LinkSwitch®-TN
Design Guide
Application Note AN-37

Introduction

LinkSwitch-TN combines a high voltage power MOSFET switch with an ON/OFF controller in one device. It is completely self-powered from the DRAIN pin, has a jittered switching frequency for low EMI and is fully fault protected. Auto-restart limits device and circuit dissipation during overload and output short circuit while over temperature protection disables the internal 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.

LinkSwitch-TN is designed for any application where a non-isolated supply is required such as appliances (coffee machines, rice cookers, dishwashers, microwave ovens etc.), nightlights, emergency exit signs and LED drivers. *LinkSwitch-TN* 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, no μF's of X class capacitance needed
- High efficiency – Full load efficiencies >75% typical for 12 V output
- 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 <100 mW at no load
- Entirely manufacturable in SMD
- More robust to drop test mechanical shock
- Fully fault protected (overload, short circuit and thermal faults)
- Scalable – *LinkSwitch-TN* family allows the same basic design to be used from <50 mA to 360 mA

Scope

This application note is for engineers designing a non-isolated power supply using the *LinkSwitch-TN* family of devices. This

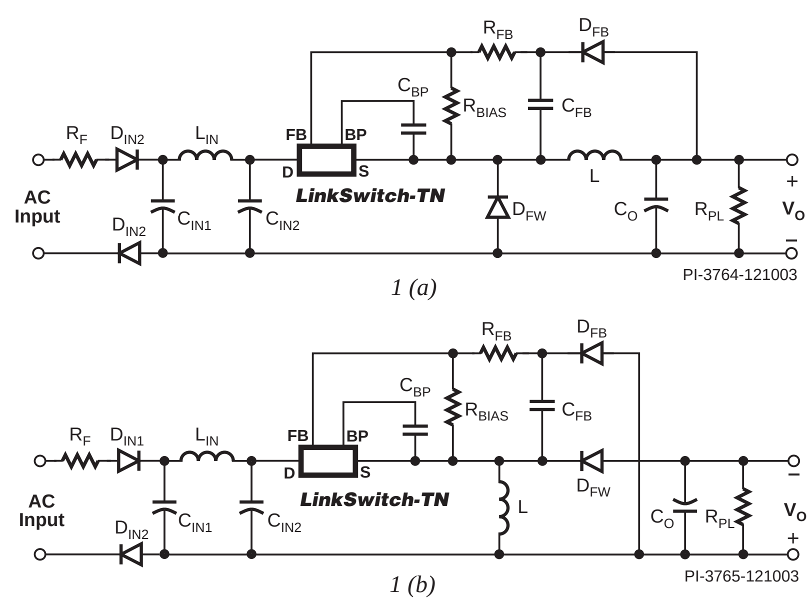


Figure 1 (a). Basic Configuration using LinkSwitch-TN in a Buck Converter. Figure 1 (b) Basic Configuration using LinkSwitch-TN in a Buck-Boost Converter.

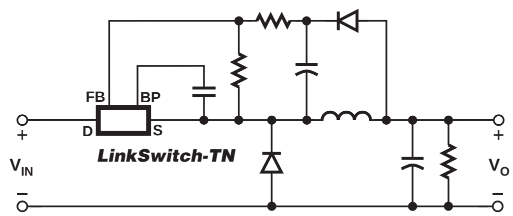
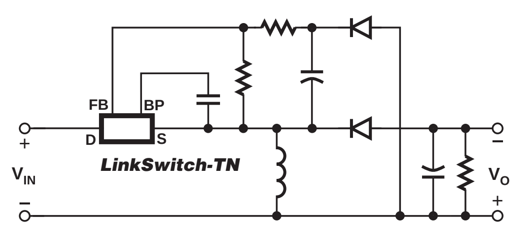
document describes the design procedure for buck and buck-boost converters using the *LinkSwitch-TN* family of integrated off-line switchers. The objective of this document is to provide power supply engineers with guidelines in order to enable them to quickly build efficient and low cost buck or buck-boost 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 common output voltages and currents.

In addition to this application note a design spreadsheet is available within the PIXls tool in the *PI Expert* design software suite. The reader may also find the *LinkSwitch-TN* DAK engineering prototype board useful as an example of a working supply. Further details of support tools and updates to this document can be found at www.powerint.com.

Quick Start

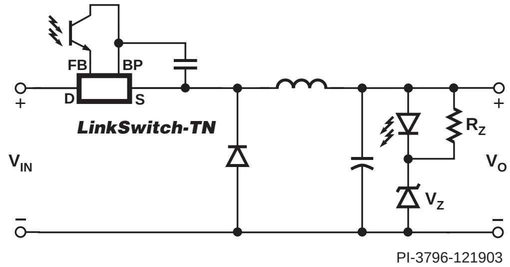
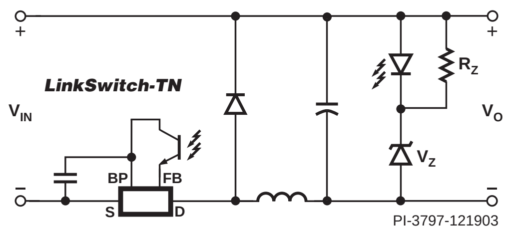
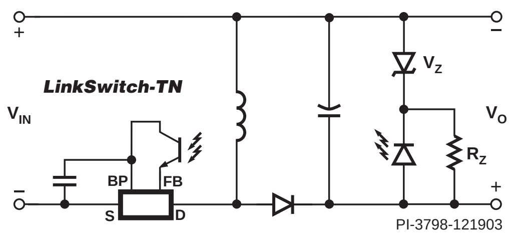
Readers wanting to start immediately can use the following information to quickly select the components for a new design, using Figure 1 and Tables 1 and 2 as references.

1) For AC input designs select the input stage (Table 9).
2) Select the topology (Tables 1 and 2).
   - If better than ±10% output regulation is required, then use opto coupler feedback with suitable reference.
3) Select the *LinkSwitch-TN* device, L, R_{FB} or V_Z, R_{BIAS}, C_{FB}, R_Z and the reverse recovery time for D_{FW} (Table 3: Buck, table 4:Buck-Boost).
4) Select freewheeling diode to meet t_{rr} determined in step 3 (Table 5).
5) For direct feedback designs, if the minimum load < 3 mA then calculate $R_{PL} = V_O / 3$ mA.
6) Select C_O as 100 µF, $1.25 \cdot V_O$, low ESR type.
7) Construct prototype and verify design.

TOPOLOGY	BASIC CIRCUIT SCHEMATIC	KEY FEATURES
High-Side Buck – Direct Feedback		1) Output referenced to input 2) Positive output (V_O) with respect to $-V_{IN}$ 3) Step down – $V_O < V_{IN}$ 4) Low cost direct feedback (±10% typ.)
High-Side Buck Boost – Direct Feedback		1) Output referenced to input 2) Negative output (V_O) with respect to $-V_{IN}$ 3) Step down – $V_O > V_{IN}$ or $V_O < V_{IN}$ 4) Low cost direct feedback (± 10% typ.) 5) Fail-safe – output is not subjected to input voltage if the internal MOSFET fails 6) Ideal for driving LEDs – better accuracy and temperature stability than low-side Buck constant current LED driver

Notes
1. Low cost, directly sensed feedback typically achieves overall regulation tolerance of ± 10%.
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) also possible but not shown.

