PWM Buck Controller, Synchronous, 100V

NCP1034

Description

The NCP1034 is a high voltage PWM controller designed for high performance synchronous Buck DC/DC applications with input voltages up to 100 V. The NCP1034 drives a pair of external N-MOSFETs. The switching frequency is programmable from 25 kHz up to 500 kHz allowing the flexibility to tune for efficiency and size. A synchronization feature allows the switching frequency to be set by an external source or output a synchronization signal to multiple NCP1034 controllers. The output voltage can be precisely regulated using the internally trimmed 1.25 V reference voltage for low voltage applications. Protection features include user programmable undervoltage lockout and hiccup current limit.

Features

• High Voltage Operating up to 100 V
• Programmable Switching Frequency up to 500 kHz
• 2 A Output Drive Capability
• Precision Reference Voltage (1.25 V)
• Programmable Soft–Start with Prebiased Load Capability
• Programmable Overcurrent Protection
• Programmable Undervoltage Protection
• Hiccup Current Limit Using MOSFET R_DS(on) Sensing
• External Frequency Synchronization
• 16 Pin SOIC Package
• This is a Pb–Free Device

Applications

• 48 V Non–Isolated DC–DC Converter
• Embedded Telecom Systems
• Networking and Computing Voltage Regulator
• Distributed Point of Load Power Architectures
• General High Voltage DC–DC Converters

ORDERING INFORMATION

See detailed ordering and shipping information in the package dimensions section on page 24 of this data sheet.
Figure 1. Typical Application Circuit

(For applications operating at cold ambient temperatures refer to the soft–start section on Page 15)
Figure 2. Internal Block Diagram

Bias Generator

5V = VBIAS
1.25V = Vref

POR

5.5V = VBIAS

Oscillator

S
R
Reset Dom

Vcc UVLO

Rt

Ct

Rt

VCC

GND

UVLO

SYNC

VCC

Fb

Comp

Vref

Error Amp

Positive Current

0.3V

SS/SD

Comp

Vref

Error Comparator

Delay

LS

UV Detect

LowUVLO

OCP

0.225x Iocset

Positive Current

OCP Reset

VBIAS

VBIAS

OCin

Active Clamp

AC ON

0.0410x Iocset

Positive Current

Negative Output

AC ON
<table>
<thead>
<tr>
<th>PIN</th>
<th>PIN NAME</th>
<th>DESCRIPTION</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>OCset</td>
<td>Current limit set point. A resistor from this pin to GND will set the positive and negative current limit threshold.</td>
</tr>
<tr>
<td>2</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>3</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>4</td>
<td>SS/SD</td>
<td>Soft–Start / Shutdown. This pin provides user programmable soft-start function. External capacitor connected from this pin to ground sets the startup time of the output voltage. The converter can be shutdown by pulling this pin below 0.3 V.</td>
</tr>
<tr>
<td>5</td>
<td>SYNC</td>
<td>The internal oscillator can be synchronized to an external clock via this pin and other IC’s can be synchronized via this pin to internal oscillator. If it is not used this pin should be connected via 10 kΩ resistor to ground.</td>
</tr>
<tr>
<td>6</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>7</td>
<td>LDRV</td>
<td>Output driver for low side MOSFET.</td>
</tr>
<tr>
<td>8</td>
<td>DRVVCC</td>
<td>This pin provides biasing for the internal low side driver. A minimum of 0.1 μF, high frequency capacitor must be connected from this pin to power ground.</td>
</tr>
<tr>
<td>9</td>
<td>VB</td>
<td>This pin powers the high side driver and must be connected to a voltage higher than input voltage. A minimum of 0.1 μF, high frequency capacitor must be connected from this pin to switch node.</td>
</tr>
<tr>
<td>10</td>
<td>HDRV</td>
<td>Output driver for high side MOSFET.</td>
</tr>
<tr>
<td>11</td>
<td>VS</td>
<td>Switch Node. This pin is connected 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>12</td>
<td>VCC</td>
<td>This pin provides power for the internal blocks of the IC. A minimum of 0.1 μF, high frequency capacitor must be connected from this pin to ground.</td>
</tr>
<tr>
<td>13</td>
<td>OCIN</td>
<td>Overcurrent sensing input. A serial resistor from this pin to drain of low MOSFET must be used to limit the current into this pin.</td>
</tr>
<tr>
<td>14</td>
<td>GND</td>
<td>Signal ground for internal reference and control circuitry.</td>
</tr>
<tr>
<td>15</td>
<td>RT</td>
<td>Connecting a resistor from this pin to ground sets the oscillator frequency.</td>
</tr>
<tr>
<td>16</td>
<td>UVLO</td>
<td>An external voltage divider is used to set the undervoltage threshold levels.</td>
</tr>
</tbody>
</table>
# ABSOLUTE MAXIMUM RATINGS

<table>
<thead>
<tr>
<th>Rating</th>
<th>Symbol</th>
<th>Min</th>
<th>Max</th>
<th>Unit</th>
</tr>
</thead>
<tbody>
<tr>
<td>FB, VUVLO, RT, OCset</td>
<td>−0.3</td>
<td>10</td>
<td>V</td>
<td></td>
</tr>
<tr>
<td>COMP, SS/SD, SYNC, OCIN</td>
<td>−0.3</td>
<td>6</td>
<td>V</td>
<td></td>
</tr>
<tr>
<td>LDRV</td>
<td>−0.3</td>
<td>VCC + 0.3</td>
<td>V</td>
<td></td>
</tr>
<tr>
<td>DRVVCC, VCC</td>
<td>−0.3</td>
<td>20</td>
<td>V</td>
<td></td>
</tr>
<tr>
<td>VB</td>
<td>VS</td>
<td>VS + 20</td>
<td>V</td>
<td></td>
</tr>
<tr>
<td>HDRV</td>
<td>VS − 0.3</td>
<td>VB + 0.3</td>
<td>V</td>
<td></td>
</tr>
<tr>
<td>Vg</td>
<td>−1.0</td>
<td>150</td>
<td>V</td>
<td></td>
</tr>
<tr>
<td>OCin Input Current</td>
<td></td>
<td>20</td>
<td>mA</td>
<td></td>
</tr>
</tbody>
</table>

All voltages referenced to GND

<table>
<thead>
<tr>
<th>Rating</th>
<th>Symbol</th>
<th>Value</th>
<th>Unit</th>
</tr>
</thead>
<tbody>
<tr>
<td>Thermal Resistance, Junction--to--Ambient</td>
<td>RθJA</td>
<td>130</td>
<td>°C/W</td>
</tr>
<tr>
<td>Operating Ambient Temperature Range</td>
<td>T_A</td>
<td>−40 to 125</td>
<td>°C</td>
</tr>
<tr>
<td>Storage Temperature Range</td>
<td>T_STG</td>
<td>−55 to 150</td>
<td>°C</td>
</tr>
<tr>
<td>Junction Operating Temperature</td>
<td>T_J</td>
<td>−40 to 150</td>
<td>°C</td>
</tr>
<tr>
<td>ESD Withstand Voltage (Note 1)</td>
<td>V_ESD</td>
<td>2000</td>
<td>V</td>
</tr>
<tr>
<td>Machine Model</td>
<td></td>
<td>200</td>
<td>V</td>
</tr>
<tr>
<td>Latchup Capability per Jedec JESD78</td>
<td></td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

Stresses exceeding those listed in the Maximum Ratings table may damage the device. If any of these limits are exceeded, device functionality should not be assumed, damage may occur and reliability may be affected.

