewheeling diode,

$$V_{FV} > 1.25 \times V_{\text{MAX}}$$  \hspace{1cm} (7)

The diode must be able to conduct the full load current. Thus:

$$I_F > 1.25 \times I_O$$  \hspace{1cm} (8)

Table 3 lists common freewheeling diode choices.

Step 7 – Select Output Capacitor

The output capacitor should be chosen based on the output voltage ripple requirement. Typically the output voltage ripple is dominated by the capacitor ESR and can be estimated as:

$$E_{SR_{\text{MAX}}} = \frac{V_{RIPPLE}}{I_{RIPPLE}}$$  \hspace{1cm} (9)

where $V_{RIPPLE}$ is the maximum output ripple specification and $I_{RIPPLE}$ is the LinkSwitch-TN2 output ripple current (Refer to C2 and C3).

The capacitor ESR value should be specified approximately at the switching frequency of 66 kHz. Capacitor values above 100 $\mu$F are not recommended as they can prevent the output voltage from reaching regulation during the 50 ms period prior to auto-restart. If more capacitance is required, then a soft-start capacitor should be added (see Tips for Designs section). Select a voltage rating such that $V_{\text{RATED}} \geq 1.25 \times V_O$.

Step 8 – Select the Feedback Resistors

The values of $R_{FB}$ and $R_{BIAS}$ are selected such that, at the regulated output voltage, the voltage on the FEEDBACK pin ($V_{FB}$) is 2 V. This voltage is specified for a FEEDBACK pin current ($I_{FB}$) of 49 $\mu$A.

Let the value of $R_{BIAS} = 2.49 \, k\Omega$; this biases the feedback network at a current of ~0.8 mA. Hence the value of $R_{FB}$ is given by:

$$R_{FB} = \frac{V_{O} - V_{FB}}{V_{FB}} = \frac{(V_{O} - V_{FB}) \times R_{BIAS}}{V_{FB} + (I_{FB} \times R_{BIAS})}$$  \hspace{1cm} (10)

Step 9 – Select the Feedback Diode and Capacitor

For the feedback capacitor, use a 10 $\mu$F general purpose electrolytic capacitor with a voltage rating $\geq 1.25 \times V_O$. For the feedback diode, use a glass passivated 1N4005GP or 1DFLR1600-7 device with a voltage rating of $\geq 1.25 \times V_{\text{MAX}}$.

Step 10 – Select the External Biased Resistor for BYPASS Pin

To reduce the no-load input power of the power supply, resistor R5 in Figure 9 of the application example, connected from the feedback capacitor C3 to the BYPASS pin, is recommended. This is applicable to the power supply whose output voltage is higher than $V_{\text{BP(SHUNT)}}$.

To achieve lowest no load power consumption, the current fed into the BYPASS pin should be slightly higher than 120 $\mu$A. For the best full load efficiency and thermal performance, the current fed into the BYPASS pin should be slightly higher than the current value stated below:

<table>
<thead>
<tr>
<th>Part Number</th>
<th>Bypass Current</th>
</tr>
</thead>
<tbody>
<tr>
<td>LNK3202</td>
<td>195 $\mu$A</td>
</tr>
<tr>
<td>LNK3204</td>
<td>222 $\mu$A</td>
</tr>
<tr>
<td>LNK3205</td>
<td>269 $\mu$A</td>
</tr>
<tr>
<td>LNK3206</td>
<td>290 $\mu$A</td>
</tr>
</tbody>
</table>

Table 11. Bypass Pin Injection Current when MOSFET is Switching.

The BYPASS pin current should not exceed 16 mA ($I_{\text{BP(MAX)}}$) at the maximum output voltage (normally when the output voltage is at no-load condition).

Step 11 – Select Pre-Load Resistor

In high-side, direct feedback designs where the minimum load is <3 mA, a pre-load resistor is required to maintain output regulation. This ensures sufficient inductor energy to pull the inductor side of the feedback capacitor $C_{FB}$ to input return via $D_{FB}$. The value of $R_{PL}$ should be selected to give a minimum output load of 3 mA.