Table 1. LinkSwitch-TN Circuit Configurations Using Directly Sensed Feedback.

TOPOLOGY	BASIC CIRCUIT SCHEMATIC	KEY FEATURES
High-Side Buck – Optocoupler Feedback		1) Output referenced to input 2) Positive output (V_O) with respect to -V_{IN} 3) Step down – $V_O < V_{IN}$ 4) Optocoupler feedback - Accuracy only limited by reference choice - Low cost non-safety rated optocoupler - No pre-load required 5) Minimum no-load consumption
Low-Side Buck – Optocoupler Feedback		1) Output referenced to input 2) Negative output (V_O) with respect to +V_{IN} 3) Step down – $V_O < V_{IN}$ 4) Optocoupler feedback - Accuracy only limited by reference choice - Low cost non-safety rated optocoupler - No pre-load required
Low-Side Buck Boost – Optocoupler Feedback		1) Output referenced to input 2) Positive output (V_O) with respect to +V_{IN} 3) Step up/down – $V_O > V_{IN}$ or $V_O < V_{IN}$ 4) Optocoupler feedback - Accuracy only limited by reference choice - Low cost non-safety rated optocoupler - No pre-load required 5) Fail-safe – output is not subjected to input voltage if the internal MOSFET fails 6) Minimum no-load consumption

Notes
1. Performance of opto feedback only limited by accuracy of reference (Zener or IC).
2. Optocoupler does not need to be safety approved.
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 Ω to 2 kΩ, 1/8 W, 5%
4. Boost topology (step-up) is also possible but not shown.
5. Optocoupler feedback provides lowest no-load consumption.

Table 2. LinkSwitch-TN Circuit Configurations Using Optocoupler Feedback.

V_{OUT}	$I_{OUT(MAX)}$	INDUCTOR μH	I_{RMS} (mA)	TOKIN	COILCRAFT	LNK30X	MODE	DIODE t_{rr}	R_{FB}	V_Z
5	≤120	680	220	SBC2-681-211	RFB0807-681	LNK304	MDCM	≤75 ns	3.84 kΩ	3.9 V
	160	680	230	SBC2-681-211	RFB0807-681		CCM	≤35 ns		
	175	680	320	SBC3-681-211	RFB0810-681	LNK305	MDCM	≤75 ns		
	225	680	340	SBC4-681-211	RFB0810-681		CCM	≤35 ns		
	280	680	440	SBC4-681-211	RFB0810-681	LNK306	MDCM	≤75 ns		
	360	680	430	SBC4-681-211	RFB0810-681		CCM	≤35 ns		
12	≤85	680	180	SBC2-681-211	RFB0807-681	LNK304	MDCM	≤75 ns	11.86 kΩ	11 V
	120	1000	230	SBC3-102-281	RFB0807-102		MDCM	≤75 ns		
	160	1500	320	SBC3-152-251	RFB0810-152		CCM	≤35 ns		
	175	680	340	SBC3-681-361	RFB0810-681	LNK305	MDCM	≤75 ns		
	225	1000	440	SBC4-102-291	RFB0810-102		CCM	≤35 ns		
	280	680	430	SBC4-681-431	RFB0810-681	LNK306	MDCM	≤75 ns		
	360	1500	400	SBC6-152-451	RFB1010-152		CCM	≤35 ns		
15	≤70	680	160	SBC2-681-211	RFB0807-681	LNK304	MDCM	≤75 ns	15.29 kΩ	13 V
	120	1200	210	-	RFB0807-122		MDCM	≤75 ns		
	160	1800	210	-	RFB0810-182	LNK305	CCM	≤35 ns		
	175	820	310	-	RFB0810-821		MDCM	≤75 ns		
	225	1200	310	-	RFB1010-122	LNK306	CCM	≤35 ns		
	280	820	390	-	RFB1010-821		MDCM	≤75 ns		
	360	1500	390	SBC6-152-451	RFB1010-152		CCM	≤35 ns		
24	≤50	680	130	SBC2-681-211	RFB0807-681	LNK304	MDCM	≤75 ns	25.6 kΩ	22 V
	120	1500	190	SBC4-152-221	RFB0810-152		MDCM	≤75 ns		
	160	2200	180	SBC4-222-211	RFB0810-222	LNK305	CCM	≤35 ns		
	175	1200	280	-	RFB0810-122		MDCM	≤75 ns		
	225	1500	280	SBC6-152-451	RFB1010-152	LNK306	CCM	≤35 ns		
	280	1200	350	-	RFB1010-122		MDCM	≤75 ns		
	360	2200	360	SBC6-222-351	-		CCM	≤35 ns		

Other Standard Components

R_{BIAS}: 2 kΩ, 1%, 1/8 W
C_{BP}: 0.1 μF, 50 V Ceramic
C_{FB}: 10 μF, $1.25 \cdot V_O$
D_{FB}: 1N4005GP
R_Z: 470 Ω to 2 kΩ, 1/8 W, 5%

Table 3. Components Quick Select for Buck Converters.

V_{OUT}	$I_{OUT(MAX)}$	INDUCTOR				LNK30X	MODE	DIODE t_{rr}	R_{FB}	V_Z
		μH	I_{RMS} (mA)	TOKIN	COILCRAFT					
5	≤120	680	220	SBC2-681-211	RFB0807-681	LNK304	MDCM	≤75 ns	3.84 kΩ	3.9 V
	160	680	230	SBC2-681-211	RFB0807-681		CCM	≤35 ns		
	175	680	340	SBC3-681-361	RFB0810-681	LNK305	MDCM	≤75 ns		
	225	680	320	SBC4-681-431	RFB0810-681		CCM	≤35 ns		
	280	680	440	SBC4-681-431	RFB0810-681	LNK306	MDCM	≤75 ns		
	360	680	430	SBC4-681-431	RFB0810-681		CCM	≤35 ns		
12	≤70	680	180	SBC2-681-211	RFB0807-681	LNK304	MDCM	≤75 ns	11.86 kΩ	11 V
	120	1200	220	-	RFB1010-122		MDCM	≤75 ns		
	160	1800	210	-	RFB0807-182		CCM	≤35 ns		
	175	820	320	-	RFB0807-821	LNK305	MDCM	≤75 ns		
	225	1200	310	-	RFB0810-122		CCM	≤35 ns		
	280	820	410	-	RFB0810-821	LNK306	MDCM	≤75 ns		
	360	1800	410	-	RFB1010-182		CCM	≤35 ns		
15	≤50	680	180	SBC2-681-211	RFB0807-681	LNK304	MDCM	≤75 ns	15.29 kΩ	13 V
	120	1500	220	SBC3-152-251	RFB0807-152		MDCM	≤75 ns		
	160	2200	220	SBC4-222-211	RFB0810-222	LNK305	CCM	≤35 ns		
	175	1000	320	SBC4-102-291	RFB0810-102		MDCM	≤75 ns		
	225	1500	320	SBC4-152-251	RFB0810-152	LNK306	CCM	≤35 ns		
	280	1200	400	-	RFB0810-122		MDCM	≤75 ns		
	360	2200	410	SBC6-222-351	RFB1010-222		CCM	≤35 ns		
24	≤35	680	180	SBC2-681-211	RFB0807-681	LNK304	MDCM	≤75 ns	25.6 kΩ	22 V
	120	2200	210	SBC3-222-191	RFB0810-222		MDCM	≤75 ns		
	160	3300	210	SBC4-332-161	RFB0810-332		CCM	≤35 ns		
	175	1800	300	-	RFB0810-182	LNK305	MDCM	≤75 ns		
	225	2200	290	SBC4-222-211	RFB1010-222		CCM	≤35 ns		
	280	1800	370	-	RFB1010-182	LNK306	MDCM	≤75 ns		
	360	3300	410	-	-		CCM	≤35 ns		