1. Excluding pins V_b, V_S and HDRV.

## TYPICAL ELECTRICAL PARAMETERS

### 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_IN</td>
<td>Converting Voltage</td>
<td>100</td>
<td>V</td>
<td></td>
</tr>
<tr>
<td>V_CC</td>
<td>Supply Voltage</td>
<td>10</td>
<td>18</td>
<td>V</td>
</tr>
<tr>
<td>DRVVCC</td>
<td>Supply Voltage</td>
<td>10</td>
<td>18</td>
<td>V</td>
</tr>
<tr>
<td>V_B to V_S</td>
<td>Supply Voltage</td>
<td>10</td>
<td>18</td>
<td>V</td>
</tr>
<tr>
<td>F_SW</td>
<td>Operating Frequency</td>
<td>25</td>
<td>500</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>

Functional operation above the stresses listed in the Recommended Operating Ranges is not implied. Extended exposure to stresses beyond the Recommended Operating Ranges limits may affect device reliability.
### ELECTRICAL CHARACTERISTICS
(Unless otherwise specified, these specifications apply over $V_{CC} = 12\,V$, $DRVV_{CC} = V_B = 12\,V$, $-40^{\circ}\,C < TJ < 125^{\circ}\,C$)

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Symbol</th>
<th>Test Condition</th>
<th>Min</th>
<th>Typ</th>
<th>Max</th>
<th>Unit</th>
</tr>
</thead>
<tbody>
<tr>
<td>REFERENCE VOLTAGE</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Feedback Voltage</td>
<td>$V_{FB}$</td>
<td>$12^{\circ},V &lt; V_{CC} &lt; 18^{\circ},V$ (Note 3)</td>
<td>1.25</td>
<td></td>
<td></td>
<td>V</td>
</tr>
<tr>
<td>Accuracy</td>
<td></td>
<td>$-40^{\circ},C &lt; TJ &lt; 125^{\circ},C$</td>
<td>-1.5</td>
<td>+1.5</td>
<td></td>
<td>%</td>
</tr>
<tr>
<td>FB Voltage Line Regulation</td>
<td>$L_{REG}$</td>
<td>$10,V &lt; V_{CC} &lt; 18,V$ (Note 3)</td>
<td>2.0</td>
<td></td>
<td></td>
<td>mV</td>
</tr>
</tbody>
</table>

| SUPPLY CURRENT                   |        |                                                                                  |       |       |       |       |
| $V_{CC}$ Supply Current (Stat)    | $I_{CC(Static)}$ | $S_S = 0\,V$, No Switching, $R_T = 10\,k\Omega$, $R_{OCSET} = 10\,k\Omega$ | 2.0   | 3.0   |       | mA    |
| $DRV_{CC}$ Supply Current (Stat)  | $I_{C(Static)}$ | $S_S = 0\,V$, No Switching                                                      | 0.1   | 0.3   |       | mA    |
| $V_B$ Supply Current (Stat)       | $I_{B(Static)}$ | $S_S = 0\,V$, No Switching                                                      | 0.1   | 0.3   |       | mA    |

| UNDervoltage LOCKOUT             |        |                                                                                  |       |       |       |       |
| $V_{CC}$–Start–Threshold         | $V_{CC_{UVLO}}(R)$ | Supply Ramping Up                                                               | 7.9   | 8.9   | 9.8   | V     |
| $V_{CC}$–Stop–Threshold          | $V_{CC_{UVLO}}(F)$ | Supply Ramping Down                                                             | 7.3   | 8.2   | 9.0   | V     |
| $V_{CC}$–Hysteresis              |        | Supply Ramping Up and Down                                                     | 0.7   |       |       | V     |
| $DRV_{CC}$–Start–Threshold       | $DRV_{CC_{UVLO}}(R)$ | Supply Ramping Up                                                               | 7.9   | 8.9   | 9.8   | V     |
| $DRV_{CC}$–Stop–Threshold        | $DRV_{CC_{UVLO}}(F)$ | Supply Ramping Down                                                             | 7.3   | 8.2   | 9.0   | V     |
| $V_B$–Hysteresis                 |        | Supply Ramping Up and Down                                                     | 0.7   |       |       | V     |
| Undervoltage Threshold Value     | $U_{UVLO}$ (Rising) |                                                                                  | 1.19  | 1.25  | 1.31  | V     |
| Undervoltage Threshold Value     | $U_{UVLO}$ (Falling) |                                                                                  | 1.10  | 1.15  | 1.20  | V     |

| OSCILLATOR                       |        |                                                                                  |       |       |       |       |
| Frequency                        | $F_S$  | $R_T = 20\,k\Omega$, $R_{T} = 10\,k\Omega$                                  | 170   | 200   | 230   | kHz   |
| Ramp Amplitude                   | $V_{ramp}$ | (Note 3)                                                                        | 2.0   |       |       | V     |
| Min Duty Cycle                   | $D_{min}$ | $FB = 2\,V$                                                                      | 0     |       |       | %     |
| Min Pulse Width                  | $D_{min}$(ctrl) | $F_S = 200\,kHz$, (Note 3)                                                     | 200   |       |       | ns    |
| Max Duty Cycle                   | $D_{max}$ | $F_S = 400\,kHz$, $FB = 1.2\,V$                                               | 80    |       |       | %     |
| SYNC Frequency Range             | $SYNC(F_S)$ | $20\%$ Above Free Running Frequency                                             |       |       | 500   | kHz   |
| SYNC Pulse Duration              | $SYNC_{(pulse)}$ |                                                                       | 200   |       |       | ns    |
| SYNC High Level                  | $SYNC_{(H)}$ |                                                                           | 2.0   |       |       | V     |
| SYNC Low Level                   | $SYNC_{(L)}$ |                                                                           | 0.8   |       |       | V     |
| SYNC Input Threshold             | $SYNC_{(Thre)}$ |                                                                         | 1.6   |       |       | V     |
| SYNC Input Hysteresis            | $SYNC_{(Hyst)}$ |                                                                    | 300   |       |       | mV    |
| SYNC Input Impedance             | $SYNC_{(ZIN)}$ | (Note 3)                                                                       | 16    |       |       | k\Omega |
| SYNC Output Impedance            | $SYNC_{(OUT)}$ | (Note 3)                                                                       | 2.5   |       |       | k\Omega |
| SYNC Output Pulse Width          | $SYNC_{(Pulse\,Width)}$ | $F_S = 500\,kHz$, (Note 3)                                                     | 300   |       |       | ns    |

Product parametric performance is indicated in the Electrical Characteristics for the listed test conditions, unless otherwise noted. Product performance may not be indicated by the Electrical Characteristics if operated under different conditions.

