HIGH VOLTAGE SYNCHRONOUS PWM BUCK CONTROLLER

Features

- High Voltage Operating up to 100V
- Programmable Switching Frequency up to 400kHz
- 1A Output Drive Capability
- Precision Reference Voltage (1.25V)
- Programmable Soft-Start
- Programmable Over Current Protection
- Hiccup Current Limit Using MOSFET $R_{DS(on)}$ sensing
- External Frequency Synchronization
- 14-pin SOIC Package

Description

The IR3651 is a high voltage PWM controller designed for high performance synchronous Buck DC/DC applications. The IR3651 drives a pair of external N-MOSFETs using a programmable switching frequency up to 400kHz allows flexibility to tune the operation of the IC to meet system level requirements, and synchronization allows the simplification of system level filter design. The output voltage can be precisely regulated using the internal 1.25V reference voltage for low voltage applications. Protection such as under voltage lockout and hiccup current limit are provided to give required system level security in the event of fault conditions.

Applications

- 48V non-isolated DC to DC Converter
- Embedded Telecom Systems
- Networking and Computing Voltage Regulator
- Distributed Point of Load Power Architectures
- General high voltage DC/DC Converters

Typical application Circuit

ORDERING INFORMATION

<table>
<thead>
<tr>
<th>PKG</th>
<th>PACKAGE DESCRIPTION</th>
<th>PIN COUNT</th>
<th>PARTS PER TUBE</th>
<th>PARTS PER REEL</th>
<th>T&amp;R ORIENTATION</th>
</tr>
</thead>
<tbody>
<tr>
<td>S</td>
<td>IR3651SPBF</td>
<td>14</td>
<td>55</td>
<td>2500</td>
<td>Fig A</td>
</tr>
<tr>
<td>S</td>
<td>IR3651STRPBF</td>
<td>14</td>
<td>155</td>
<td>2500</td>
<td></td>
</tr>
</tbody>
</table>
ABSOLUTE MAXIMUM RATINGS
(Voltages referenced to GND)

- Vcc, DRVcc Supply Voltage .................................. -0.3V to 20V
- Vs Supply Voltage ............................................. -0.3V to 150V
- Vb Supply Voltage ............................................. -0.3V to Vs+20V
- OCset .......................................................... 10mA
- Storage Temperature Range ..................................... -65°C To 150°C
- Operating Junction Temperature Range ................... -40°C To 150°C
- ESD Classification ............................................... JEDEC, JESD22-A114 (1K)
- Moisture Sensitivity Level ..................................... JEDEC Level 2 @ 260°C

CAUTION: Stresses beyond those listed under “Absolute Maximum Ratings” may cause permanent damage to
the device. These are stress ratings only and functional operation of the device at these or any other conditions
beyond those indicated in the operational sections of the specifications are not implied.

Package Information

14-Pin SOIC NB (S)

\[ \Theta_{JA} = 88.2^\circ \text{C/W} \]
\[ \Theta_{JC} = 37^\circ \text{C/W} \]
Fig. 1: Simplified block diagram of the IR3651
## Pin Description

<table>
<thead>
<tr>
<th>Pin</th>
<th>Name</th>
<th>Description</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>Fb</td>
<td>Inverting input to the error amplifier. This pin is connected directly to the output of the regulator via resistor divider to set the output voltage and provide feedback to the error amplifier.</td>
</tr>
<tr>
<td>2</td>
<td>Comp</td>
<td>Output of error amplifier. An external resistor and capacitor network is typically connected from this pin to ground to provide loop compensation.</td>
</tr>
<tr>
<td>3</td>
<td>SS/SD</td>
<td>Soft start / shutdown. This pin provides user programmable soft-start function. Connect an external capacitor from this pin to ground to set the start up time of the output voltage. The converter can be shutdown by pulling this pin below 0.3V.</td>
</tr>
<tr>
<td>4</td>
<td>SYNC</td>
<td>The internal oscillator can be synchronized to an external clock via this pin.</td>
</tr>
<tr>
<td>5</td>
<td>PGnd</td>
<td>Power Ground. This pin serves as a separate ground for the MOSFET driver and should be connected to the system's power ground plane.</td>
</tr>
<tr>
<td>6</td>
<td>LDrv</td>
<td>Output driver for low side MOSFET.</td>
</tr>
<tr>
<td>7</td>
<td>DRVcc</td>
<td>This pin provides biasing for the internal low side driver. A minimum of 0.1uF, high frequency capacitor must be connected from this pin to power ground.</td>
</tr>
<tr>
<td>8</td>
<td>Vb</td>
<td>This pin powers the high side driver and must be connected to a voltage higher than bus voltage. A minimum of 0.1uF, high frequency capacitor must be connected from this pin to switch node.</td>
</tr>
<tr>
<td>9</td>
<td>HDrv</td>
<td>Output driver for high side MOSFET</td>
</tr>
<tr>
<td>10</td>
<td>Vs</td>
<td>Switch node. Connect this pin to the source of the upper MOSFET and the drain of the lower MOSFET. This pin is return path for the upper gate driver.</td>
</tr>
<tr>
<td>11</td>
<td>Vcc</td>
<td>This pin provides power for the internal blocks of the IC. A minimum of 0.1uF, high frequency capacitor must be connected from this pin to ground.</td>
</tr>
<tr>
<td>12</td>
<td>OCSet</td>
<td>Current limit set point. A resistor from this pin to drain of low side MOSFET will set the current limit threshold.</td>
</tr>
<tr>
<td>13</td>
<td>Gnd</td>
<td>Signal ground for internal reference and control circuitry.</td>
</tr>
<tr>
<td>14</td>
<td>Rt</td>
<td>Connecting a resistor from this pin to ground sets the oscillator frequency.</td>
</tr>
</tbody>
</table>
## Recommended Operating Conditions