Other Standard Components

R_{BIAS}: 2 kΩ, 1%, 1/8 W
C_{BP}: 0.1 μF, 50 V Ceramic
C_{FB}: 10 μF, $1.25 \cdot V_O$
D_{FB}: 1N4005GP
R_Z: 470 Ω to 2 kΩ, 1/8 W, 5%

Table 4. Components Quick Select for Buck-Boost Converters.

PART NO.	V_{RRM} (V)	I_F (A)	t_{rr} (ns)	PACKAGE	MANUFACTURER
MUR160	600	1	50	Leaded	Vishay
UF4005	600	1	75	Leaded	Vishay
BYV26C	600	1	30	Leaded	Vishay/Philips
FE1A	600	1	35	Leaded	Vishay
STTA10 6	600	1	20	Leaded	ST Microelectronics
STTA10 6U	600	1	20	SMD	ST Microelectronics
US1J	600	1	75	SMD	Vishay

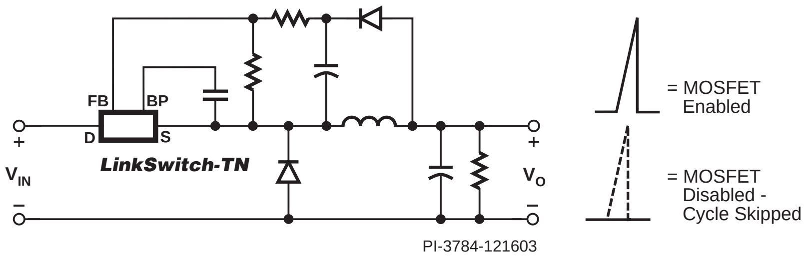
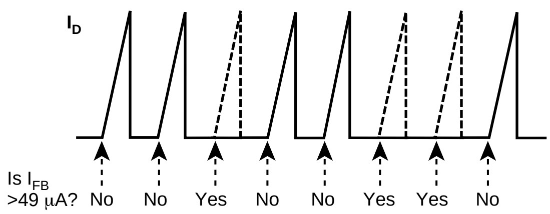
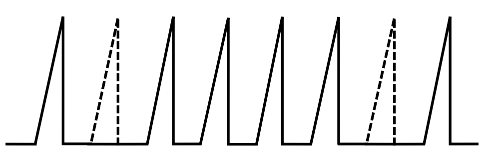
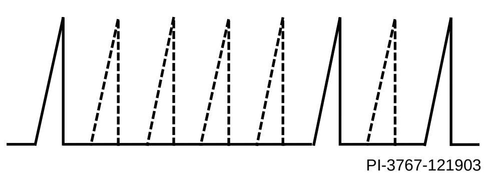
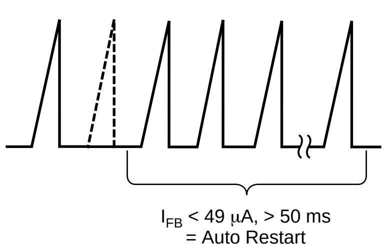
Table 5. List of Ultra-Fast Diodes Suitable for use as the Freewheeling Diode.

LinkSwitch-TN Circuit Design

LinkSwitch-TN Operation

The basic circuit configuration for a Buck converter using *LinkSwitch-TN* is shown in Figure 1a.

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.

<table>
<tr><td>Reference Schematic and Key</td><td colspan="2">
</td></tr>
<tr><td rowspan="3">Normal Operation</td><td>
</td><td>At 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</td></tr>
<tr><td></td><td>High load – few cycles skipped</td></tr>
<tr><td>
</td><td>Low load – many cycles skipped</td></tr>
<tr><td>Auto Restart</td><td>

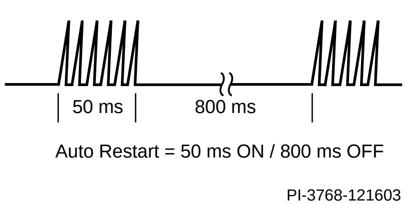

</td><td>If no feedback (I_{FB} < 49 µA) for > 50 ms then output switching is disabled for approximately 800 ms.</td></tr>
</table>

Table 6. LinkSwitch-TN Operation.

To allow direct sensing of the output voltage, without the need for a reference (Zener diode or reference IC), the FB pin voltage is tightly toleranced over the entire operating temperature range. For example, this allows a 12 V design with an overall output tolerance of ±10%. 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-TN* 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-TN* 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-TN* 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

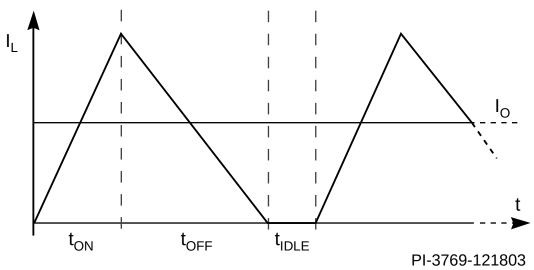
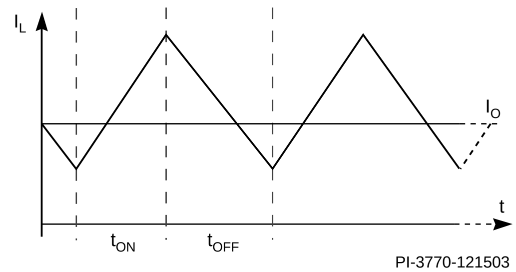
COMPARISON OF CCM AND MDCM OPERATING MODES		
OPERATING MODE	**MDCM**	**CCM**
Operating Description	 Inductor current falls to zero during t_{OFF}, Borderline between MDCM and CCM when t_{IDLE} = 0.	 Current flows continuously in the inductor for the entire duration of a switching cycle.
Inductor	**Lower Cost** Lower value, smaller size.	**Higher Cost** Higher value, larger size.
Freewheeling Diode	**Lower Cost** 75 ns ultra-fast reverse recovery type (≤35 ns for ambient >70 °C).	**Higher Cost** 35 ns ultra-fast recovery type required.
LinkSwitch-TN	**Potentially Higher Cost** May require larger device to deliver required output current–depends on required output current.	**Potentially Lowest Cost** May allow smaller device to deliver required output current–depends on required output current.
Efficiency	**Higher Efficiency** Lower switching losses.	**Lower Efficiency** Higher switching losses.
Overall	**Typically Lower Cost**	**Typically Higher Cost**

Table 7. Comparison of Mostly Discontinuous Conduction (MDCM) and Continuous Conduction (CCM) Modes of Operation.

the low output currents of *LinkSwitch-TN* 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 parameters therefore the *LinkSwitch-TN* (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.