2. Cold temperature performance is guaranteed via correlation using statistical quality control. Not tested in production.
3. Guaranteed by design but not tested in production.

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## ELECTRICAL CHARACTERISTICS

(Unless otherwise specified, these specifications apply over $V_{CC} = 12$ V, $DRV_{CC} = V_B = 12$ V, $-40^\circ \text{C} < TJ < 125^\circ \text{C}$)

### ERROR AMPLIFIER

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Symbol</th>
<th>Test Condition</th>
<th>Min</th>
<th>Typ</th>
<th>Max</th>
<th>Unit</th>
</tr>
</thead>
<tbody>
<tr>
<td>Input Bias Current</td>
<td>$I_{FB}$</td>
<td>$SS = 3$ V, $FB = 1$ V</td>
<td>$-0.1$</td>
<td>$-0.4$</td>
<td></td>
<td>$\mu$A</td>
</tr>
<tr>
<td>Source/Sink Current</td>
<td>$I_{(Source/Sink)}$</td>
<td></td>
<td>$50$</td>
<td>$100$</td>
<td>$120$</td>
<td>$\mu$A</td>
</tr>
<tr>
<td>Bandwidth</td>
<td></td>
<td>(Note 3)</td>
<td>$4.0$</td>
<td>$10$</td>
<td></td>
<td>MHz</td>
</tr>
<tr>
<td>DC gain</td>
<td></td>
<td>(Note 3)</td>
<td>$55$</td>
<td></td>
<td></td>
<td>dB</td>
</tr>
<tr>
<td>Transconductance</td>
<td>$g_m$</td>
<td>(Note 3)</td>
<td>$1500$</td>
<td>$3150$</td>
<td>$4000$</td>
<td>$\mu$mho</td>
</tr>
</tbody>
</table>

### SOFT–START/SD

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Symbol</th>
<th>Test Condition</th>
<th>Min</th>
<th>Typ</th>
<th>Max</th>
<th>Unit</th>
</tr>
</thead>
<tbody>
<tr>
<td>Soft–Start Current</td>
<td>$I_{SS}$</td>
<td>$SS = 0$ V</td>
<td>$15$</td>
<td>$20$</td>
<td>$25$</td>
<td>$\mu$A</td>
</tr>
<tr>
<td>Shutdown Output Threshold</td>
<td>$S_D$</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td>V</td>
</tr>
</tbody>
</table>

### OVERCURRENT PROTECTION

<table>
<thead>
<tr>
<th>OCSET Voltage</th>
<th>$V_{OCSET}$</th>
<th></th>
<th>$1.25$</th>
<th></th>
<th></th>
<th>V</th>
</tr>
</thead>
<tbody>
<tr>
<td>Hiccup Current</td>
<td>$I_{Hiccup}$</td>
<td>(Note 3)</td>
<td>$1.0$</td>
<td></td>
<td></td>
<td>$\mu$A</td>
</tr>
<tr>
<td>Hiccup Duty Cycle</td>
<td>$I_{Hiccup}/I_{SS}$</td>
<td>(Note 3)</td>
<td>$5.0$</td>
<td></td>
<td></td>
<td>%</td>
</tr>
</tbody>
</table>

### OUTPUT DRIVERS

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Symbol</th>
<th>Test Condition</th>
<th>Min</th>
<th>Typ</th>
<th>Max</th>
<th>Unit</th>
</tr>
</thead>
<tbody>
<tr>
<td>LO, Drive Rise Time</td>
<td>$t_{(Lo)}$</td>
<td>$C_L = 1.5$ nF (See Figure 3)</td>
<td>$17$</td>
<td></td>
<td></td>
<td>ns</td>
</tr>
<tr>
<td>HI Drive Rise Time</td>
<td>$t_{(Hi)}$</td>
<td>$C_L = 1.5$ nF (See Figure 3)</td>
<td>$17$</td>
<td></td>
<td></td>
<td>ns</td>
</tr>
<tr>
<td>LO Drive Fall Time</td>
<td>$t_{(Lo)}$</td>
<td>$C_L = 1.5$ nF (See Figure 3)</td>
<td></td>
<td>$10$</td>
<td></td>
<td>ns</td>
</tr>
<tr>
<td>HI Drive Fall Time</td>
<td>$t_{(Hi)}$</td>
<td>$C_L = 1.5$ nF (See Figure 3)</td>
<td></td>
<td>$10$</td>
<td></td>
<td>ns</td>
</tr>
<tr>
<td>Dead Band Time</td>
<td>$t_{dead}$</td>
<td>(See Figure 3)</td>
<td>$30$</td>
<td>$60$</td>
<td>$120$</td>
<td>ns</td>
</tr>
<tr>
<td>LO Output High Short Circuit Pulsed Current</td>
<td>$I_{LDRVhigh}$</td>
<td>$V_{LDRV} = 0$ V, $P_W \leq 10$ $\mu$s, $T_J = 25^\circ$ C (Note 3)</td>
<td>$1.4$</td>
<td></td>
<td></td>
<td>A</td>
</tr>
<tr>
<td>HI Output High Short Circuit Pulsed Current</td>
<td>$I_{HDRVhigh}$</td>
<td>$V_{HDRV} = 0$ V, $P_W \leq 10$ $\mu$s, $T_J = 25^\circ$ C (Note 3)</td>
<td>$2.2$</td>
<td></td>
<td></td>
<td>A</td>
</tr>
<tr>
<td>LO Output Low Short Circuit Pulsed Current</td>
<td>$I_{LDRVhigh}$</td>
<td>$V_{LDRV} = DRV_{CC}$, $P_W \leq 10$ $\mu$s, $T_J = 25^\circ$ C (Note 3)</td>
<td>$1.4$</td>
<td></td>
<td></td>
<td>A</td>
</tr>
<tr>
<td>HI Output Low Short Circuit Pulsed Current</td>
<td>$I_{HDRVhigh}$</td>
<td>$V_{HDRV} = V_B$, $P_W \leq 10$ $\mu$s, $T_J = 25^\circ$ C (Note 3)</td>
<td>$2.2$</td>
<td></td>
<td></td>
<td>A</td>
</tr>
<tr>
<td>LO Output Resistor, Source</td>
<td>$R_{LOH}$</td>
<td>Typical Value @ $25^\circ$ C, (Note 3)</td>
<td>$7$</td>
<td>$12$</td>
<td></td>
<td>$\Omega$</td>
</tr>
<tr>
<td>LO Output Resistor, Sink</td>
<td>$R_{LOL}$</td>
<td>Typical Value @ $25^\circ$ C, (Note 3)</td>
<td>$2$</td>
<td>$8$</td>
<td></td>
<td>$\Omega$</td>
</tr>
<tr>
<td>HI Output Resistor, Source</td>
<td>$R_{HIH}$</td>
<td>Typical Value @ $25^\circ$ C, (Note 3)</td>
<td>$7$</td>
<td>$12$</td>
<td></td>
<td>$\Omega$</td>
</tr>
<tr>
<td>HI Output Resistor, Sink</td>
<td>$R_{HIL}$</td>
<td>Typical Value @ $25^\circ$ C, (Note 3)</td>
<td>$2$</td>
<td>$8$</td>
<td></td>
<td>$\Omega$</td>
</tr>
</tbody>
</table>