<table>
<thead>
<tr>
<th>Symbol</th>
<th>Definition</th>
<th>Min</th>
<th>Max</th>
<th>Unit</th>
</tr>
</thead>
<tbody>
<tr>
<td>$V_{bus}$</td>
<td>Converting Voltage</td>
<td>12</td>
<td>100</td>
<td>V</td>
</tr>
<tr>
<td>$V_{cc}$</td>
<td>Supply Voltage</td>
<td>4.5</td>
<td>13.2</td>
<td>V</td>
</tr>
<tr>
<td>DRV$_{cc}$</td>
<td>Supply Voltage</td>
<td>10</td>
<td>16</td>
<td>V</td>
</tr>
<tr>
<td>$V_{o}$ to $V_{s}$</td>
<td>Supply Voltage</td>
<td>10</td>
<td>16</td>
<td>V</td>
</tr>
<tr>
<td>$F_s$</td>
<td>Operating Frequency</td>
<td>100</td>
<td>400</td>
<td>kHz</td>
</tr>
<tr>
<td>$T_j$</td>
<td>Junction Temperature</td>
<td>-40</td>
<td>125</td>
<td>°C</td>
</tr>
</tbody>
</table>

## Electrical Specifications

Unless otherwise specified, these specifications apply over $V_{cc}=5V$; DRV$_{cc}$=Vb=12V, 0°C<$T_j<$125°C

### Reference Voltage

<table>
<thead>
<tr>
<th>Parameter</th>
<th>SYM</th>
<th>Test Condition</th>
<th>Min</th>
<th>TYP</th>
<th>MAX</th>
<th>Unit</th>
</tr>
</thead>
<tbody>
<tr>
<td>Feedback Voltage</td>
<td>$V_{FB}$</td>
<td></td>
<td>1.25</td>
<td></td>
<td></td>
<td>V</td>
</tr>
<tr>
<td>Accuracy</td>
<td></td>
<td>0°C&lt;$T_j&lt;$125°C</td>
<td>-1.5</td>
<td>+1.5</td>
<td></td>
<td>%</td>
</tr>
<tr>
<td>Fb Voltage Line Regulation</td>
<td>LREG</td>
<td>5V&lt;$V_{cc}$&lt;12V</td>
<td>2.0</td>
<td></td>
<td></td>
<td>mV</td>
</tr>
</tbody>
</table>

### Supply Current

<table>
<thead>
<tr>
<th>Parameter</th>
<th>SYM</th>
<th>Test Condition</th>
<th>Min</th>
<th>TYP</th>
<th>MAX</th>
<th>Unit</th>
</tr>
</thead>
<tbody>
<tr>
<td>$V_{cc}$ Supply Current (Stat)</td>
<td>$I_{CC(Static)}$</td>
<td>SS=0V, No Switching</td>
<td>6</td>
<td>7</td>
<td></td>
<td>mA</td>
</tr>
<tr>
<td>$V_{cc}$ Supply Current (Dyn)</td>
<td>$I_{CC(Dynamic)}$</td>
<td>$F_s$=200kHz, $C_{LOAD}$=1.5nF</td>
<td>6</td>
<td>7</td>
<td></td>
<td>mA</td>
</tr>
<tr>
<td>DRV$_{cc}$ Supply Current (Stat)</td>
<td>$I_{C(Static)}$</td>
<td>SS=0V, No Switching</td>
<td>0.3</td>
<td>0.5</td>
<td></td>
<td>mA</td>
</tr>
<tr>
<td>DRV$_{cc}$ Supply Current (Dyn)</td>
<td>$I_{C(Dynamic)}$</td>
<td>$F_s$=200kHz, $C_{LOAD}$=1.5nF</td>
<td>4</td>
<td>5</td>
<td></td>
<td>mA</td>
</tr>
<tr>
<td>Vb Supply Current (Stat)</td>
<td>$I_{b(Static)}$</td>
<td>SS=0V, No Switching</td>
<td>0.3</td>
<td>0.5</td>
<td></td>
<td>mA</td>
</tr>
<tr>
<td>Vb Supply Current (Dyn)</td>
<td>$I_{b(Dynamic)}$</td>
<td>$F_s$=200kHz, $C_{LOAD}$=1.5nF</td>
<td>4.5</td>
<td>5.5</td>
<td></td>
<td>mA</td>
</tr>
</tbody>
</table>