Step-by-Step Design Procedure

Step 1. Determine System Requirements VAC_{MIN}, VAC_{MAX}, P_O, V_O, f_L, η

Determine the input voltage range from Table 8.

Input (VAC)	VAC_{MIN}	VAC_{MAX}
100/115	85	132
230	195	265
Universal	85	265

Table 8. Standard Worldwide Input Line Voltage Ranges.

Line Frequency, f_L: 50 or 60 Hz, for half-wave rectification use $f_L/2$.
Output Voltage, V_O: in Volts.
Output Power, P_O: in Watts.
Power supply efficiency, η: 0.7 for a 12 V output, 0.55 for a 5 V output if no better reference data available.

	Total Capacitance $C_{IN(TOTAL)}$ $\mu F/P_{OUT}$ $(C_{IN1} + C_{IN2})$	
AC Input Voltage (VAC)	Half Wave Rectification	Full Wave Rectification
100/115	6-8	3-4
230	1-2	1
Universal	6-8	3-4

Table 10. 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. The fusible resistor should be chosen as flame proof 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.

For designs ≤1 W it is lower cost to use half-wave rectification, >1 W full wave rectification (smaller input capacitors). 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 9 shows the recommended input stage based on output power for a universal input design while Table 10 shows how to adjust the input capacitance for other input voltage ranges.

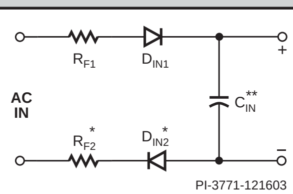
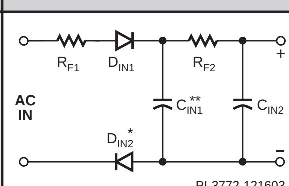
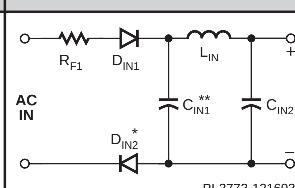
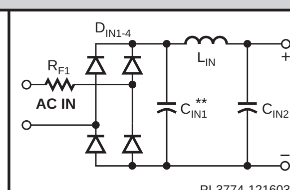
P_{OUT}	≤ 0.25 W	0.25-1 W	> 1 W	
85-265 VAC Input Stage				
	R_{F1}, R_{F2}: 100-470 Ω, 0.5 W, Fusible C_{IN}: ≥2.2 μF, 400 V D_{IN1}, D_{IN2}: 1N4007, 1 A, 1000 V	R_{F1}: 8.2 Ω, 1 W Fusible R_{F2}: 100 Ω, 0.5 W, Flame proof C_{IN1}, C_{IN2}: ≥3.3 μF, 400 V each D_{IN1}, D_{IN2}: 1N4007, 1 A, 1000 V	R_{F1}: 8.2 Ω, 1 W Fusible L_{IN}: 470 μH-2.2 mH, 0.05 A-0.3 A C_{IN1}, C_{IN2}: ≥4 μF/W_{OUT}, 400 V each D_{IN1}, D_{IN2}: 1N4007, 1 A, 1000 V	R_{F1}: 8.2 Ω, 1 W Fusible L_{IN}: 470 μH-2.2 mH, 0.05 A-0.3 A C_{IN1}, C_{IN2}: ≥2 μF/W_{OUT}, 400 V each D_{IN1}, D_{IN2}: 1N4005, 1 A, 600 V
Comments	*Optional for improved EMI and line surge performance. Remove for designs requiring no impedance in return rail. **Increase value to meet required differential line surge performance.			

Table 9. Recommended AC Input Stages For Universal Input.

Step 3. Determine Minimum and Maximum DC Input Voltages V_{MIN} and V_{MAX} Based on AC Input Voltage

Calculate V_{MAX} as

$$V_{MAX} = \sqrt{2} \cdot V_{ACMAX} \qquad (1)$$

Assuming that the value of input fusible resistor is small, the voltage drop across it can be ignored.

Assume bridge diode conduction time of t_c = 3 ms if no other data available.

Derive minimum input voltage V_{MIN}

$$V_{MIN} = \sqrt{\left(2 \cdot V_{ACMIN}{}^2\right) \frac{2 \cdot P_O\left(\frac{1}{2 \cdot f_L} - t_C\right)}{\eta \cdot C_{IN\,(TOTAL)}}} \qquad (2)$$

If V_{MIN} is ≤70 V then increase value of $C_{IN(TOTAL)}$.

Step 4. Select *LinkSwitch-TN* Device Based on Output Current and Current Limit

Decide on 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.

$$I_{LIMIT\_MIN} > 2 \cdot I_O \qquad (3)$$

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_{LIMIT_MIN}.

$$0.5 \cdot I_{LIMIT\_MIN} < I_O < 0.8 \cdot I_{LIMIT\_MIN} \qquad (4)$$

Please see data sheet for *LinkSwitch-TN* current limit values.

Step 5. Select the Output Inductor

Tables 3 and 4 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 higher voltage and/or current above the required output specification. Alternatively the PIXls spreadsheet tool in the *PI Expert* software design suite or Appendix A can be used to calculate the exact inductor value (Eq. A7) and RMS current rating (Eq. A20).

It is recommended that the value of inductor chosen should be closer to L_{TYP} rather than $1.5 \cdot L_{TYP}$ due to lower DC resistance and higher RMS rating. The lower limit of 680 μH limits the maximum di/dt to prevent very high peak current values. Tables 3 and 4 provide reference part numbers for standard inductors from two suppliers.

$$680\ \mu\text{H} < L_{TYP} < L < 1.5 \cdot L_{TYP} \qquad (5)$$

For *LinkSwitch-TN* 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 2 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_{L_TOL} in equation (A7) and the design spreadsheet adjusts for both this drop and the initial inductance value tolerance.

For example if a 680 μH, 360 mA inductor is required, referring to Figure 2, the tolerance is 10% and an estimated 9.5% for the reduction in inductance at the operating current (approximately $[0.36/0.38] \cdot 10$). Therefore the value of K_{L_TOL} = 1.195 (19.5%).

If no data is available assume a K_{L_TOL} of 1.15 (15%).