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

3. Guaranteed by design but not tested in production.
Figure 3. Definition of Rise–Fall Time and Deadband Time
Figure 4. $V_{FB}$

Figure 5. $UVLO_{VB}$

Figure 6. $UVLO_{VCC}$

Figure 7. $UVLO_{DRVCC}$

Figure 8. $UVLO$

Figure 9. $I_{CC}$ (Stat)
NCP1034

TYPICAL OPERATING CHARACTERISTICS

**Figure 10. Switching Frequency @ \( R_T = 20 \text{ k\,\Omega} \)**

**Figure 11. Maximum Duty Cycle @ \( f = 400 \text{ kHz} \)**

**Figure 12. Minimum on Time**

**Figure 13. Error Amplifier Transconductance**

**Figure 14. Deadtime**

**Figure 15. Driver Pullup Resistance**
TYPICAL OPERATING CHARACTERISTICS

Figure 16. Driver Pulldown Resistance

Figure 17. OCP @ $R_8 = 10 \, k\Omega$, $R_{OCIN} = 10 \, k\Omega$

Figure 18. POSOCP @ $R_8 = 10 \, k\Omega$, $R_{OCIN} = 10 \, k\Omega$
Undervoltage Lock-out

There are four undervoltage lock-out circuits. Two of them protect external high-side and low-side drivers, the third ensures that the IC does not start until \( V_{CC} \) is under a set threshold. The last one can be programmed by the user. It has a rising threshold at 1.25 V and a falling threshold at 1.15 V, and the user can define the undervoltage level by an external resistor divider. If the voltage is not over the threshold value, the device stops operating. The high-side driver UVLO only stops switching the high-side MOSFET.

Programmed falling and rising UVLO voltage can be calculated by Equations 1 and 2:

\[
V_{UVLO,\text{falling}} = 1.15 \cdot \left(1 + \frac{R4}{R5}\right) \quad \text{(eq. 1)}
\]

and

\[
V_{UVLO,\text{rising}} = 1.25 \cdot \left(1 + \frac{R4}{R5}\right) \quad \text{(eq. 2)}
\]

Shutdown

The output voltage can be disabled by pulling the SS/SD pin below 0.3 V. A small transistor can be used to pull it down as shown in Figure 19. During this time, both external MOSFETs are turned off. After the SS/SD pin is released, the IC starts its operation with a soft-start sequence.

![Figure 19. Shutdown Interface](image)

Operating Frequency Selection

The operating frequency is set by an external resistor connected from the \( R_t \) pin to ground. The value of this resistor can be selected from Figure 20, which shows switching frequency versus the timing resistor value.

![Figure 20. Frequency Dependence of \( R_t \) Value](image)

Frequency Synchronization

The NCP1034 can be synchronized to an external clock signal. The input synchronization signal should be a TTL logic level. The oscillator is synchronized to the rising edge of the synchronizing signal. When synchronization is used, the free running frequency must be set by the timing resistor to a frequency at least 80% of the external synchronization frequency (Example: \( R_T = 20 \, \text{k}\Omega / 200 \, \text{kHz} \) and external TTL = 220 kHz).

The NCP1034 can also output synchronization pulses on the SYNC pin. Pulses are generated when the internal oscillator ramp reaches the high threshold voltage. The frequency of these pulses is set by an external \( R_T \) resistor. Up to five NCP1034 controllers can be connected directly to the SYNC pin, all of which are synchronized to the controller with the highest frequency. The lowest frequency must be at least 80% of the highest one.

The equivalent internal circuit of the Sync pin is shown in Figure 21.

![Figure 21. Equivalent Connection of the Sync Pin](image)
Figure 22 shows the part with no synchronization. In this circuit the internal clock is fixed by the external timing resistor $R_T$. The SYNC pin can be tied to GND through a series resistor to prevent false triggering in a noisy environment.

![Figure 22. Fixed Frequency](image)

Figure 23 shows the part synchronized to an external clock through the SYNC pin. The synchronization frequency can be up to 20% greater than the programmed fixed frequency (Example: $R_T = 20 \, \text{k}\Omega / 200 \, \text{kHz}$ and the SYNC input frequency can range from 200–220 kHz). The clock frequency at the SYNC pin replaces the master clock generated by the internal oscillator circuit. Pulling the SYNC pin low programs the part to run freely at the frequency programmed by $R_T$. When pulling the SYNC pin low a 4.7 kΩ resistor should be used.

![Figure 23. External Synchronization](image)

Figure 24 shows the part operating in the master slave synchronization configuration. In this configuration all three parts are connected together through the SYNC pin in order to synchronize the system switching frequency. The $R_T$ timing resistor can be the same value for all three parts ($R_T = 20 \, \text{k}\Omega / 20 \, \text{k}\Omega / 20 \, \text{k}\Omega$) which would make the highest frequency part the master, or to guarantee one part is the master the timing resistor can be slightly lower in value. ($R_T = 20 \, \text{k}\Omega / 22 \, \text{k}\Omega / 22 \, \text{k}\Omega$)

![Figure 24. Master Slave Synchronization](image)
The value of the output capacitor should be calculated using the following equation:

\[
C_{\text{OUT}} \geq \frac{\Delta I_L}{8 \cdot f \cdot (\Delta V_{\text{OUT}} - \Delta I_L \cdot \text{ESR})} \quad \text{(eq. 5)}
\]

For higher switching frequency, it is suitable to use multi-layer ceramic capacitor (MLCC) with very low ESR. The advantages are small size, low output voltage ripple and fast transient response. The disadvantage of MLCC type is the requirement to use a Type III compensation network.