### Under Voltage Lockout

<table>
<thead>
<tr>
<th>Parameter</th>
<th>SYM</th>
<th>Test Condition</th>
<th>Min</th>
<th>TYP</th>
<th>MAX</th>
<th>Unit</th>
</tr>
</thead>
<tbody>
<tr>
<td>$V_{cc}$-Start-Threshold</td>
<td>$V_{CC_UVLO(R)}$</td>
<td>Supply ramping up</td>
<td>4.0</td>
<td>4.17</td>
<td>4.35</td>
<td>V</td>
</tr>
<tr>
<td>$V_{cc}$-Stop-Threshold</td>
<td>$V_{CC_UVLO(F)}$</td>
<td>Supply ramping down</td>
<td>3.75</td>
<td>4.1</td>
<td></td>
<td>V</td>
</tr>
<tr>
<td>$V_{cc}$-Hysteresis</td>
<td></td>
<td>Supply ramping up and down</td>
<td>0.15</td>
<td>0.25</td>
<td>0.3</td>
<td>V</td>
</tr>
<tr>
<td>DRV$_{cc}$-Start-Threshold</td>
<td>$DRV_{cc}$-UVLO(R)</td>
<td>Supply ramping up</td>
<td>8.3</td>
<td>9</td>
<td>9.7</td>
<td>V</td>
</tr>
<tr>
<td>DRV$_{cc}$-Stop-Threshold</td>
<td>$DRV_{cc}$-UVLO(F)</td>
<td>Supply ramping down</td>
<td>7.5</td>
<td>8.2</td>
<td>8.9</td>
<td>V</td>
</tr>
<tr>
<td>DRV$_{cc}$-Hysteresis</td>
<td></td>
<td>Supply ramping up and down</td>
<td>0.6</td>
<td>0.9</td>
<td></td>
<td>V</td>
</tr>
<tr>
<td>$V_b$-Start-Threshold</td>
<td>$V_b$-UVLO(R)</td>
<td>Supply ramping up</td>
<td>8.3</td>
<td>9</td>
<td>9.7</td>
<td>V</td>
</tr>
<tr>
<td>$V_b$-Stop-Threshold</td>
<td>$V_b$-UVLO(F)</td>
<td>Supply ramping down</td>
<td>7.5</td>
<td>8.2</td>
<td>8.9</td>
<td>V</td>
</tr>
<tr>
<td>$V_b$-Hysteresis</td>
<td></td>
<td>Supply ramping up and down</td>
<td>0.6</td>
<td></td>
<td>0.9</td>
<td>V</td>
</tr>
</tbody>
</table>