Not all the energy stored in the inductor is delivered to the load, due to losses in the inductor itself. To compensate for this a loss

Inductance and Tolerance — Current Rating for 20 °C Rise — Current Rating for 40 °C Rise — Current Rating for Value -10%

● SBC3 Series (SBC3-□□□-○○○)

Model	Inductance L(mH/ at 10 kHz	Rdc (W) max.	Rated Current (A) ΔT = 20 °C	Current (Reference Value) (A) ΔT = 40 °C	Current (Reference Value) (A) L change rate -10%
681-361	680±10%	1.62	0.36	0.50	0.38
102-281	1000±10%	2.37	0.28	0.39	0.31
152-251	1500±10%	3.64	0.25	0.35	0.26
222-191	2200±10%	5.62	0.19	0.26	0.21
332-151	3300±10%	7.66	0.15	0.21	0.17

PI-3783-121003

Figure 2. Example of Standard Inductor Data Sheet.

factor K_{LOSS} is used. This has a recommended value of between 50% and 66% of the total supply losses as given by equation (5). For example, a design with an overall efficiency (η) of 0.75 would have a K_{LOSS} value of between 0.875 and 0.833.

$$K_{LOSS} = 1 - \left(\frac{(1-\eta)}{2}\right) \; to \; 1 - \left(\frac{2(1-\eta)}{3}\right) \tag{6}$$

Step 6. Select Freewheeling Diode

For MDCM operation at t_{AMB} ≤70 °C, select an ultra-fast diode with t_{rr} ≤75 ns. At t_{AMB} >70 °C, t_{rr} ≤ 35 ns.

For CCM operation, select an ultra-fast diode with t_{rr} ≤35 ns.

Allowing 25% design margin for the freewheeling diode,

$$V_{PIV} > 1.25 \cdot V_{MAX} \tag{7}$$

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

$$I_F > 1.25 \cdot I_O \tag{8}$$

Table 5 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:

$$ESR_{MAX} = \frac{V_{RIPPLE}}{I_{LIMIT}} \tag{9}$$

where V_{RIPPLE} is the maximum output ripple specification and I_{LIMIT} is the *LinkSwitch-TN* current limit. The capacitor ESR value should be specified approximately at the switching frequency of 66 kHz.

Capacitor values above 100 μF 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 Other Information section).

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 1.65 V. This voltage is specified for a FEEDBACK pin current (I_{FB}) of 49 μA.

Let the value of R_{BIAS} = 2 kΩ; this biases the feedback network at a current of ~0.8 mA. Hence the value of R_{FB} is given by

$$R_{FB} = \frac{V_O - V_{FB}}{\frac{V_{FB}}{R_{BIAS}} + I_{FB}} = \frac{(V_O - V_{FB}) \cdot R_{BIAS}}{V_{FB} + (I_{FB} \cdot R_{BIAS})} = \frac{(V_O - 1.65\ \text{V}) \cdot 2\ \text{k}\Omega}{1.748\ \text{V}} \tag{10}$$

Step 9. Select the Feedback Diode and Capacitor

For the feedback capacitor, use a 10 μF general purpose electrolytic capacitor with a voltage rating ≥1.25 • V_O.

For the feedback diode, use a glass passivated 1N4005GP or 1N4937GP device with a voltage rating of ≥1.25 • V_{MAX}.

Step 10. Select Bypass Capacitor

Use 0.1 μF, 50 V ceramic capacitor.

Step 11. Select Pre-load Resistor

For direct feedback designs if the minimum load <3 mA then calculate R_{PL} = V_O / 3 mA.

Other information

Startup Into Non-Resistive Loads

If the total system capacitance is >100 μF or the output voltage is >12 V, then the output may fail to reach regulation during start-up. This may also be true when the load is not resistive, for example the output is supplying a motor or fan.

To increase the startup time, a soft-start capacitor can be added across the feedback resistor, as shown in Figure 3. 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_O. Figure 4 shows the effect of C_{SS} used on a 12 V, 150 mA design driving a motor load.

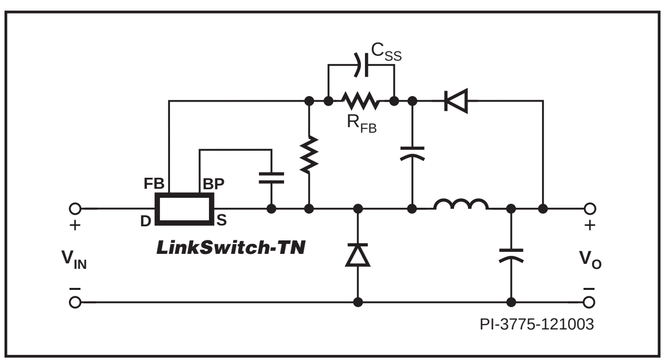


Figure 3. Example Schematic Showing Placement of Soft-Start Capacitor.

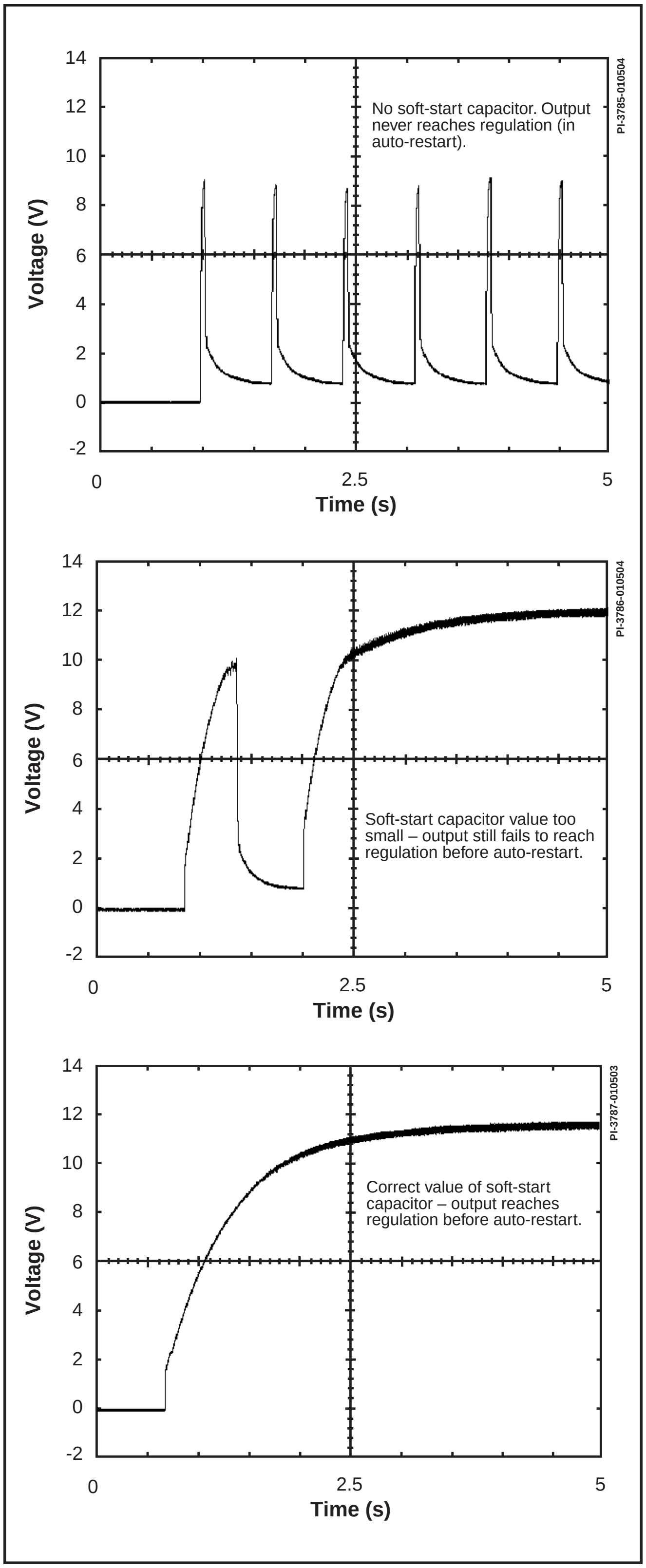


Figure 4. Example of Using Soft-Start Capacitor to Enable Driving a 12 V, 0.15 A Motor Load. All Measurements were made at 85 VAC (worst case condition).