In designs with an optocoupler a Zener diode or reference bias current provides a 1 mA to 2 mA minimum load, preventing “pulse bunching” and increased output ripple at zero load.
**Tips for Designs**

**Start-Up Into Non-Resistive Loads**

If the total system capacitance is >100 µF or the output voltage is >12 V, then during start-up the output may fail to reach regulation within 50 ms which can trigger auto-restart protection feature. This may also be true when the load is not resistive, for example, the output is supplying a motor or fan. To increase the start-up time, a soft-start capacitor can be added across the feedback resistor, as shown in Figure 5. The value of this soft-start capacitor is typically in the range of 0.47 µF to 47 µF with a voltage rating of 1.25 × V<sub>o</sub>. Addition of this capacitor can lead to instability in some designs that resembles bunching of switching cycles hence this recommendation should be carefully verified by measuring output ripple under different operating conditions.

**Generating Negative and Positive Outputs**

In appliance applications there is often a requirement to generate both an AC line referenced positive and negative output. This can be accomplished using the circuit in Figure 6. The two Zener diodes have a voltage rating close to the required output voltage for each rail and ensure that regulation is maintained when one rail is lightly and the other heavily loaded. The LinkSwitch-TN2 circuit is designed as if it were a single output voltage with an output current equal to the sum of both outputs. The magnitude sum of the output voltages in this example is 12 V.

**Constant Current Circuit Configuration (LED Driver)**

The circuit shown in Figure 7 is ideal for driving constant current loads such as LEDs. It uses the tight tolerance and temperature stable FEEDBACK pin of LinkSwitch-TN2 IC as the reference to provide an accurate output current. To generate a constant current output, the average output current is converted to a voltage by resistor R<sub>SENSE</sub> and capacitor C<sub>SENSE</sub> and fed into the FEEDBACK pin via R<sub>B</sub> and R<sub>BIAS</sub>. With the values of R<sub>B</sub> and R<sub>BIAS</sub> as shown, the value of R<sub>SENSE</sub> should be chosen to generate a voltage drop of 2 V at the required output current. Capacitor C<sub>SENSE</sub> filters the voltage across R<sub>SENSE</sub>, which is modulated by inductor ripple current. The value of C<sub>SENSE</sub> should be large enough to minimize the ripple voltage, especially in M<sub>DC</sub>-CH designs. A value of C<sub>SENSE</sub> is selected such that the time constant (t) of R<sub>SENSE</sub> and C<sub>SENSE</sub> is greater than 20 times that of the switching period (15 µs). The peak voltage seen by C<sub>SENSE</sub> is equal to R<sub>SENSE</sub> × I<sub>LIMIT(MAX)</sub>. The output capacitor is optional; however with no output capacitor the load will see the full peak current (I<sub>LIMIT</sub>) of the selected LinkSwitch-TN2 device. Increase the value of C<sub>O</sub> (typically in the range of 100 nF to 10 µF) to reduce the peak current to an acceptable level for the load.

If the load is disconnected, feedback is lost and the large output voltage which results may cause circuit failure. To prevent this, a second voltage control loop, D<sub>FB</sub> and V<sub>REF</sub>, can be added as shown in Figure 7. This also requires that capacitor C<sub>O</sub> is used in the circuit. The voltage of the Zener is selected as the next standard value above the maximum voltage across the LED string when it is in constant current operation. The same design equations / design spreadsheet can be used as for a standard buck-boost design, with the following additional considerations:

1. V<sub>o</sub> = LED V<sub>F</sub> × Number of LEDs per string.
2. I<sub>o</sub> = LED I<sub>L</sub> × Number of strings.
3. Lower efficiency estimate due to R<sub>SENSE</sub> losses (enter R<sub>SENSE</sub> into design spreadsheet as inductor resistance).
4. Set R<sub>B</sub> = 2 kΩ and R<sub>FB</sub> = 300 Ω.
5. R<sub>SENSE</sub> = 2/I<sub>L</sub>.
6. C<sub>SENSE</sub> = 20 × (15 µs/R<sub>SENSE</sub>).  
7. Select C<sub>O</sub> based on acceptable output ripple current through the load.
8. If the load can be disconnected or for additional fault protection, add voltage feedback components D<sub>FB</sub> and V<sub>REF</sub> in addition to C<sub>O</sub>.

**Thermal Environment**

To ensure good thermal performance, the SOURCE pin temperature should be maintained below 100 °C, by providing adequate heat sinking. For applications with high ambient temperature (>50 °C), it is recommended to build and test the power supply at the maximum operating ambient temperature and ensure that there is adequate thermal margin. The maximum output current specified in the data sheet is based on specific operating conditions and may need to be thermally derated. Also, it is recommended to use ultrafast (≤35 ns) low recovery reverse diodes at high operating temperatures (>70 °C). If the device temperature exceeds 85 °C with ambient temperature of 25 °C, it is recommended the next bigger device in the family should be selected for the application.
A battery powered thermocouple meter is recommended to make measurements when the SOURCE pins are a switching node. Alternatively, the ambient temperature may be raised to indicate margin to thermal shutdown.