### Oscillator

<table>
<thead>
<tr>
<th>Parameter</th>
<th>SYM</th>
<th>Test Condition</th>
<th>Min</th>
<th>TYP</th>
<th>MAX</th>
<th>Unit</th>
</tr>
</thead>
<tbody>
<tr>
<td>Frequency</td>
<td>$F_s$</td>
<td>$R_t$=120K $R_t$=51K</td>
<td>170</td>
<td>200</td>
<td>230</td>
<td>kHz</td>
</tr>
<tr>
<td>Ramp Amplitude</td>
<td>$V_{ramp}$</td>
<td>Note2</td>
<td>1.25</td>
<td></td>
<td></td>
<td>V</td>
</tr>
<tr>
<td>Min Duty Cycle</td>
<td>$D_{min}$</td>
<td>$F_b$=2V</td>
<td>0</td>
<td></td>
<td></td>
<td>%</td>
</tr>
<tr>
<td>Min Pulse Width</td>
<td>$D_{min(crit)}$</td>
<td>$F_s$=200kHz, Note2</td>
<td>200</td>
<td></td>
<td></td>
<td>ns</td>
</tr>
<tr>
<td>Max Duty Cycle</td>
<td>$D_{max}$</td>
<td>$F_s$=200kHz, $F_b$=1.2V $F_s$=400kHz, $F_b$=1.2V</td>
<td>80</td>
<td>70</td>
<td></td>
<td>%</td>
</tr>
<tr>
<td>Sync Frequency Range</td>
<td>$Sync(F_s)$</td>
<td>20% above free running freq</td>
<td></td>
<td></td>
<td>480</td>
<td>kHz</td>
</tr>
<tr>
<td>Sync Pulse Duration</td>
<td>$Sync(puls)$</td>
<td></td>
<td>200</td>
<td></td>
<td></td>
<td>ns</td>
</tr>
<tr>
<td>Sync high Level Threshold</td>
<td>$Sync(H)$</td>
<td></td>
<td>2</td>
<td></td>
<td></td>
<td>V</td>
</tr>
<tr>
<td>Sync low Level Threshold</td>
<td>$Sync(L)$</td>
<td></td>
<td>0.8</td>
<td></td>
<td></td>
<td>V</td>
</tr>
<tr>
<td>Parameter</td>
<td>SYM</td>
<td>Test Condition</td>
<td>Min</td>
<td>TYP</td>
<td>MAX</td>
<td>Units</td>
</tr>
<tr>
<td>---------------------------------</td>
<td>---------</td>
<td>--------------------</td>
<td>-----</td>
<td>------</td>
<td>------</td>
<td>-----------</td>
</tr>
<tr>
<td><strong>Error Amplifier</strong></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Input Bias Current</td>
<td>I_{FB}</td>
<td>SS=3V, Fb=1V</td>
<td>-0.1</td>
<td>-0.4</td>
<td></td>
<td>μA</td>
</tr>
<tr>
<td>Source/Sink Current</td>
<td>I(source/Sink)</td>
<td>50</td>
<td>85</td>
<td>120</td>
<td>μA</td>
<td></td>
</tr>
<tr>
<td>Transconductance</td>
<td>gm</td>
<td>1500</td>
<td>2400</td>
<td>3000</td>
<td></td>
<td>μmho</td>
</tr>
<tr>
<td><strong>Soft Start/SD</strong></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Soft Start Current</td>
<td>I_{SS}</td>
<td>SS=0V</td>
<td>15</td>
<td>20</td>
<td>25</td>
<td>μA</td>
</tr>
<tr>
<td>Shutdown Output Threshold</td>
<td>SD</td>
<td></td>
<td></td>
<td></td>
<td>0.25</td>
<td>V</td>
</tr>
<tr>
<td><strong>Over Current Protection</strong></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>OCSET Current</td>
<td>I_{OCSET}</td>
<td>7.5</td>
<td>10</td>
<td>12.5</td>
<td>μA</td>
<td></td>
</tr>
<tr>
<td>Hiccup Current</td>
<td>I_{Hiccup}</td>
<td>3</td>
<td></td>
<td></td>
<td>μA</td>
<td></td>
</tr>
<tr>
<td>Hiccup Duty Cycle</td>
<td>I_{Hiccup} / I_{SS}</td>
<td>Note2</td>
<td>5</td>
<td>%</td>
<td></td>
<td></td>
</tr>
<tr>
<td><strong>Output Drivers</strong></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>LO, Drive Rise Time</td>
<td>Tr(Lo)</td>
<td>CL=1.5nF</td>
<td>10</td>
<td>20</td>
<td></td>
<td>ns</td>
</tr>
<tr>
<td></td>
<td></td>
<td>See Fig 2, Note2</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>HI Drive Rise Time</td>
<td>Tr(Hi)</td>
<td>CL=1.5nF</td>
<td>10</td>
<td>20</td>
<td></td>
<td>ns</td>
</tr>
<tr>
<td></td>
<td></td>
<td>See Fig 2, Note2</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>LO Drive Fall Time</td>
<td>Tf(Lo)</td>
<td>CL=1.5nF</td>
<td>10</td>
<td>20</td>
<td></td>
<td>ns</td>
</tr>
<tr>
<td></td>
<td></td>
<td>See Fig 2, Note2</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>HI Drive Fall Time</td>
<td>Tf(Hi)</td>
<td>CL=1.5nF</td>
<td>10</td>
<td>20</td>
<td></td>
<td>ns</td>
</tr>
<tr>
<td></td>
<td></td>
<td>See Fig 2, Note2</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Dead Band Time</td>
<td>T_dead</td>
<td>See Fig 2</td>
<td>30</td>
<td>45</td>
<td>100</td>
<td>ns</td>
</tr>
<tr>
<td>Upper Driver Source Current</td>
<td>I_{upper(source)}</td>
<td>HDRv short circuit pulsed current, PW&lt;10us</td>
<td>1.0</td>
<td>A</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Upper Driver Sink Current</td>
<td>I_{upper(sink)}</td>
<td>HDRv short circuit pulsed current, PW&lt;10us</td>
<td>1.0</td>
<td>A</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Lower Driver Source Current</td>
<td>I_{lower(source)}</td>
<td>LDrv short circuit pulsed current, PW&lt;10us</td>
<td>1.0</td>
<td>A</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Lower Driver Sink Current</td>
<td>I_{lower(sink)}</td>
<td>LDrv short circuit pulsed current, PW&lt;10us</td>
<td>1.0</td>
<td>A</td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

Note1: Cold temperature performance is guaranteed via correlation using statistical quality control. Not tested in production.

Note2: Guaranteed by Design but not tested in production.

![Fig. 2: Definition of Rise/Fall time and Deadband Time](image-url)
TYPICAL OPERATING CHARACTERISTICS (-40°C TO +125°C)

- **V_{fb}**
- **Transconductance**
- **I_{cc}(dynamic)**
- **I_{cc}(static)**
- **I_{b}(dynamic)**
- **I_{c}(dynamic)**
- **V_{b_{UVLO}}**
- **DRVcc_{UVLO}**
TYPICAL OPERATING CHARACTERISTICS (-40°C TO +125°C)

- Vcc_UVLO
- Frequency RT=120K
- Dead time
- Max DC @ 200KHz
- Iss
- Iocset
Circuit Description

THEORY OF OPERATION

Introduction
The IR3651 is a voltage mode PWM synchronous controller. The output voltage is set by feedback pin (Fb) and the internal reference voltage (1.25V). These are two inputs to error amplifier. The error signal between these two inputs is compared to a fixed frequency linear sawtooth ramp and generates fixed frequency pulses of variable duty-cycle (D) which drives N-channel external MOSFETs.

The timing of the IC is controlled by an internal oscillator circuit that can be externally programmed up to 400kHz.

The IR3651 operates with a wide input voltage from 12V to 100V allowing an extended operating input voltage range.

The current limit is programmable and uses on-resistance of the low-side MOSFET, eliminating the need for an external current sense resistor.

Under-Voltage Lockout
The under-voltage lockout circuit monitors the Vcc supply and assures that the IC doesn’t starts until the Vcc reaches the set threshold. Lockout occurs if Vcc falls below 4.1V. Normal operation resumes once Vcc rises above the set value.

Shutdown
The output can be shutdown by pulling the soft-start pin below 0.3V. This can be easily done by using an external small signal transistor. During shutdown both MOSFET drivers will be turned off. Normal operation will resume by cycling soft start pin.