**Input Capacitor Selection**

The input capacitor is used to supply current pulses while high-side MOSFET is on. When the MOSFET is off, the input capacitor is being charged. The value of this capacitor can be selected with Equation 6:

\[
C_{\text{IN}} \geq \frac{I_{\text{OUT}} \cdot V_{\text{OUT}}}{V_{\text{IN}}} \cdot \left(1 - \frac{V_{\text{OUT}}}{V_{\text{IN}}}\right) f \cdot \Delta V_{\text{IN}} \quad \text{(eq. 6)}
\]

Where \(\Delta V_{\text{IN}}\) is the input voltage ripple and the recommended value is about 2% - 5% of \(V_{\text{IN}}\). The input capacitor must be large enough to handle the input ripple current. Its value should be calculated using Equation 7:

\[
I_{\text{RMS}} = I_{\text{OUT}} \cdot \sqrt{\frac{V_{\text{OUT}} \cdot \left(1 - \frac{V_{\text{OUT}}}{V_{\text{IN}}}\right)}{V_{\text{IN}}}} \quad \text{(eq. 7)}
\]

**Power MOSFET Selection**

The NCP1034 uses two N-channel MOSFET’s. They can be primary selected by \(R_{\text{DS(on)}}\), maximum drain-to-source voltage and gate charge. \(R_{\text{DS(on)}}\) impacts conductive losses and gate charge impacts switching losses. The low side MOSFET is selected primarily for conduction losses, and the high-side MOSFET is selected to reduce switching losses especially when the output voltage is less than 30% of the input voltages. The drain-to-source breakdown voltage must be higher than the maximum input voltage. Conductive power losses can be calculated using the Equations 8 and 9:

\[
P_{\text{COND-HIGHFET}} = I_{\text{OUT}}^2 \cdot R_{\text{DS(on)}} \cdot \frac{V_{\text{OUT}}}{V_{\text{IN}}} \quad \text{(eq. 8)}
\]

\[
P_{\text{COND-LOWFET}} = I_{\text{OUT}}^2 \cdot R_{\text{DS(on)}} \cdot \left(1 - \frac{V_{\text{OUT}}}{V_{\text{IN}}}\right) \quad \text{(eq. 9)}
\]

Switching losses are dependent on drain-to-source voltage at turn–off state, output current and switch–on and switch–off time as is shown by Equation 10.

\[
P_{\text{SW}} = \frac{V_{\text{DS(off)}}}{2} \cdot (t_{\text{ON}} + t_{\text{OFF}}) \cdot f \cdot I_{\text{OUT}} \quad \text{(eq. 10)}
\]

\(t_{\text{ON}}\) and \(t_{\text{OFF}}\) times are dependent on the transistor gate.

The MOSFET output capacitance loss is caused by the charging and discharging during the switching process and can be computed using Equation 11.

\[
P_{\text{COSS}} = C_{\text{OSS}} \cdot \frac{V_{\text{IN}}^2 \cdot f}{2} \quad \text{(eq. 11)}
\]

Where \(C_{\text{OSS}} = C_{\text{DS}} + C_{\text{GD}}\).

Significant power dissipation is caused by the reverse recovery charge in the low–side MOSFET body diode, which conducts at dead time. This charge is needed to close the diode. The current from the input power supply flows through the high–side MOSFET to the low–side MOSFET body diode. This power dissipation can be calculated using Equation 12.

\[
P_{\text{QRR}} = Q_{\text{RR}} \cdot V_{\text{IN}} \cdot f \quad \text{(eq. 12)}
\]

\(Q_{\text{RR}}\) is the diode recovery charge as given in the manufacturer’s datasheet. For some types of MOSFETs, this dissipation may be dominant at high input voltages. It is necessary to take care when selecting a MOSFET. An external Schottky diode across the low–side MOSFET can be used to eliminate the reverse recovery charge power loss. The Schottky diode’s forward voltage should be lower than that of the body diode, and reverse recovery time \(t_{\text{rr}}\) should be lower then that of the body diode. The Schottky diode’s capacitance loss can be calculated as shown in Equation 13.

\[
P_{\text{C(Schottky)}} = \frac{C_{\text{Schottky}} \cdot V_{\text{IN}}^2 \cdot f}{2} \quad \text{(eq. 13)}
\]
MOSFETs delays, turn–on and turn–off times must be short enough to prevent cross conduction. If not, there will be cross conduction from the input through both MOSFETs to ground. Due to this fact, the following conditions must be true:

\[
\begin{align*}
    t_{d(on)\text{high}} + t_{\text{dead}} &> t_{d(off)\text{low}} + t_f/low \\
    t_{d(on)\text{low}} + t_{\text{dead}} &> t_{d(off)\text{high}} + t_f/high
\end{align*}
\]  

(eq. 14)

Where \( t_{\text{dead}} \) is the controller dead band time, \( t_{d(on)} \), \( t_r \), \( t_d(\text{off}) \) and \( t_f \) are MOSFETs parameters. These parameters can be found in the datasheet for specific conditions.

It is NOT recommended to add external resistor or other circuit on MOSFETs' gates to slow–down their turn–off. If gate resistance is a must, please make sure the above condition in eq. 14 is still satisfied to avoid cross conduction.

**Bootstrap Circuit**

This circuit is used to obtain a voltage higher than the input voltage in order to switch–on high side N MOSFET. The bootstrap capacitor is charged from the IC’s supply voltage through D1, when the low side MOSFET is switched–on up to the IC’s supply voltage. It must have enough capacity to supply power for the high–side circuit when the high–side MOSFET is being switched on. The minimum value recommended for the bootstrap capacitor is 100 nF. Diode D1 has to be designed to withstand a reverse voltage given by the following equation:

\[
D1_{VR\text{min}} = V_{IN} - V_{CC}
\]  

(eq. 15)

**Soft–Start**

The soft–start time is set by capacitor connected between SS/SD Pin and ground. This function is used for controlling the output voltage slope and limiting startup currents. The start–up sequence initiates when Power On Ready (POR) internal signal rises to logic high level. That means the supply voltage, low side drive supply voltage and external UVLO are over the set thresholds. The soft–start capacitor is charged by 20 \( \mu \)A current source. If POR is low, the SS/SD Pin is internally pulled to GND, which means that the NCP1034 is in a shutdown state. The SS/SD Pin voltage (0 V to 2.6 V) controls internal current source (64 \( \mu \)A to 0 \( \mu \)A) with negative linear characteristic. This current source injects current into the resistor (25 k\( \Omega \)) connected between the Fb pin and negative input of the error amplifier and into the external feedback resistor network. Voltage drop on these resistors is over 1.6 V, which is enough to force the error amplifier into negative saturation state and to block switching.

Note that at cold ambient temperatures (–35°C) the internal 25 k\( \Omega \) drops up to 25% in value and so does the internal current source (64 \( \mu \)A) up to 10%. For those reasons, users must compensate for these variations by increasing the external lower resistive divider value in order to force the error amplifier into negative saturation at Soft–Start. Here is an example at –35°C showing how to select the proper R2 resistor:

\[
R_{\text{internal}} \sim 19 \, k\Omega
\]

Internal Current Source \sim 58 \( \mu \)A

\[
V_{R\text{internal}} = 19 \, k\Omega \times 58 \, \mu A = 1.1 \, V
\]

Select R2 such that:

\[
V_{R2} = 1.6 \, V - 1.1 \, V = 0.5 \, V
\]

\[
0.5 \, V = 58 \, \mu A = 8.6 \, k\Omega
\]

The minimum value required for R2 to keep Comp at GND is 8.6 k\( \Omega \).
When the soft–start pin reaches around 1.2 V (exact value depends on feedback and compensation network and on soft–start capacitor; a larger soft–start capacitor and a lower compensation capacity decrease this level) the IC starts switching. The impact of controlled current source decreases and the output voltage starts to rise. When the soft–start capacitor voltage reaches 2.6 V, the output voltage is at nominal value.