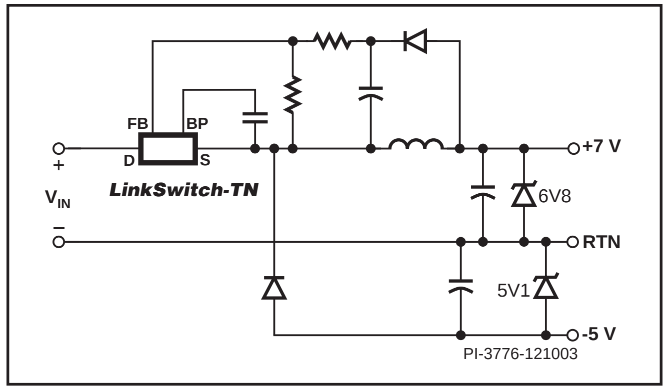


Figure 5. Example Circuit – Generating Dual Output Voltages.

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 5. 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-TN* 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 being 12 V.

Constant Current Circuit Configuration (LED Driver)

The circuit shown in Figure 5 is ideal for driving constant current loads such as LEDs. It uses the tight tolerance and temperature stable FEEDBACK pin of *LinkSwitch-TN* as the reference to provide an accurate output current.

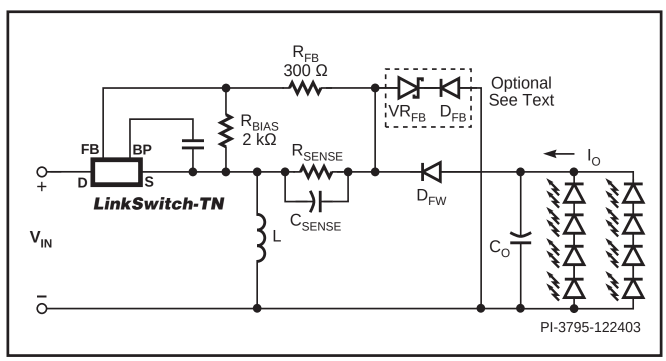


Figure 6. High-Side Buck-Boost Constant Current Output Configuration.

To generate a constant current output, the average output current is converted to a voltage by resistor R_{SENSE} and capacitor C_{SENSE} and fed into the FEEDBACK pin via R_{FB} and R_{BIAS}.

With the values of R_{BIAS} and R_{FB} as shown, the value of R_{SENSE} should be chosen to generate a voltage drop of 2 V at the required output current. Capacitor C_{SENSE} filters the voltage across R_{SENSE}, which is modulated by inductor ripple current. The value of C_{SENSE} should be large enough to minimize the ripple voltage, especially in MDCM designs. A value of C_{SENSE} is selected such that the time constant (t) of R_{SENSE} and C_{SENSE} is greater than 20 times that of the switching period (15 μs). The peak voltage seen by C_{SENSE} is equal to $R_{SENSE} \cdot I_{LIMIT(MAX)}$.

The output capacitor is optional; however with no output capacitor the load will see the full peak current (I_{LIMIT}) of the selected *LinkSwitch-TN*. Increase the value of C_O (typically in the range of 100 nF to 10 uF) 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_{FB} and VR_{FB}, can be added as shown if Figure 6. This also requires that C_O is fitted. 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_O = LED $V_F \cdot$ Number of LEDs per string
2. I_O = LED $I_F \cdot$ Number of strings
3. Lower efficiency estimate due to R_{SENSE} losses (enter R_{SENSE} into design spreadsheet as inductor resistance)
4. Set R_{BIAS} = 2 kΩ and R_{FB} = 300 Ω
5. $R_{SENSE} = 2/I_O$
6. $C_{SENSE} = 20 \cdot (15\ \mu s/R_{SENSE})$
7. Select C_O 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_{FB} and VR_{FB}, in addition to C_O.

Thermal Environment

To ensure good thermal performance, the SOURCE pin temperature should be maintained below 100 °C, by providing adequate heatsinking.

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 figures for maximum output current provided in the data sheet correspond to an ambient temperature of 50 °C, and may need to be thermally derated. Also, it is recommended to use ultra fast (≤35 ns) low reverse recovery diodes at higher operating temperatures (>70 °C).

Recommended Layout Considerations

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-TN*, inductor, freewheeling diode and the output capacitor.

Most off-the-shelf inductors are drum core inductors or dog-bone 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 and for this reason, they should be placed as far away as possible from the AC input lines.

Appendix A

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.

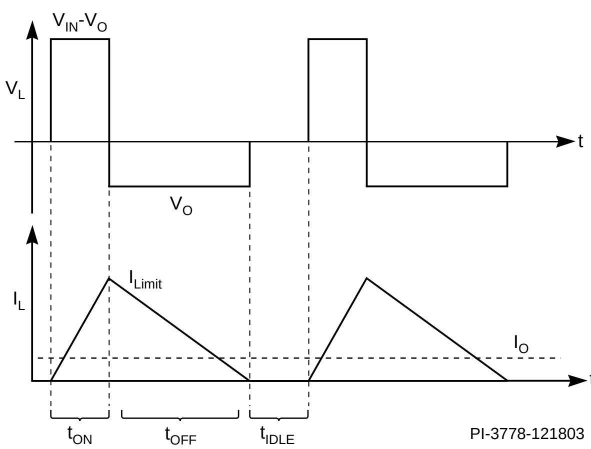


Figure 7. Inductor Voltage and Inductor Current of a Buck Converter in DCM.

As a general case, Figure 7 shows the inductor current in discontinuous 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 7. Interval t_{ON} is when the *LinkSwitch-TN* 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-TN* is OFF and the freewheeling diode is ON. Current ramps down to zero during this interval. Interval t_{IDLE} is when both the *LinkSwitch-TN* and freewheeling diode are OFF, and the inductor current is zero.