Error Amplifier
The IR3651 is a voltage mode controller. The error amplifier is of transconductance type. The amplifier is capable of operating with Type III compensation control scheme using low ESR output capacitance.

Operating Frequency Selection
The switching frequency is determined by connecting an external resistor (Rt) to ground. Figure 3 provides a graph of oscillator frequency versus Rt.

Frequency Synchronization
The IR3651 is capable of accepting an external digital synchronization signal. Synchronization will be enabled by the rising edge at an external clock. Switching frequency is set by external resistor (Rt). During synchronization, Rt is selected such that the free running frequency is 20% below the synchronization frequency. When unused, the sync pin will remain floating and is noise immune.

Pre-Bias Startup
IR3651 is able to start up into pre-charged output, which prevents oscillation and disturbances of the output voltage.

Depends on system configuration, specific amount of output capacitors may be required to prevent discharging the output voltage. 
Soft-Start

The IR3651 has programmable soft-start to control the output voltage rise and limit the inrush current during start-up.

To ensure correct start-up, the soft-start sequence initiates when Vcc rises above its threshold and generate the Power On Ready (POR) signal. The soft-start function operates by sourcing current to charge an external capacitor to about 3V.

Initially, the soft-start function clamps the output of error amplifier by injecting a current (64uA) into the Fb pin and generates a voltage about 1.6V (64ux25K) across the negative input of error amplifier (see figure 4).

The magnitude of the injected current is inversely proportional to the voltage at the soft-start pin. As the soft-start voltage ramps up, the injected current decreases linearly and so does the voltage at negative input of error amplifier.

When the soft-start capacitor is around 1V, the voltage at the positive input of the error amplifier is approximately 1.25V.

The output of error amplifier will start increasing and generating the first PWM signal. As the soft-start capacitor voltage continues to go up, the current flowing into the Fb pin will keep decreasing.

The feedback voltage increases linearly as the soft start voltage ramps up. When soft-start voltage is around 2V the output voltage is reached the steady state and the injected current is zero.

Figure 5 shows the theoretical operational waveforms during soft-start.

The output voltage start-up time is the time period when soft-start capacitor voltage increases from 1V to 2V.

The start-up time will be dependent on the size of the external soft-start capacitor and can be estimate by:

\[ 20\mu A \times \frac{T_{\text{start}}}{C_{ss}} = 2V - 1V \]

For a given start-up time, the soft-start capacitor (nF) can be estimated as:

\[ C_{ss} = 20 \mu A \times T_{\text{start}}(ms) \]
Over-Current Protection

The over current protection is performed by sensing current through the R\_DS(on) of low side MOSFET. This method enhances the converter's efficiency and reduce cost by eliminating a current sense resistor. As shown in figure 6, an external resistor (R\_SET) is connected between OCS\_Set pin and the drain of low side MOSFET (Q2) which sets the current limit set point.

The internal current source develops a voltage across R\_SET. When the low side MOSFET is turned on, the inductor current flows through the Q2 and results a voltage which is given by:

\[ V_{OC\_Set} = (I_{OC\_Set} \times R_{OC\_Set}) - (R_{DS\_on} \times I_i) \]  --(2)

![Fig. 6: Connection of over current sensing resistor](image)

The critical inductor current can be calculated by setting:

\[ V_{OC\_Set} = (I_{OC\_Set} \times R_{OC\_Set}) - (R_{DS\_on} \times I_i) = 0 \]

\[ I_{SET} = I_{L(\text{init} \text{ia})} = \frac{R_{DS\_on} \times I_{OC\_Set}}{R_{OC\_Set}} \]  --(3)

\[ I_{SET} = I_{L(\text{init} \text{ia})} = 1.5 \times I_i + \frac{\Delta i}{2} \]

An over current is detected if the OCS\_Set pin goes below ground. This trips the OCP comparator and cycles the soft start function in hiccup mode.

The hiccup is performed by charging and discharging the soft-start capacitor in certain slope rate. As shown in figure 7 a 3\( \mu \)A current source is used to discharge the soft-start capacitor.

The OCP comparator resets after every soft start cycles, the converter stays in this mode until the overload or short circuit is removed. The converter will automatically recover.

![Fig. 7: 3\( \mu \)A current source for discharging soft-start capacitor during hiccup](image)

![Fig. 8: OCS\_Set pin during normal condition](image)

The OCP circuit starts sampling current approximately 200ns before the low gate drive turns off. The OCS\_Set pin is internally clamped to about 1.5V during deadtime to prevent false trigging, figure 8 shows the OCS\_Set pin during one switching cycle.

The value of R\_SET should be checked in an actual circuit to ensure that the over current protection circuit activates as expected. The IR3651 current limit is designed primarily as disaster preventing, "no blow up" circuit, and doesn't operate as a precision current regulator.
Application Information

Output Voltage Programming

Output voltage is programmed by reference voltage and external voltage divider. The Fb pin is the inverting input of the error amplifier, which is internally referenced to 1.25V. The divider is ratioed to provide 1.25V at the Fb pin when the output is at its desired value. The output voltage is defined by using the following equation:

When an external resistor divider is connected to the output as shown in figure 9.

\[ V_o = V_{ref} \left( 1 + \frac{R_8}{R_9} \right) \]  \hspace{0.5cm} -(4)

Equation (4) can be rewritten as:

\[ R_9 = R_9 \cdot \left( \frac{V_{ref}}{V_0 - V_{ref}} \right) \]  \hspace{0.5cm} -(5)

For the calculated values of R_8 and R_9 see feedback compensation section.