**Recommended Layout Considerations**

In the buck or buck-boost converter configuration, since the SOURCE pins in LinkSwitch-TN2 devices are switching nodes, the copper area connected to SOURCE should be minimized to minimize EMI within the thermal constraints of the design.

In the boost configuration, since the SOURCE pins are tied to DC return, the copper area connected to SOURCE can be maximized to improve heat sinking.

Figure 8 are printed circuit board layout design examples for the circuit schematic shown in Figure 9. The loop formed between the LinkSwitch-TN2, inductor (L1), freewheeling diode (D1), and output capacitor (C2) should be kept as small as possible. The loop between the input capacitor C5, IC DRAIN pin, SOURCE pin, freewheeling diode cathode and anode should be as small as possible. The BYPASS pin capacitor C1 (Figure 9) should be located physically close to the SOURCE and BYPASS pins.

Most off-the-shelf inductors are drum core inductors or dogbone inductors. These inductors do not have a good closed magnetic path, and are a source of significant magnetic coupling. They are a source of differential mode noise. To minimize direct coupling from switching nodes, the LinkSwitch-TN2 IC should be placed away from AC input lines. It may be advantageous to place capacitors C4 and C5 in-between LinkSwitch-TN2 device and the AC input. In a drum core inductor, the winding is typically wound in multiple layers. In a given design, result of EMI performance and regulation can change depending on which end of the inductor is connected to the output capacitor and which end is connected to the SOURCE pin of the IC. It is therefore recommended that effect of change of orientation of the inductor be verified and the favorable inductor orientation be used. Once the favorable orientation is determined, the same should be consistently followed on all boards manufactured so as to ensure repeatable performance. Typically inductors are marked to indicate the start and end of the windings. These markings can be used to correctly orient the inductors during assembly.

The second rectifier diode D4 is optional, but may be included for better EMI performance and higher line surge withstand capability. Traces carrying high currents should be as short in length and thick in width as possible. These are the traces which connect the input capacitor, LinkSwitch-TN2 IC, inductor, freewheeling diode, and the output capacitor.

**Design for Safety Compliance**

Power supplies are required to have capability of withstanding surge voltages which typically are a result of events such as lightning strikes. It is expected that such events do not lead to failure of any components or loss of functionality. Standards such as IEC61000-4-5 defines surge voltage and current waveforms as well as source impedance, which emulate typical worst case transients for testing of protection mechanisms for line connected power circuits and data line connected equipment.

Components of the fusible resistor, EMI filter and the capacitors used in the power supply input stage, help in limiting the voltage and current stress that the components of the power supply are subjected to during these events.
MOVs will often be required to be added at the input of the power supply if the surge level is high (DM surge > 1 kV). These MOVs are placed after the input fuse and help in clamping the voltage at the input of the power supply when a surge event occurs.

The following checklist can be used to ensure that the design is compliant to the applicable requirements:

- Define the target market for LinkSwitch-TN2 converter.
- Determine the equipment class to determine common-mode (CM) and differential-mode (DM) surge levels.
- If DM surge >1000 V, then you will likely need to include an MOV across the AC line at the front-end of the EMI filter.
- Select a MOV for North America 115 VAC or universal input with adequate stand-off voltage during normal operation as well as adequate rated surge current and energy capacity.
- An example of selecting an MOV: Assume that you have a North America application within a Class 3 equipment installation for which you need to select a MOV for differential mode protection, connected across the AC line. The DM Spike Energy will be less than 6.9 J. A device rated for 150 VAC continuous operation would provide adequate stand-off voltage for 115 VAC nominal applications. Littelfuse part number V150LA5 provides 25 J and 2500 A surge capability with adequate margin to minimize degraded performance due to accumulated strikes over the life of the MOV.
- For a universal input design, the V320LA10 provides 48 J and 2500 A surge capability.
- Conduct both common-mode and differential-mode surge tests on the converter and observe voltages across key components and currents where necessary to validate SOA operation of components.

Verify all voltage and current extremes are within the rated specification of each X and Y capacitor. If not, specify a component with a higher rating.