The soft–start time must be at least 10 times longer than the time needed to charge the compensation network from the output of the error amplifier. If the soft–start time is not long enough, the soft–start sequence would be faster than the charging compensation network and the IC would start without slowly increasing the output voltage. The soft–start capacitance can be calculated using Equation 16.

\[ C_{SS} = 15 \cdot 10^{-6} \cdot T_{SS} \] (eq. 16)

---

**Start to Prebiased Output**

The NCP1034 is able to startup into a prebiased output capacitor. The low–side MOSFET does not turn on before high–side MOSFET gets the first turn–on pulse. During this time, the energy is not discharged by the low–side MOSFET until the soft–start sequence crosses the programmed output voltage.
Overcurrent Protection

The voltage drop across the low side MOSFET $R_{DS(on)}$ is connected through resistor $R_8$ and into the IC through pin 13 OCin. Within the IC, this value is compared with the value programmed by resistor $R_7$ to set the overcurrent limit. The programmed current limit is set by selecting the value of $R_7$, which is connected between pin 1 OCset and GND. If the voltage drop is larger than the set value, the NCP1034 goes into hiccup mode. During this time, both external MOSFETs are turned off and the soft start capacitor is discharged with a current equal to 5% of the charging current. The capacitor continues to discharge until the voltage reaches 0.25 V, and then the IC initiates a standard soft start sequence.

The recommended value for the protection resistor $R_8$ is 10 kΩ. The $R_7$ resistance value can be calculated using Equation 17:

$$R_7 = \frac{R_8 \cdot 3.56 \cdot R_{DS(on)} \cdot I_{pk}}{I_{pos}}$$

(eq. 17)

**Figure 28. Overcurrent Protection (Hi–Cup Mode)**

The NCP1034 provides protection of the low side MOSFET against positive overcurrent (from output to this MOSFET). Its value can be calculated using Equation 18:

$$I_{pos} = \frac{5125 - 0.184 \cdot R_8 \cdot 1.25}{R_7 \cdot R_{DS(on)}}$$

(eq. 18)

NCP1034’s overcurrent protection threshold could be affected by external circuits and PCB layout. Please pay attention to the following:

- Do not slow down the low–side MOSFET turning-on by any resistance or other circuit on its gate. About 80 ns after the rising edge of LDRV pin, the NCP1034 overcurrent protection function starts. If the low–side MOSFET hasn’t been fully turned–on then, the overcurrent protection may be falsely triggered, even at very low load current.

- OCin trace layout

  The OCin trace, between OCin pin and $R_8$, is a high impedance node. Any noise coupling to it may falsely trigger overcurrent protection. Please avoid any noise source near this OCin trace, such as VS, VB, HDRV and LDRV nodes. Any capacitance on the OCin pin impacts the overcurrent protection threshold as well. Therefore, it is not recommended.

- The voltage difference between PGND pin and low–side MOSFET source pin affects overcurrent protection threshold. As shown in Figure 2, the overcurrent comparator input pin OCin is reference to PGND pin. Therefore, the overcurrent protection threshold should factor in the voltage difference between the external MOSFET’s source pins and the NCP1034’s PGND pin.

- fix $R_8 = 10$ kΩ

  As shown in Eq. 17 and Eq. 18, $R_8$ resistance affects overcurrent limit threshold and positive overcurrent limit threshold in opposite directions. To simplify the design, please fix $R_8$ at 10 kΩ as possible, and use $R_7$ to program overcurrent limit threshold.

**Compensation Circuit**

The NCP1034 is a voltage mode buck converter with a transconductance error amplifier compensated by an external compensation network. Compensation is needed to achieve accurate output voltage regulation and fast transient response. The goal of the compensation circuit is to provide a loop gain function with the highest crossing frequency and adequate phase margin (minimally 45°).
The transfer function of the power stage (the output LC filter) is a double pole system. The resonance frequency of this filter is expressed as follows:

\[ f_{P0} = \frac{1}{2 \cdot \pi \cdot \sqrt{L \cdot C_{OUT}}} \]  
(eq. 19)

One zero of this LC filter is given by the output capacitance and output capacitor ESR. Its value can be calculated by using the following equation:

\[ f_{Z0} = \frac{1}{2 \cdot \pi \cdot C_{OUT} \cdot ESR} \]  
(eq. 20)

The next parameter that must be chosen is the zero crossover frequency \( f_0 \). It can be chosen to be \( 1/10 \)−\( 1/5 \) of the switching frequency. These three parameters show the necessary type of compensation that can be selected from Table 1.

### Table 1. COMPENSATION TYPES

<table>
<thead>
<tr>
<th>Zero Crossover Frequency Condition</th>
<th>Compensation Type</th>
<th>Typical Output Capacitor Type</th>
</tr>
</thead>
<tbody>
<tr>
<td>( f_{P0} &lt; f_{Z0} &lt; f_0 &lt; f_S/2 )</td>
<td>Type II (PI)</td>
<td>Electrolytic, Tantalum</td>
</tr>
<tr>
<td>( f_{P0} &lt; f_0 &lt; f_{Z0} &lt; f_S/2 )</td>
<td>Type III (PID) Method I</td>
<td>Tantalum, Ceramic</td>
</tr>
<tr>
<td>( f_{P0} &lt; f_0 &lt; f_S/2 &lt; f_{Z0} )</td>
<td>Type III (PID) Method II</td>
<td>Ceramic</td>
</tr>
</tbody>
</table>

**Compensation Type II (PI)**

This compensation is suitable for low-cost electrolytic capacitor. The zero created by the capacitor’s ESR is a few kHz and the zero crossover frequency is chosen to be \( 1/10 \) of the switching frequency. Components of the PI compensation (Figure 29) network can be specified by the following equations:

\[ R_{C1} = \frac{2 \cdot \pi \cdot f_{0} \cdot L \cdot V_{RAMP} \cdot V_{OUT}}{ESR \cdot V_{IN} \cdot V_{ref} \cdot gm} \]  
\[ C_{C1} = \frac{1}{0.75 \cdot 2 \cdot \pi \cdot f_{P0} \cdot R_{C1}} \]  
(eq. 21)

\[ C_{C2} = \frac{\pi \cdot R_{C1} \cdot f_S}{V_{OUT} - V_{ref}} \cdot R_{1} \]

\[ R_{1} = \frac{V_{ref}}{V_{OUT} - V_{ref}} \cdot R_{2} \]

\[ V_{RAMP} \] is the peak-to-peak voltage of the oscillator ramp and \( gm \) is the transconductance error amplifier gain.