Soft-Start Programming

The soft-start timing can be programmed by selecting the soft-start capacitance value. The start-up time of the converter can be calculated by using:

\[ C_{ss} \approx 20 \mu \text{A} \cdot T_{\text{start}} \]  \hspace{0.5cm} -(f)

Where T_{start} is the desired start-up time (ms)
For a start-up time of 5ms, the soft-start capacitor will be 0.1uF. Choose a ceramic capacitor at 0.1uF.
Input Capacitor Selection
The input filter capacitor should be selected based on how much ripple the supply can tolerate on the DC input line. The ripple current generated during the on time of upper MOSFET should be provided by input capacitor. The RMS value of this ripple is expressed by:

\[ I_{\text{RMS}} = I_o \sqrt{D(1-D)} \quad (7) \]

Where:
- \( D \) is the Duty Cycle
- \( I_{\text{RMS}} \) is the RMS value of the input capacitor current.
- \( I_o \) is the output current.

For applications with input supplies above 30V, choice of input capacitor type is limited to ceramics or aluminum electrolytics. Ceramic capacitors offer high peak current capabilities, they also feature low ESR and ESL at higher frequency which enhance better efficiency, however high voltage ceramic capacitors are available with only in low value capacitance. A combination of ceramic capacitors and electrolytic capacitors are recommended.

Inductor Selection
The inductor is selected based on output power, operating frequency and efficiency requirements. Low inductor value causes large ripple current, resulting in the smaller size, faster response to a load transient but poor efficiency and high output noise. Generally, the selection of inductor value can be reduced to desired maximum ripple current in the inductor \((\Delta I)\) . The optimum point is usually found between 20% and 50% ripple of the output current.

For the buck converter, the inductor value for desired operating ripple current can be determined using the following relation:

\[ V_o - V_i = L \frac{\Delta I}{\Delta t}; \quad \Delta t = D \frac{1}{F_s} \]

\[ L = \frac{(V_o - V_i) \cdot V_i}{V_o - \Delta I \cdot F_s} \quad (8) \]

Where:
- \( V_i \) = Maximum input voltage
- \( V_o \) = Output Voltage
- \( \Delta I \) = Inductor ripple current
- \( F_s \) = Switching frequency
- \( \Delta t \) = Turn on time
- \( D \) = Duty cycle

Output Capacitor Selection
The voltage ripple and transient requirements determine the output capacitors types and values. The criteria is normally based on the value of the Effective Series Resistance (ESR). However the actual capacitance value and the Equivalent Series Inductance (ESL) are other contributing components, these components can be described as:

\[ \Delta V_s = \Delta V_{\text{(ESR)}} + \Delta V_{\text{(ESL)}} + \Delta V_{\text{(C)}} \]

\[ \Delta V_{\text{(ESR)}} = \Delta I \cdot ESR \quad (9) \]

\[ \Delta V_{\text{(ESL)}} = \frac{(V_o \cdot \Delta I)}{L} \cdot ESL \]

\[ \Delta V_{\text{(C)}} = \frac{\Delta I}{8 \cdot C \cdot F_s} \]

\[ \Delta V_s = \text{Output voltage ripple} \]

\[ \Delta I = \text{Inductor ripple current} \]

Since the output capacitor has major role in overall performance of converter and determines the result of transient response, selection of capacitor is critical. The IR3651 can perform well with all types of capacitors.

As a rule the capacitor must have low enough ESR to meet output ripple and load transient requirements, yet have high enough ESR to satisfy stability requirements.

The goal for this design is to meet the voltage ripple requirement in smallest possible capacitor size. Therefore ceramic capacitor is selected due to low ESR and small size.

In the case of tantalum or low ESR electrolytic capacitors, the ESR dominates the output voltage ripple, equation (9) can be used to calculate the required ESR for the specific voltage ripple.
Power MOSFET Selection

The IR3651 uses two N-Channel MOSFETs. The selection criteria to meet power transfer requirements is based on maximum drain-source voltage ($V_{DSS}$), gate-source drive voltage ($V_{gs}$), maximum output current, On-resistance $R_{DS(on)}$ and thermal management.

The MOSFET must have a maximum operating voltage ($V_{DSS}$) exceeding the maximum input voltage ($V_{in}$).

The gate drive requirement is almost the same for both MOSFETs. Logic-level transistor can be used and caution should be taken with devices at very low $V_{gs}$ to prevent undesired turn-on of the complementary MOSFET, which results in shoot-through current.

The total power dissipation for MOSFETs includes conduction and switching losses. For the Buck converter the average inductor current is equal to the DC load current. The conduction loss is defined as:

$$P_{cond} = (upper\ switch) = \frac{1}{2} I_{load} R_{DS(on)} D + \theta$$

$$P_{cond} = (lower\ switch) = \frac{1}{2} I_{load} R_{DS(on)} (1-D) + \theta$$

$\theta = R_{DS(on)}$ temperature dependency

The $R_{DS(on)}$ temperature dependency should be considered for the worst case operation. This is typically given in the MOSFET data sheet. Ensure that the conduction losses and switching losses do not exceed the package ratings or violate the overall thermal budget.