- Verify surge transient current rating of the diode bridge used.
- Verify MOSFET switch BV rating is greater than surge voltage on switching node. If not, you may need to increase bulk capacitor size to prevent the surge energy from increasing the capacitor voltage to objectionable levels.
- Ensure that bulk capacitor surge voltage rating is not exceeded during testing. If surge voltage rating is exceeded, you may need to increase capacitance. Some capacitors may tolerate higher than the rated surge voltage for short durations however capacitor manufacturers should be consulted for guidance.

When making measurements on a power supply during a line surge or safety test, care should be taken to ensure that the test equipment is galvanically isolated. If alternate paths for the surge energy are created as a result of connection of test probes, the test result will be incorrect. Care must be taken to use voltage probes that are rated for measurement of high-voltages in excess of the voltages likely to be encountered during the test.
Appendix A – Application Example

A 1.44 W Universal Input Buck Converter

The circuit shown in Figure 9 is a typical implementation of a 12 V, 120 mA non-isolated power supply used in appliance control such as rice cookers, dishwashers or other white goods. This circuit may also be applicable to other applications such as night-lights, LED drivers, electricity meters, and residential heating controllers, where a non-isolated supply is suitable.

![Figure 9. Universal Input, 12 V, 120 mA Constant Voltage Power Supply using LinkSwitch-TN2.](PI-7557-092616)
Appendix B

Calculations for Inductor Value for Buck and Buck-Boost Topologies

There is a minimum value of inductance that is required to deliver the
specified output power, regardless of line voltage and operating mode.

As a general case, Figure 10 shows the inductor current in discontinu-
ous conduction mode (DCM). The following expressions are valid for
both CCM as well as DCM operation. There are three unique intervals
in DCM as can be seen from Figure 10. Interval $t_{ON}$ is when the
LinkSwitch-TN2 IC is ON and the freewheeling diode is OFF. Current
ramps up in the inductor from an initial value of zero. The peak
current is the current limit $I_{LIMIT}$ of the device. Interval $t_{OFF}$ is when
the LinkSwitch-TN2 IC is OFF and the freewheeling diode is ON.
Current ramps down to zero during this interval. Interval $t_{IDLE}$ is when
both the LinkSwitch-TN2 IC and freewheeling diode are OFF, and the
inductor current is zero.

In CCM, this idle state does not exist and thus $t_{IDLE} = 0$.

We can express the current swing at the end of interval $t_{ON}$ in a buck
converter as:

$$\Delta I(t_{ON}) = I_{RIPPLE} = \frac{V_{MIN} - V_{DS} - V_{O}}{L_{MIN}} \times t_{ON} \quad (C1)$$

$$I_{RIPPLE} = 2 \times (I_{LIMIT} - I_{O}) \quad t_{IDLE} = 0 \text{(for CCM)} \quad (C2)$$

$$I_{RIPPLE} = 2 \times (I_{LIMIT} - I_{O}) \quad t_{IDLE} > 0 \text{(for MDCM)} \quad (C3)$$

where

$I_{RIPPLE}$ = Inductor ripple current
$I_{LIMIT} = $ Minimum current limit
$V_{MIN} =$ Minimum DC bus voltage
$V_{DS} =$ On-state Drain to Source voltage drop
$V_{O} =$ Output voltage
$L_{MIN} =$ Minimum inductance

Similarly, we can express the current swing at the end of interval $t_{OFF}$
as:

$$\Delta I(t_{OFF}) = I_{RIPPLE} = \frac{V_{O} + V_{FD}}{L_{MIN}} \times t_{OFF} \quad (C4)$$

The initial current through the inductor at the beginning of each
switching cycle can be expressed as:

$$I_{INITIAL} = I_{LIMIT} - I_{RIPPLE} \quad (C5)$$

The average current through the inductor over one switching cycle is
equal to the output current $I_{O}$. This current can be expressed as:

$$I_{O} = \frac{1}{T_{SW, MAX}} \left( \frac{1}{2} \times (I_{LIMIT} + I_{INITIAL}) \times t_{ON} + \frac{1}{2} \times (I_{LIMIT} + I_{INITIAL}) \times t_{OFF} + 0 \times t_{IDLE} \right) \quad (C6)$$

Where

$I_{O} =$ Output current.
$T_{SW, MAX} =$ The switching interval corresponding to minimum switching
frequency $f_{SW}$.