**Compensation Type III (PID)**

Tantalum and ceramics capacitors have lower ESR than electrolytic, so the zero of the output LC filter goes to a higher frequency above the zero crossover frequency. This situation needs to be compensated by the PID compensation network that is show in Figure 30.

There are two methods to select the zeros and poles of compensation network. The first one (method I) is useable for tantalum output capacitors, which have a higher ESR than ceramic, and its zeros and poles can be calculated shown below:

\[ f_{Z1} = 0.75 \cdot f_{P0} \]
\[ f_{Z2} = f_{P0} \]  
(eq. 22)
\[ f_{Z2} = f_{P0} \]
\[ f_{P2} = f_{Z0} \]
\[ f_{P3} = \frac{f_{S}}{2} \]

The second one (method II) is for ceramic capacitors:
The remaining calculations are the same for both methods.

\[ R_{C1} > > \frac{2}{g_m} \]
\[ C_{C1} = \frac{1}{2 \cdot \pi \cdot f_{Z1} \cdot R_{C1}} \]
\[ C_{C2} = \frac{1}{2 \cdot \pi \cdot f_{P3} \cdot R_{C1}} \]
\[ C_{FB1} = \frac{2 \cdot \pi \cdot f_0 \cdot L \cdot V_{RAMP} \cdot C_{OUT}}{V_{IN} \cdot R_{C1}} \]  
(eq. 24)
\[ R_{FB1} = \frac{1}{2 \pi \cdot C_{FB1} \cdot f_{P2}} \]
\[ R_1 = \frac{1}{2 \cdot \pi \cdot C_{FB1} \cdot f_{Z2} - R_{FB1}} \]
\[ R_2 = \frac{V_{ref}}{V_{OUT} - V_{ref}} \cdot R_1 \]

To check the design of this compensation network, the equation must be true
\[ R_1 \cdot R_2 \cdot R_{FB1} > \frac{1}{g_m} \]  
(eq. 25)

If it is not true, then a higher value of \( R_{C1} \) must be selected.

**Input Power Supply**

The NCP1034 controller and built-in drivers need to be powered through \( V_{CC} \), \( DRV_{VCC} \) and \( V_B \) pins with a voltage between 10 V – 18 V. The supply current requirement is a summation of the static and dynamic currents. Static current consumption can be calculated by the following equation:
\[ I_{CS} = I_{CC} + I_C + I_B \]  
(eq. 26)

Dynamic current consumption is calculated using the following equation, based on the switching frequency and MOSFET gate charge.
\[ I_{CD} = \left( Q_{G(lo)} + Q_{G(hi)} \right) \cdot f \]  
(eq. 27)

To power the device, an external power supply or voltage regulator from \( V_{IN} \) can be used. Two options are a linear shunt voltage regulator and a shunt voltage regulator with transistor, as shown in Figure 31. A voltage regulator without a transistor can be used when the power consumption is low and zener diode power dissipation is acceptable. Otherwise, a shunt regulator with transistor can be used.

**Figure 31. Linear Shunt Voltage Regulator**

For the linear shunt voltage regulator (option a) the \( V_{CC} \) voltage is the same as the zener diode reverse voltage \( V_Z \). The value of the resistor \( R \) can be calculated using Equation 28, where \( I_{ZT} \) is the minimum reverse current at \( V_Z \). The value selected should be lower than the calculated value. The maximum power losses of resistor \( R \) and the zener diode \( D \) can be calculated by Equations 29 and 30.
\[ R < \frac{V_{INmin} - V_{CC}}{I_{CS} + I_{CD} + I_{ZT}} \]  
(eq. 28)
\[ P_R = \left( V_{INmax} - V_{CC} \right) \cdot \left( I_{CS} + I_{CD} \right) \]  
(eq. 29)
\[ P_D = \left( \frac{V_{INmax} - V_{CC}}{R} \right) - I_{CS} \]  
(eq. 30)

The shunt voltage regulator with transistor (option b) is advantageous when the zener diode loss is too high or when input voltage varies across a wide range and it is difficult to set a bias point. The output voltage is lower than \( V_Z \) due to the \( V_{BE} \) of the transistor. The maximum resistor value of \( R \) can be calculated by Equation 31, where \( \beta \) is the transistor DC current gain. The maximum power dissipation of the resistor, zener diode and transistor are calculated by Equations 32 to 34. The transistor reverse breakdown voltage must be selected to be able to withstand the voltage difference between maximum input voltage and \( V_{CC} \).
\[ R < \frac{V_{INmin} - V_{ZT}}{\frac{I_{CS} + I_{CD}}{\beta} + I_{ZT}} \]  
(eq. 31)
\[ P_R = (V_{\text{INmax}} - V_{\text{CC}}) \cdot \left( \frac{l_{\text{CS}} + l_{\text{CD}}}{\beta} + l_{\text{ZT}} \right) \quad (\text{eq. 32}) \]

\[ P_D = \left( \frac{V_{\text{INmax}} - V_{\text{ZT}}}{R} - \frac{l_{\text{CS}}}{\beta} \right) \cdot V_{\text{ZT}} \quad (\text{eq. 34}) \]

\[ P_T = (V_{\text{INmax}} - V_{\text{CC}}) \cdot (l_{\text{CS}} + l_{\text{CD}}) \quad (\text{eq. 35}) \]

### Table 2. POWER SUPPLY REGULATOR EXAMPLES

<table>
<thead>
<tr>
<th>Components</th>
<th>MOSFETs</th>
<th>( Q_{\text{G(TOT)}} ) (nC)</th>
<th>( f ) (kHz)</th>
<th>( V_{\text{INmax}} ) (V)</th>
<th>( V_{\text{INmin}} ) (V)</th>
<th>( I_{\text{SUPPLYmax}} ) (mA)</th>
<th>( R_{\text{BIAS}} ) (k( \Omega ))</th>
<th>ZD</th>
<th>Transistor</th>
</tr>
</thead>
<tbody>
<tr>
<td>LS–FET</td>
<td>NTD24N06</td>
<td>24</td>
<td>200</td>
<td>60</td>
<td>36</td>
<td>8.7</td>
<td>2.6</td>
<td>MMSZ4699</td>
<td>–</td>
</tr>
<tr>
<td>HS–FET</td>
<td>NTD3055</td>
<td>7.1</td>
<td></td>
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<td></td>
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</tr>
<tr>
<td>LS–FET</td>
<td>NTD24N06</td>
<td>24</td>
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<td>60</td>
<td>20</td>
<td>16.9</td>
<td>10</td>
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<td>MJD31</td>
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<td>HS–FET</td>
<td>NTD24N06</td>
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<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
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</tr>
</tbody>
</table>

### PCB Layout

The layout of high-frequency and high-current switching converters has a large impact on the circuit parameters. It is important, therefore, to pay close attention to the PCB layout.