The switching loss is more difficult to calculate, even though the switching transition is well understood. The reason is the effect of the parasitic components and switching times during the switching procedures such as turn-on / turn-off delays and rise and fall times. The control MOSFET contributes to the majority of the switching losses in synchronous Buck converter. The synchronous MOSFET turns on under zero voltage conditions, therefore, the turn on losses for synchronous MOSFET can be neglected. With a linear approximation, the total switching loss can be expressed as:

$$P_{sw} = \frac{V_{ds(off)} t_r + t_f I_{load}}{T}$$

Where:

$V_{ds(off)}$ = Drain to source voltage at the off time
$t_r$ = Rise time
$t_f$ = Fall time
$T$ = Switching period
$I_{load}$ = Load current

The switching time waveforms is shown in figure 10.

$$P_{sw} = \frac{V_{ds(off)} t_r + t_f I_{load}}{T}$$

Where:

$V_{ds(off)}$ = Drain to source voltage at the off time
$t_r$ = Rise time
$t_f$ = Fall time
$T$ = Switching period
$I_{load}$ = Load current

![Fig. 10: switching time waveforms](image-url)
Feedback Compensation

The IR3651 is a voltage mode controller; the control loop is a single voltage feedback path including error amplifier and error comparator. To achieve fast transient response and accurate output regulation, a compensation circuit is necessary. The goal of the compensation network is to provide a closed loop transfer function with the highest 0dB crossing frequency and adequate phase margin (greater than 45°).

The output LC filter introduces a double pole, –40dB/decade gain slope above its corner resonant frequency, and a total phase lag of 180° (see figure 11). The resonant frequency of the LC filter expressed as follows:

\[ F_{LC} = \frac{1}{2\pi \sqrt{L_o \cdot C_o}} \quad -(11) \]

Figure 11 shows gain and phase of the LC filter. Since we already have 180° phase shift just from the output filter, the system risks being unstable.

The ESR zero of the output capacitor is expressed as follows:

\[ F_{ESR} = \frac{1}{2\pi \cdot \text{ESR} \cdot C_o} \quad -(12) \]

Fig. 12: Typell compensation network and its asymptotic gain plot

The transfer function \((V_e/V_o)\) is given by:

\[ H(s) = \left( g_m \cdot \frac{R_2}{R_s + R_2} \right) \cdot \frac{1 + sR_sC_s}{sC_s} \quad -(13) \]

The \((s)\) indicates that the transfer function varies as a function of frequency. This configuration introduces a gain and zero, expressed by:

\[ [H(s)] = \left( g_m \cdot \frac{R_2}{R_s + R_2} \right) \cdot R_s \]

\[ F_z = \frac{1}{2\pi \cdot R_s \cdot C_s} \quad -(14) \]

The gain is determined by the voltage divider and error amplifier’s transconductance gain. First select the desired zero-crossover frequency \((F_o)\):

\[ F_o > F_{ESR} \quad \text{and} \quad F_o \leq (1/5 \sim 1/10) \cdot F_z \]

Use the following equation to calculate \(R_4\):

\[ R_4 = \frac{V_{in} \cdot F_o \cdot F_{ESR}}{V_m \cdot F_{LC} \cdot (R_s + R_2) \cdot g_m} \quad -(15) \]

Where:
- \(V_{in}\) = Maximum Input Voltage
- \(V_{osc}\) = Oscillator Ramp Voltage
- \(F_o\) = Crossover Frequency
- \(F_{ESR}\) = Zero Frequency of the Output Capacitor
- \(F_{LC}\) = Resonant Frequency of the Output Filter
- \(R_s\) and \(R_2\) = Feedback Resistor Dividers
- \(g_m\) = Error Amplifier Transconductance
To cancel one of the LC filter poles, place the zero before the LC filter resonant frequency pole:

\[ F_z = \frac{75\% F_{LC}}{2\pi L \cdot C_6} \]

\[ F_z = 0.75 \cdot \frac{1}{2\pi \sqrt{L \cdot C_6}} \quad -(16) \]

Using equations (15) and (16) to calculate \( C_4 \).

One more capacitor is sometimes added in parallel with \( C_4 \) and \( R_3 \). This introduces one more pole which is mainly used to suppress the switching noise.

The additional pole is given by:

\[ F_p = \frac{1}{2\pi R_3 \cdot C_4 \cdot C_{POLE}} \quad - \left( s \right) \]

The pole sets to one half of switching frequency which results in the capacitor \( C_{POLE} \):

\[ C_{POLE} = \frac{1}{\pi R_3 \cdot F_p - \frac{1}{C_4}} \]

For \( F_p \ll \frac{F_s}{2} \)

For a general solution for unconditionally stability for any type of output capacitors, in a wide range of ESR values we should implement local feedback with a compensation network (typeIII). The typically used compensation network for voltage-mode controller is shown in figure 15.