Substituting for $t_{ON}$ and $t_{OFF}$ from equations (C1) and (C4) we have:

$$I_{O} = \frac{1}{T_{SW, MAX}} \left( \frac{1}{2} \times (I_{LIMIT} + I_{INITIAL}) \times I_{RIPPLE} \times I_{MIN} \times \frac{V_{MIN} - V_{DS} - V_{O}}{V_{O} + V_{FD}} \right) \quad (C7)$$

For MDCM design, $I_{INITIAL} = 0$, $I_{RIPPLE} = I_{LIMIT}$.  

$$I_{MIN} = \frac{2 \times (V_{O} + V_{FD}) \times I_{O} \times (V_{MIN} - V_{DS} - V_{O})}{(I_{LIMIT} - I_{INITIAL}) \times F_{SM} \times (V_{MIN} - V_{DS} + V_{FD})} \quad (C8)$$

For CCM design, $t_{IDLE} = 0$.

$$I_{O} = \frac{1}{2} \times (I_{LIMIT} + I_{INITIAL}) \quad (C9)$$

$$I_{INITIAL} = I_{LIMIT} \quad (C10)$$

$$L_{MIN} = \frac{2 \times (V_{O} + V_{FD}) \times (V_{MIN} - V_{DS} - V_{O})}{(I_{LIMIT} - I_{O}) \times F_{SM} \times (V_{MIN} - V_{DS} + V_{FD})} \quad (C11)$$

For output voltages greater than 20 V, use $V_{MAX}$ for calculation of $L_{MIN}$
(Equation C8). For output voltages less than 20 V, use $V_{MIN}$ for
calculation of $L_{MIN}$ to compensate for current limit delay time over-
shoot.

This however does not account for the losses within the inductor
(resistance of winding and core losses) and the freewheeling diode,
which will limit the maximum power delivering capability and thus
reduce the maximum output current. The minimum inductance must
compensate for these losses in order to deliver specified full load
power. An estimate of these losses can be made by estimating the
total losses in the power supply, and then allocating part of these
losses to the inductor and diode. This is done by the loss factor $K_{LOSS}$
which increases the size of the inductor accordingly. Furthermore,
typical inductors for this type of application are bobbin core or dog
bone chokes. The specified current rating refer to a temperature rise
of 20 °C or 40 °C and to an inductance drop of 10%. We must

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As a general case, Figure 10 shows the inductor current in discontinu-
ous conduction mode (DCM). The following expressions are valid for
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in DCM as can be seen from Figure 10. Interval $t_{ON}$ is when the
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Current ramps down to zero during this interval. Interval $t_{IDLE}$ is when
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converter as:

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$$I_{RIPPLE} = 2 \times (I_{LIMIT} - I_{O}) \quad t_{IDLE} = 0 \text{(for CCM)} \quad (C2)$$

$$I_{RIPPLE} = 2 \times (I_{LIMIT} - I_{O}) \quad t_{IDLE} > 0 \text{(for MDCM)} \quad (C3)$$

where

$I_{RIPPLE}$ = Inductor ripple current
$I_{LIMIT} = $ Minimum current limit
$V_{MIN} =$ Minimum DC bus voltage
$V_{DS} =$ On-state Drain to Source voltage drop
$V_{O} =$ Output voltage
$L_{MIN} =$ Minimum inductance

Similarly, we can express the current swing at the end of interval $t_{OFF}$
as:

$$\Delta I(t_{OFF}) = I_{RIPPLE} = \frac{V_{O} + V_{FD}}{L_{MIN}} \times t_{OFF} \quad (C4)$$
incorporate an inductance tolerance factor $K_{L_TOL}$ within the expression for minimum inductance, to account for this manufacturing tolerance. The typical inductance value thus can be expressed as:

$$L_{TYP} = \frac{(1 + K_{L_TOL}) \times L_{MIN}}{K_{LOSS}}$$  \hspace{1cm} (C13) $$

where $K_{LOSS}$ is a loss factor, which accounts for the off-state total losses of the inductor.