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

Neglecting the forward voltage drop of the freewheeling diode, we can express the current swing at the end of interval t_{ON} in a buck converter as

$$\Delta I(t_{ON}) = I_{RIPPLE} = \frac{V_{MIN} - V_{DS} - V_O}{L_{MIN}} \cdot t_{ON}$$

$$I_{RIPPLE} = 2 \cdot (I_{LIMIT\_MIN} - I_O) \quad t_{IDLE} = 0 \; (for\ CCM)$$

$$I_{RIPPLE} = I_{LIMIT\_MIN}, \qquad t_{IDLE} > 0 \; (for\ DCM) \tag{A1}$$

where
I_{RIPPLE} = Inductor Ripple Current
I_{LIMIT_MIN} = 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

$$\Delta I(t_{OFF}) = I_{RIPPLE} = \frac{V_O}{L_{MIN}} \cdot t_{OFF} \tag{A2}$$

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

$$I_{INITIAL} = I_{LIMIT\_MIN} - I_{RIPPLE} \tag{A3}$$

The average current through the inductor over one switching cycle is equal to the output current I_O. This current can be expressed as

$$I_O = \frac{1}{T_{SW\_MAX}} \begin{pmatrix} \frac{1}{2} \cdot (I_{LIMIT\_MIN} + I_{INITIAL}) \cdot t_{ON} + \frac{1}{2} \cdot \\ (I_{LIMIT\_MIN} + I_{INITIAL}) \cdot t_{OFF} + 0 \cdot t_{IDLE} \end{pmatrix} \tag{A4}$$

where

I_O = Output Current.
T_{SW_MAX} = the switching interval corresponding to minimum switching frequency FS_{MIN}.

Substituting for t_{ON} and t_{OFF} from equations (A1) and (A2) we have

$$I_O = \frac{1}{T_{SW\_MAX}} \begin{pmatrix} \frac{1}{2} \cdot (I_{LIMIT\_MIN} + I_{INITIAL}) \frac{I_{RIPPLE} \cdot L_{MIN}}{V_{MIN} - V_{DS} - V_O} \\ + \frac{1}{2} \cdot (I_{LIMIT\_MIN} + I_{INITIAL}) \frac{I_{RIPPLE} \cdot L_{MIN}}{V_O} \end{pmatrix} \tag{A5}$$

$$L_{MIN} = \frac{2 \cdot (V_O \cdot I_O) \cdot (V_{MIN} - V_{DS} - V_O)}{(I_{LIMIT\_MIN}^2 - I_{INITIAL}^2) \cdot FS_{MIN} \cdot (V_{MIN} - V_{DS})} \tag{A6}$$

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 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

$$L_{TYP} = \frac{2 \cdot K_{L\_TOL} \cdot \left(\frac{V_O \cdot I_O}{K_{LOSS}}\right) \cdot (V_{MIN} - V_{DS} - V_O)}{(I_{LIMIT\_MIN}^2 - I_{INITIAL}^2) \cdot FS_{MIN} \cdot (V_{MIN} - V_{DS})} \tag{A7}$$

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 1.1 and 1.2. A typical value is 1.15.

With this typical inductance we can express maximum output power as

$$P_{O\_MAX} = \frac{1}{2} \cdot L_{TYP} \cdot (I_{LIMIT\_MIN}^2 - I_{INITIAL}^2) \cdot FS_{MIN} \cdot \frac{V_{MIN} - V_{DS}}{V_{MIN} - V_{DS} - V_O} \cdot \frac{K_{LOSS}}{K_{L\_TOL}} \tag{A8}$$

Similarly for Buck-Boost topology the expressions for L_{TYP} and P_{O_MAX} are

$$L_{TYP} = \frac{2 \cdot K_{L\_TOL} \cdot \left(\frac{V_O \cdot I_O}{K_{LOSS}} \right)}{\left(I_{LIMIT\_MIN}{}^2 - I_{INITIAL}{}^2 \right) \cdot FS_{MIN}} \quad \text{(A9)}$$

$$P_{O\_MAX} = \frac{1}{2} \cdot L_{TYP} \cdot (I_{LIMIT\_MIN}{}^2 - I_{INITIAL}{}^2) \quad \text{(A10)}$$

Average Switching Frequency

Since *LinkSwitch-TN* 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 (A8). Simplifying, we have

$$FS_{AVG} = \frac{2 \cdot V_O \cdot I_O \cdot K_{L\_TOL}}{L \cdot \left(I_{LIMIT}{}^2 - I_{INITIAL}{}^2 \right) K_{LOSS}} \cdot \frac{V_{MIN} - V_{DS} - V_O}{V_{MIN} - V_{DS}}$$

(A11)

Similarly for Buck-Boost converter, simplifying equation (A9) we have

$$FS_{AVG} = \frac{2 \cdot V_O \cdot I_O}{L \cdot \left(I_{LIMIT}{}^2 - I_{INITIAL}{}^2 \right) K_{LOSS}} \cdot \frac{K_{L\_TOL}}{K_{LOSS}}$$

(A12)

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 over heat. Also, RMS currents through the *LinkSwitch-TN* 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

$$I_{INITIAL} = I_{LIMIT\_MIN} - \frac{V_O}{L} \cdot t_{OFF} \quad \text{(A13)}$$

For DCM operation this initial current will be zero.

The current through the *LinkSwitch-TN* as a function of time is given by

$$i_{SW}(t) = I_{INITIAL} + \frac{V_{MIN} - V_{DS} - V_O}{L} \cdot t \text{ , } 0 < t \le t_{ON}$$

$$i_{SW}(t) = 0 \text{ , } t_{ON} < t \le t_{ON} \quad \text{(A14)}$$

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

$$i_D(t) = 0, \quad 0 < t \le t_{ON}$$

$$i_D(t) = I_{LIMIT\_MIN} - \frac{V_O}{L}, t_{ON} < t \le t_{SW} \quad \text{(A15)}$$

$$i_D(t) = 0, \quad I_{LIMIT\_MIN} - \frac{V_O}{L} \cdot t < 0 \quad \text{(A16)}$$

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

$$i_L(t) = i_{SW}(\mathrm{t}) + i_D(t) \quad \text{(A17)}$$

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

$$i_{SW\_RMS} = \sqrt{\frac{1}{T_{AVG}} \int_0^{t_{ON}} i_{SW}(t)^2 \cdot dt} \quad \text{(A18)}$$

$$i_{D\_RMS} = \sqrt{\frac{1}{T_{AVG}} \int_{t_{ON}}^{t_{ON}+t_{OFF}} i_D(t)^2 \cdot dt} \quad \text{(A19)}$$

$$i_{L\_RMS} = \sqrt{\frac{1}{T_{AVG}} \int_0^{T_{AVG}} \left(i_{SW}(t) + i_D(t) \right)^2 \cdot dt} \quad \text{(A20)}$$

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

$$i_{L\_RMS} = \sqrt{i_{SW\_RMS}{}^2 + i_{D\_RMS}{}^2} \quad \text{(A21)}$$

Table A1 lists the design equations for important parameters using the Buck and Buck-Boost topologies.