In such configuration, the transfer function is given by:

\[ \frac{V_o}{V_i} = \frac{1 - g_m Z_f}{1 + g_m Z_{IN}} \]

The error amplifier gain is independent of the transconductance under the following condition:

\[ g_m \cdot Z_f > 1 \quad \text{and} \quad g_m \cdot Z_{IN} > 1 \quad -(17) \]

By replacing \( Z_{IN} \) and \( Z_f \) according to figure 15, the transformer function can be expressed as:

\[ H(s) = \frac{1}{sR_2 \left( \frac{C_2}{C_3} \right)} \left( 1 + \frac{sR_2 C_2}{1 + s \left[ \frac{C_2}{C_3} \left( 1 + sR_2 \right) \right]} \right) \]

Cross over frequency is expressed as:

\[ F_c = \frac{R_3 \cdot C_7 \cdot \frac{V_o}{I_{osc}} \cdot \frac{1}{2\pi L \cdot C_6}}{1 + s \left( \frac{C_2}{C_3} \right)} \]
Based on the frequency of the zero generated by output capacitor and its ESR versus crossover frequency, the compensation type can be different. The table below shows the compensation types and location of crossover frequency.

<table>
<thead>
<tr>
<th>Compensator Type</th>
<th>F_{ESR} vs. F_o</th>
<th>Output Capacitor</th>
</tr>
</thead>
<tbody>
<tr>
<td>Typ II (PI)</td>
<td>F_{LC} &lt; F_{ESR} &lt; F_o &lt; F_{s/2}</td>
<td>Electrolytic, Tantalum</td>
</tr>
<tr>
<td>Typ III (PID) Method A</td>
<td>F_{LC} &lt; F_o &lt; F_{ESR} &lt; F_{s/2}</td>
<td>Tantalum, ceramic</td>
</tr>
<tr>
<td>Typ III (PID) Method B</td>
<td>F_{LC} &lt; F_o &lt; F_{s/2} &lt; F_{ESR}</td>
<td>Ceramic</td>
</tr>
</tbody>
</table>

Table 1 - The compensation type and location of F_{ESR} versus F_o

The details of these compensation types are discussed in application note AN-1043 which can be downloaded from IR Web-Site.

For F_{LC} < F_o < F_{s/2} < F_{ESR}

typ III method B is selected to place the pole and zeros.

F_o < F_{ESR} and F_o \leq (1/5 \sim 1/10) * F_s

F_{z2} = F_o / \left(1 - \sin \theta \over 1 + \sin \theta \right)

F_{p2} = F_o / \left(1 + \sin \theta \over 1 - \sin \theta \right)

Select: F_{z2} = 0.5 * F_{z2} \quad \text{and} \quad F_{p2} = 0.5 * F_s

R_s \geq 2 / g_m

The following design rules will give a crossover frequency approximately one-tenth of the switching frequency. The higher the band width, the potentially faster the load transient response. The DC gain will be large enough to provide high DC-regulation accuracy (typically -5dB to -12dB). The phase margin should be greater than 45° for overall stability.

Desired Phase Margin: \theta_{max} = \pi / 3

Calculate C_1, C_4 and C_7:

C_4 = \frac{1}{2 \pi * F_s} \ast R_s

C_3 = \frac{1}{2 \pi * F_p} \ast R_s

C_7 = 2 \pi * F_s \ast \frac{L_s \ast C_s \ast V_{in}}{R_s \ast V_{in}}

Calculate R_{n0}, R_s and R_g:

R_{n0} = \frac{1}{2 \pi * C_s \ast F_s}

R_g = \frac{1}{2 \pi * C_s \ast F_s} - R_{n0}

R_s = \frac{V_{ref}}{V_o - V_{ref}} \ast R_g

Programming the Current-Limit

The Current-Limit threshold can be set by connecting a resistor (R_{SET}) from drain of low side MOSFET to the OCSet pin. The resistor can be calculated by using equation (3).

The R_{DS(on)} has a positive temperature coefficient and it should be considered for the worse case operation. This resistor must be placed close to the IC, place a small ceramic capacitor from this pin to ground for noise rejection purposes.

I_{SET} = \frac{I_{(OCSet)}}{R_{OCSet}} = \frac{R_{OCSet} \ast I_{OCSet}}{R_{DS(on)}} \quad \text{--}(3)
**Layout Consideration**

The layout is very important when designing high frequency switching converters. Layout will affect noise pickup and can cause a good design to perform with less than expected results.

Start to place the power components, make all the connection in the top layer with wide, copper filled areas.

The inductor, output capacitor and the MOSFET should be close to each other as possible. This helps to reduce the EMI radiated by the power traces due to the high switching currents through them. Place input capacitor directly to the drain of the high-side MOSFET, to reduce the ESR. Replace the single input capacitor with two parallel units.

The feedback part of the system should be kept away from the inductor and other noise sources. The critical bypass components such as capacitors for Vcc, DRVc and Vb should be close to respective pins. It is important to place the feedback components including feedback resistors and compensation components close to Fb and Comp pins.

In multilayer PCB use one layer as power ground plane and have a control circuit ground (analog ground), to which all signals are referenced. The goal is to localize the high current path to a separate loop that does not interfere with the more sensitive analog control function. These two grounds must be connected together on the PCB board layout at a single point.

---

**Part Marking Information**

![Part Marking Diagram]

- **Part number**: IRxxxxxx
- **Date code**: YWW?
- **Pin 1 Identifier**: □
- **Lot Code**: (Prod mode - 4 digit SPN code)
- **MARKING CODE**: ?
- **Lead Free Released**: P
- **Non-Lead Free Released**: 
- **Assembly site code**: XXXX
(S) SOIC Package
14-Pin Surface Mount, Narrow Body

NOTE: ALL MEASUREMENTS ARE IN MILLIMETERS.