$K_{L_TOL}$ is the inductor tolerance factor and can be between 10% and 20%. A typical value is 0.15. With this typical inductance we can express maximum output power as:

$$P_{O,MAX} = L_{TYP} \times FS_{MIN} \times (I_{MIN,SW}^2 - I_{INITIAL}^2) \times \frac{V_o \times (V_{SW} - V_{DS} - V_o)}{2 \times (V_{SW} - V_{DS} - V_o) \times (V_o + V_{FD})} \times \frac{K_{LOSS}}{(1 + K_{L_TOL})}$$

Similarly for buck-boost topology the expressions for $L_{TYP}$ and $P_{O,MAX}$ are:

$$L_{TYP} = 2(1 + K_{L_TOL}) \times (V_o + V_{FD}) \times \frac{I_o \times (V_{MIN} - V_{DS})}{K_{LOSS} \times (I_{LIMIT,MIN}^2 - I_{INITIAL}^2) \times FS_{MIN} \times (V_{MIN} - V_{DS} + V_{FD} + V_o)}$$

$$P_{O,MAX} = L_{TYP} \times FS_{MIN} \times (I_{LIMIT,MIN}^2 - I_{INITIAL}^2) \times \frac{V_o \times (V_{MIN} - V_{DS} + V_{FD} + V_o)}{2 \times (V_{MIN} - V_{DS} \times (V_o + V_{FD})} \times \frac{K_{LOSS}}{(1 + K_{L_TOL})}$$

Average Switching Frequency

Since LinkSwitch-TN2 uses an on-off type of control, the frequency of switching is non-uniform due to cycle skipping. We can average this switching frequency by substituting the maximum power as the output power in Equation C14. Simplifying, we have:

$$FS_{AVG} = 2 \times (1 + K_{L_TOL}) \times (V_o + V_{FD}) \times \frac{I_o \times (V_{MIN} - V_{DS} - V_o)}{K_{LOSS} \times (I_{LIMIT,MIN}^2 - I_{INITIAL}^2) \times L_{TYP} \times (V_{MIN} - V_{DS} + V_{FD} + V_o)}$$

Similarly for buck-boost converter, simplifying Equation C16 we have:

$$FS_{AVG} = 2(1 + K_{L_TOL}) \times (V_o + V_{FD}) \times \frac{I_o \times (V_{MIN} - V_{DS})}{K_{LOSS} \times (I_{LIMIT,MIN}^2 - I_{INITIAL}^2) \times L_{TYP} \times (V_{MIN} - V_{DS} + V_{FD} + V_o)}$$

Calculation of RMS Currents

The RMS current value through the inductor is mainly required to ensure that the inductor is appropriately sized and will not overheat. Also, RMS currents through the LinkSwitch-TN2 IC and freewheeling diode are required to estimate losses in the power supply. Assuming CCM operation, the initial current in the inductor in steady state is given by:

$$I_{INITIAL} = I_{MIN,MIN} - \frac{V_o + V_{FD}}{L_{TYP}} \times t_{OFF}$$

$t_{OFF}$ is when MOSFET is off.

For DCM operation this initial current will be zero.

The current through the LinkSwitch-TN2 as a function of time is given by:

$$i_{SW}(t) = I_{INITIAL} + \frac{V_{MIN} - V_{DS} - V_o}{L_{TYP}} \times t, 0 < t \leq t_{ON}$$

$$i_{SW}(t) = 0, t_{ON} < t \leq T$$

$t_{ON}$ is when MOSFET is on.

The current through the freewheeling diode as a function of time is given by:

$$i_d(t) = 0, 0 < t \leq t_{ON}$$

$$i_d(t) = I_{MIN,MIN} - \frac{V_o + V_{FD}}{L_{TYP}} \times t, t_{ON} < t \leq t_{OFF}$$

$$i_d(t) = 0, t_{OFF} < t \leq T$$

$t_{OFF}$ is when freewheeling diode is on.

And the current through the inductor as a function of time is given by:

$$i_L(t) = i_{SW}(t) + i_d(t)$$

From the definition of RMS currents we can express the RMS currents through the switch, freewheeling diode and inductor as follows:

$$i_{SW,RMS} = \sqrt{\frac{1}{T_{AVG}} \int_0^T i_{SW}(t)^2 \times dt}$$

$$i_{D,RMS} = \sqrt{\frac{1}{T_{AVG}} \int_0^T i_d(t)^2 \times dt}$$

$$i_{L,RMS} = \sqrt{\frac{1}{T_{AVG}} \int_0^T (i_{SW}(t) + i_d(t))^2 \times dt}$$