The input capacitor, MOSFETs, inductor and output capacitor should be placed as close as possible to one another. This is suitable to reduce EMI and to minimize \( V_S \) overshoots. Connecting the signal and power ground at one point near the output connector improves load regulation. Connection between the source pin of the low side MOSFET and the IC should be very short with wide traces and optimally using two layers to achieve minimum inductance between them.

The blocking and bootstrap capacitors should be placed as close as possible to the IC. The feedback and compensation network should be close to the IC to minimize noise.

### TYPICAL APPLICATION

![Figure 33. Single Output Buck Converter from 38 V – 58 V to 5 V/5 A @ 200 kHz](image-url)
Figure 34. Efficiency and Power Loss of Circuit at Figure 33
# Bill of Materials

<table>
<thead>
<tr>
<th>Designator</th>
<th>Qty</th>
<th>Description</th>
<th>Value</th>
<th>Tolerance</th>
<th>Footprint</th>
<th>Manufacturer</th>
<th>Part Number</th>
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</thead>
<tbody>
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<td>R9</td>
<td>1</td>
<td>Resistor</td>
<td>1k2</td>
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<td>1206</td>
<td>Vishay</td>
<td>CRCW10261K20FKEA</td>
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<td>1</td>
<td>Resistor</td>
<td>3k9</td>
<td>1%</td>
<td>1206</td>
<td>Vishay</td>
<td>CRCW10263K90FKEA</td>
</tr>
<tr>
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<td>Resistor</td>
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<td>1%</td>
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<td>Vishay</td>
<td>CRCW10264K60FKEA</td>
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<tr>
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<td>Resistor</td>
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<td>1%</td>
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<td>Vishay</td>
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<td>Resistor</td>
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<td>Vishay</td>
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<td>SOIC16</td>
<td>onsemi</td>
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Figure 39. Typical Application Board Photos

ORDERING INFORMATION

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<tr>
<td>NCP1034DR2G</td>
<td>SOIC-16</td>
<td>2500 / Tape &amp; Reel</td>
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<td>(Pb–Free)</td>
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†For information on tape and reel specifications, including part orientation and tape sizes, please refer to our Tape and Reel Packaging Specifications Brochure, BRD8011/D.
NOTES:
2. CONTROLLING DIMENSION: MILLIMETER.
3. DIMENSIONS A AND B DO NOT INCLUDE MOLD PROTRUSION.
4. MAXIMUM MOLD PROTRUSION 0.15 (0.006) PER SIDE.
5. DIMENSION D DOES NOT INCLUDE DAMBAR PROTRUSION. ALLOWABLE DAMBAR PROTRUSION SHALL BE 0.127 (0.005) TOTAL IN EXCESS OF THE D DIMENSION AT MAXIMUM MATERIAL CONDITION.

SOLDERING FOOTPRINT

DIMENSIONS: MILLIMETERS

---

**SOLDERING FOOTPRINT**

8X

---

**DIMENSIONS: MILLIMETERS**

---

**STYLE 1:**
- 1. COLLECTOR
- 2. BASE
- 3. Emitter
- 4. NO CONNECTION
- 5. Emitter
- 6. BASE
- 7. COLLECTOR
- 8. COLLECTOR
- 9. BASE
- 10. Emitter
- 11. NO CONNECTION
- 12. Emitter
- 13. BASE
- 14. COLLECTOR
- 15. Emitter
- 16. COLLECTOR

**STYLE 2:**
- 1. CATHODE
- 2. ANODE
- 3. NO CONNECTION
- 4. CATHODE
- 5. CATHODE
- 6. NO CONNECTION
- 7. ANODE
- 8. CATHODE
- 9. CATHODE
- 10. ANODE
- 11. NO CONNECTION
- 12. ANODE
- 13. CATHODE
- 14. CATHODE
- 15. ANODE
- 16. CATHODE

**STYLE 3:**
- 1. COLLECTOR, DYE #1
- 2. BASE, #1
- 3. EMITTER, #1
- 4. COLLECTOR, #1
- 5. COLLECTOR, #2
- 6. BASE, #2
- 7. COLLECTOR, #3
- 8. COLLECTOR, #4
- 9. BASE, #3
- 10. EMITTER, #3
- 11. BASE, #2
- 12. EMITTER, #2
- 13. BASE, #1
- 14. EMITTER, #1

**STYLE 4:**
- 1. COLLECTOR, DYE #1
- 2. COLLECTOR, #1
- 3. COLLECTOR, #2
- 4. COLLECTOR, #2
- 5. COLLECTOR, #3
- 6. COLLECTOR, #3
- 7. COLLECTOR, #4
- 8. COLLECTOR, #4
- 9. BASE, #4
- 10. EMITTER, #4
- 11. BASE, #3
- 12. EMITTER, #3
- 13. BASE, #2
- 14. EMITTER, #2
- 15. BASE, #1
- 16. EMITTER, #1

**STYLE 5:**
- 1. DRAIN, DYE #1
- 2. DRAIN, #1
- 3. DRAIN, #2
- 4. DRAIN, #2
- 5. DRAIN, #3
- 6. DRAIN, #3
- 7. DRAIN, #4
- 8. DRAIN, #4
- 9. GATE, #4
- 10. SOURCE, #4
- 11. GATE, #5
- 12. SOURCE, #5
- 13. GATE, #6
- 14. SOURCE, #6
- 15. GATE, #7
- 16. SOURCE, #7

**STYLE 6:**
- 1. CATHODE
- 2. CATHODE
- 3. CATHODE
- 4. CATHODE
- 5. CATHODE
- 6. CATHODE
- 7. CATHODE
- 8. CATHODE
- 9. ANODE
- 10. ANODE
- 11. ANODE
- 12. ANODE
- 13. ANODE
- 14. ANODE
- 15. ANODE
- 16. ANODE

**STYLE 7:**
- 1. SOURCE N-CH
- 2. COMMON DRAIN (OUTPUT)
- 3. COMMON DRAIN (OUTPUT)
- 4. GATE P-CH
- 5. COMMON DRAIN (OUTPUT)
- 6. COMMON DRAIN (OUTPUT)
- 7. COMMON DRAIN (OUTPUT)
- 8. SOURCE P-CH
- 9. SOURCE P-CH
- 10. SOURCE N-CH
- 11. COMMON DRAIN (OUTPUT)
- 12. COMMON DRAIN (OUTPUT)
- 13. COMMON DRAIN (OUTPUT)
- 14. COMMON DRAIN (OUTPUT)
- 15. COMMON DRAIN (OUTPUT)
- 16. SOURCE N-CH

**DIMENSIONS: MILLIMETERS**

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**LIST OF TABLES**

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<th>DIMENSIONS</th>
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<td>B</td>
<td>3.80 4.00</td>
<td>0.150 0.157</td>
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<tr>
<td>C</td>
<td>1.35 1.75</td>
<td>0.054 0.068</td>
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<td>D</td>
<td>0.35 0.49</td>
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**DOCUMENT NUMBER:** 98ASB42566B

**DESCRIPTION:** SOIC–16

**PAGE 1 OF 1**