PARAMETER	BUCK	BUCK-BOOST
L_{TYP}	$L_{TYP} = \dfrac{2 \cdot K_L \cdot \left(\dfrac{V_O \cdot I_O}{K_{L\_LOSS}}\right) \cdot (V_{MIN} - V_{DS} - V_O)}{\left(I_{LIMIT\_MIN}^2 - I_{INITIAL}^2\right) \cdot FS_{MIN} \cdot (V_{MIN} - V_{DS})}$	$L_{TYP} = \dfrac{2 \cdot K_L \cdot \left(\dfrac{V_O \cdot I_O}{K_{L\_LOSS}}\right)}{\left(I_{LIMIT\_MIN}^2 - I_{INITIAL}^2\right) \cdot FS_{MIN}}$
F_{AVG}	$FS_{TYP} = \dfrac{2 \cdot V_O \cdot I_O \cdot K_L}{L \cdot \left(I_{LIMIT}^2 - I_{INITIAL}\right) \cdot K_{L\_LOSS}} \cdot \dfrac{V_{MIN} - V_{DS} - V_O}{V_{MIN} - V_{DS}}$	$FS_{AVG} = \dfrac{2 \cdot V_O \cdot I_O}{L \cdot \left(I_{LIMIT}^2 - I_{INITIAL}^2\right)} \cdot \dfrac{K_L}{K_{L\_LOSS}}$
$i_{SW}(t)$ *LinkSwitch-TN* Current	$i_{SW}(t) = i_{INIT} + \dfrac{V_{MIN} - V_{DS} - V_O}{L} \cdot t \ , t \le t_{ON}$ $i_{SW}(t) = 0 \ , t > t_{ON}$	$i_{SW}(t) = i_{INIT} + \dfrac{V_{MIN} - V_{DS}}{L} \cdot t \ , t \le t_{ON}$ $i_{SW}(t) = 0 \ , t > t_{ON}$
$i_d(t)$ Diode Forward Current	$i_D(t) = I_{LIMIT\_MIN} - \dfrac{V_O}{L} \cdot t \ , t > t_{ON}$ $i_D(t) = 0 \ , I_{LIMIT\_MIN} - \dfrac{V_O}{L} \cdot t < 0$ $i_D(t) = 0 \ , t \le t_{ON}$	$i_D(t) = I_{LIMIT\_MIN} - \dfrac{V_O}{L} \cdot t \ , t > t_{ON}$ $i_D(t) = 0 \ , I_{LIMIT\_MIN} - \dfrac{V_O}{L} \cdot t < 0$ $i_D(t) = 0 \ , t \le t_{ON}$
$i_L(t)$ Inductor Current	$i_L(t) = i_{SW}(t) + i_D(t)$	$i_L(t) = i_{SW}(t) + i_D(t)$
Max Drain Voltage	V_{MAX}	$V_{MAX} + V_O$

Table A1. Circuit Characteristics for Buck and Buck-Boost Topologies.

For the latest updates, visit our Web site: *www.powerint.com*

PATENT INFORMATION

Power Integrations reserves the right to make changes to its products at any time to improve reliability or manufacturability. Power Integrations does not assume any liability arising from the use of any device or circuit described herein, nor does it convey any license under its patent rights or the rights of others.

The products and applications illustrated herein (including circuits external to the products and transformer construction) may be covered by one or more U.S. and foreign patents or potentially by pending U.S. and foreign patent applications assigned to Power Integrations. A complete list of Power Integrations' patents may be found at www.powerint.com.

LIFE SUPPORT POLICY

POWER INTEGRATIONS' PRODUCTS ARE NOT AUTHORIZED FOR USE AS CRITICAL COMPONENTS IN LIFE SUPPORT DEVICES OR SYSTEMS WITHOUT THE EXPRESS WRITTEN APPROVAL OF THE PRESIDENT OF POWER INTEGRATIONS. As used herein:

1. Life support devices or systems which, (a) are intended for surgical implant into the body, or (b) support or sustain life, and whose failure to perform, when properly used in accordance with instructions for use provided in the labeling, can be reasonably expected to result in a significant injury to the user.

2. A critical component is any component of a life support device or system whose failure to perform can be reasonably expected to cause the failure of the life support device or system, or to affect its safety or effectiveness.

WORLD HEADQUARTERS
Power Integrations
5245 Hellyer Avenue
San Jose, CA 95138, USA.
Main: +1-408-414-9200
Customer Service:
Phone: +1-408-414-9665
Fax: +1-408-414-9765
e-mail: usasales@powerint.com

CHINA (SHENZHEN)
Power Integrations
International Holdings, Inc.
Rm# 1705, Bao Hua Bldg.
1016 Hua Qiang Bei Lu
Shenzhen Guangdong,
518031, China
Phone: +86-755-8367-5143
Fax: +86-755-8377-9610
e-mail: chinasales@powerint.com

ITALY
Power Integrations S.r.l.
Via Vittorio Veneto 12,
Bresso
Milano, 20091, Italy
Phone: +39-028-928-6001
Fax: +39-028-928-6009
e-mail: eurosales@powerint.com

SINGAPORE (ASIA PACIFIC HEADQUARTERS)
Power Integrations, Singapore
51 Newton Road
#15-08/10 Goldhill Plaza
Singapore, 308900
Phone: +65-6358-2160
Fax: +65-6358-2015
e-mail: singaporesales@powerint.com

AMERICAS
Power Integrations
4335 South Lee Street,
Suite G
Buford, GA 30518, USA
Phone: +1-678-714-6033
Fax: +1-678-714-6012
e-mail: usasales@powerint.com

GERMANY
Power Integrations GmbH
Rueckerstrasse 3
D-80336, Muenchen, Germany
Phone: +49-895-527-3910
Fax: +49-895-527-3920
e-mail: eurosales@powerint.com

JAPAN
Power Integrations, K.K.
Keihin-Tatemono 1st Bldg.
12-20 Shin-Yokohama
2-Chome,
Kohoku-ku, Yokohama-shi,
Kanagawa 222-0033, Japan
Phone: +81-45-471-1021
Fax: +81-45-471-3717
e-mail: japansales@powerint.com

TAIWAN
Power Integrations
International Holdings, Inc.
5F-1, No. 316, Nei Hu Rd., Sec. 1
Nei Hu Dist.
Taipei, Taiwan 114, R.O.C.
Phone: +886-2-2659-4570
Fax: +886-2-2659-4550
e-mail: taiwansales@powerint.com

CHINA (SHANGHAI)
Power Integrations
International Holdings, Inc.
Rm 807, Pacheer
Commercial Centre
555 Nanjing West Road
Shanghai, 200041, China
Phone: +86-21-6215-5548
Fax: +86-21-6215-2468
e-mail: chinasales@powerint.com

INDIA (TECHNICAL SUPPORT)
Innovatech
261/A, Ground Floor
7th Main, 17th Cross,
Sadashivanagar
Bangalore 560080
Phone: +91-80-5113-8020
Fax: +91-80-5113-8023
e-mail: indiasales@powerint.com

KOREA
Power Integrations
International Holdings, Inc.
8th Floor, DongSung Bldg.
17-8 Yoido-dong,
Youngdeungpo-gu,
Seoul, 150-874, Korea
Phone: +82-2-782-2840
Fax: +82-2-782-4427
e-mail: koreasales@powerint.com

UK (EUROPE & AFRICA HEADQUARTERS)
Power Integrations (Europe) Ltd.
1st Floor, St. James's House
East Street
Farnham
Surrey
GU9 7TJ
United Kingdom
Phone: +44 (0) 1252-730-140
Fax: +44 (0) 1252-727-689
e-mail: eurosales@powerint.com

APPLICATIONS HOTLINE
World Wide +1-408-414-9660

APPLICATIONS FAX
World Wide +1-408-414-9760