Since the switch and freewheeling diode currents fall to zero during the turn-off and turn-on intervals respectively, the RMS inductor current is simplified to:

$$i_{L,RMS} = \sqrt{i_{SW,RMS}^2 + i_{D,RMS}^2}$$

Table C1 lists the design equations for important parameters using the buck and buck-boost topologies.
<table>
<thead>
<tr>
<th>Parameter</th>
<th>Buck</th>
<th>Buck-Boost</th>
</tr>
</thead>
<tbody>
<tr>
<td>$L_{\text{typ}}$</td>
<td>$L_{\text{typ}} = 2(1 + K_{\text{L.Tol}}) \times (V_o + V_{FD}) \times \frac{I_o}{K_{\text{LOSS}} \times (I_{\text{LIM.MIN}}^2 - I_{\text{INITIAL}}^2) \times FS_{\text{MIN}} \times (V_{\text{MIN}} - V_{DS} + V_{FD})}$</td>
<td>$L_{\text{typ}} = 2(1 + K_{\text{L.Tol}}) \times (V_o + V_{FD}) \times \frac{I_o}{K_{\text{LOSS}} \times (I_{\text{LIM.MIN}}^2 - I_{\text{INITIAL}}^2) \times FS_{\text{MIN}} \times (V_{\text{MIN}} - V_{DS} + V_{FD} + V_o)}$</td>
</tr>
<tr>
<td>$F_{\text{AVG}}$</td>
<td>$F_{\text{AVG}} = 2 \times (1 + K_{\text{L.Tol}}) \times (V_o + V_{FD}) \times \frac{I_o}{K_{\text{LOSS}} \times (I_{\text{LIM.MIN}}^2 - I_{\text{INITIAL}}^2) \times L_{\text{typ}} \times (V_{\text{MIN}} - V_{DS} + V_{FD})}$</td>
<td>$F_{\text{AVG}} = 2 \times (1 + K_{\text{L.Tol}}) \times (V_o + V_{FD}) \times \frac{I_o}{K_{\text{LOSS}} \times (I_{\text{LIM.MIN}}^2 - I_{\text{INITIAL}}^2) \times L_{\text{typ}} \times (V_{\text{MIN}} - V_{DS} + V_{FD} + V_o)}$</td>
</tr>
<tr>
<td>$i_{\text{sw}}(t)$ LinkSwitch-TN2 Current</td>
<td>$i_{\text{sw}}(t) = \begin{cases} I_{\text{INITIAL}} + \frac{V_{\text{MIN}} - V_{DS} - V_o}{L_{\text{typ}}} \times t, &amp; 0 &lt; t \leq t_{\text{ON}} \ i_{\text{sw}}(t) = 0, &amp; t_{\text{ON}} &lt; t \leq T \end{cases}$</td>
<td>$i_{\text{sw}}(t) = \begin{cases} I_{\text{INITIAL}} + \frac{V_{\text{MIN}} - V_{DS}}{L_{\text{typ}}} \times t, &amp; 0 &lt; t \leq t_{\text{ON}} \ i_{\text{sw}}(t) = 0, &amp; t_{\text{ON}} &lt; t \leq T \end{cases}$</td>
</tr>
<tr>
<td>$i_{\text{d}}(t)$ Diode Forward Current</td>
<td>$i_{\text{d}}(t) = \begin{cases} 0, &amp; 0 &lt; t \leq t_{\text{ON}} \ I_{\text{LIM.MIN}} - \frac{V_o + V_{FD}}{L_{\text{typ}}} \times t, &amp; t_{\text{ON}} &lt; t \leq t_{\text{OFF}} \ i_{\text{d}}(t) = 0, &amp; t_{\text{OFF}} &lt; t \leq T \end{cases}$</td>
<td>$i_{\text{d}}(t) = \begin{cases} 0, &amp; 0 &lt; t \leq t_{\text{ON}} \ I_{\text{LIM.MIN}} - \frac{V_o + V_{FD}}{L_{\text{typ}}} \times t, &amp; t_{\text{ON}} &lt; t \leq t_{\text{OFF}} \ i_{\text{d}}(t) = 0, &amp; t_{\text{OFF}} &lt; t \leq T \end{cases}$</td>
</tr>
<tr>
<td>$i_{L}(t)$ Inductor Current</td>
<td>$i_{L}(t) = i_{\text{sw}}(t) + i_{\text{d}}(t)$</td>
<td>$i_{L}(t) = i_{\text{sw}}(t) + i_{\text{d}}(t)$</td>
</tr>
<tr>
<td>Max Drain Voltage</td>
<td>$V_{\text{MAX}}$</td>
<td>$V_{\text{MAX}} + V_o$</td>
</tr>
</tbody>
</table>

Table C1. Circuit Characteristics for Buck and Buck-Boost Topologies.
Appendix C – Protection Feature for Flyback Applications

Hysteretic Output Overvoltage Protection

In flyback topology, the output overvoltage protection provided by the LinkSwitch-TN2 IC uses auto-restart that is triggered by a current $> I_{BP(SD)}$ into the BYPASS pin. To prevent inadvertent triggering of this feature, in addition to an internal filter, the BYPASS pin capacitor provides external filtering. For the bypass capacitor to be effective as a high frequency filter, the capacitor should be located as close as possible to the SOURCE and BYPASS pins of the device.

The OVP function can be realized in a non-isolated flyback converter by connecting a Zener diode from the output to the BYPASS pin. The circuit example shown in Figure 11 describes a simple method for implementing the output overvoltage protection. Additional filtering for the OVP detection feature, can be achieved by inserting a low value (10 Ω to 47 Ω) resistor in series with the OVP Zener diode. The resistor in series with the OVP Zener diode also limits the maximum current into the BYPASS pin. The current should be limited to less than 16 mA.

During a fault condition resulting from loss of feedback, the output voltage will rapidly rise above the nominal voltage. A voltage at the output that exceeds the sum of the voltage rating of the Zener diode and the BYPASS pin voltage will cause a current in excess of $I_{BP(SD)}$ injected into the BYPASS pin, which will trigger the auto-restart and protect the power supply from overvoltage.

Line Overvoltage Protection

In a flyback converter configuration, during the power MOSFET on-time, the LinkSwitch-TN2 IC can sense indirectly the DC bus overvoltage condition by monitoring the current flowing into the FEEDBACK pin depending on circuit configuration. Figure 12 shows one possible circuit implementation. During the power MOSFET on-time, the voltage across the secondary winding is proportional to the voltage across the primary winding. The current flowing through the base-emitter and base of transistor Q3 is therefore directly proportional to the $V_{BUS}$ voltage.

$$V_{BUS} = V_{BE3} - V_{DS}$$  \[(D1)\]

$V_{DS}$ is much smaller compared to the bus voltage which can be neglected.

The voltage across the secondary winding is proportional to the voltage across the primary winding.

$$V_{SEC} = \frac{V_{PRIBUS}}{n}$$  \[(D2)\]

$$-V_{BP} + V_{VR3(n+1)} + V_{DS} + V_{BE3} + V_{B3} = V_{SEC}$$  \[(D3)\]

The voltage across the Zener diode VR3 is therefore dependent on $V_{BUS}$. When the line voltage is higher than its threshold and the Zener diode VR3 is turned on, transistor Q3 is turned on and current will flow into FEEDBACK pin from the BYPASS pin capacitor through transistor Q3. When the fed current is higher than FEEDBACK pin...
The current into FEEDBACK pin is the collector current of Q3 if the transistor is not saturated, which is calculated as:

\[
I_{Q3,EC} = h_{FE} \times I_{Q3,EB} = h_{FE} \times \left[ \frac{(V_{REF} - V_{DS})}{N} + V_{BE} - V_{Q3,EB} - V_{DS} - V_{MR3} \right] \frac{R3}{R4} \tag{D8}
\]

The current of \(I_{Q3,EC}\) should not exceed 120% of \(I_{FB(SD)}\) in order to limit the current into the FEEDBACK pin.

In order to have accurate line OV threshold voltage and also for good efficiency, regulation performance and stability, the transformer leakage inductance should be minimized. Low leakage will minimize ringing on the secondary winding and provide accurate line OVP detection. The current into the FEEDBACK pin is sampled and compared to \(I_{FB(SD)}\) typically 280 ns after the high-voltage power MOSFET is turned on.

In some designs if the ringing at the secondary winding is longer than 280 ns, a RC snubber across the rectifier diode may be needed to damp the ringing to ensure precise detection of line voltage.

Below is an example with 33 V Zener (VR3) BZX74-C33, and the threshold is at 308 V. When the bus voltage is higher than the threshold, the power supply goes into auto-restart. The first time a fault is asserted the off-time is 150 ms (\(t_{AR(OFF)}\) – first off period). If the fault condition persists, subsequent off-times are 1500 ms long (\(t_{AR(OFF)}\) subsequent periods).

**Figure 13. Indirect Line-Sensing for Overvoltage Protection Result.**
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