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Tips and Tricks for the NCP1250

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APPLICATION NOTE

Despite a limited number of pins, the NCP1250 can be used in a variety of applications in the ac−dc or dc−dc power conversion field. Besides the available literature on the controller itself, this application note reviews a few tricks to help you improve the performance of the part in particular operating conditions.

Reducing the Standby Power

This is THE subject of discussion when tackling ac−dc converters for the consumer market: "what standby power in no−load condition can your part reach?". Well, if we disconnect the start−up resistors and the 2 MΩ X2−capacitors discharge resistors string in the NCP1250 demonstration board, we reach an amazing 35 mW when powered at 230 V_{rms} and fully unloaded on the secondary side (V_{out} = 19 V, I_{out} = 0). The main contributors to this consumption are the following ones:

1. X2−capacitor choice and discharge elements: as a substantial amount of reactive current circulates in this capacitor, it can induce dielectric losses, in particular with cheap components. Besides this loss, the voltage on its terminals must decrease at a sufficient pace when you unplug the power cord so that the available level becomes benign for a user touching the plug after 1 s. This is the reason why discharge resistors are connected in parallel with the filtering capacitor. These elements are selected so that the time constant involving the X2−capacitor and the resistors is 1 s, as specified by the IEC−950 standard. For a 0.47 μF capacitor, you must install a 2 MΩ resistors string, dissipating 26 mW at 230 V_{rms}. If you chase the tens of mW, you will be happy to discover an active circuitry described in this application note.
2. Bulk capacitor leakage: it is an often forgotten parameter because designers are not used to chasing tens of mW when thinking about a 65 W adapter. However, depending on the adopted brand, you can experience a consumption of a few mW, sometimes up to 10, just because the bulk capacitor is leaky. Please pay attention to this parameter if you plan to beat no−load consumption records.
3. Controller consumption: there is nothing you can do on this one. It depends on the design and technical choices the semiconductor vendor made when developing the device. The NCP1250 has been the object of a particular care in this domain. When supplied from a 12 V auxiliary source while driving a 6 A/600 V MOSFET, the controller only draws 550 μA typically. From the 12 V line, it is a bare 6 mW. Difficult to do less...
4. Feedback currents: if you selected a TL431, you must inject at least 1 mA in the device to get it working properly. If you do not, you will experience a poor output impedance, leading to an unacceptable transient performance. With a 19 V output, a 1 mA bias associated with the regular feedback current generates a significant primary−side loss. If your output voltage is below 18 V, you can use a TLV431 whose minimum injected current is down to 100 μA. If your buyer imposes a TL431, one of the proposed tricks will help getting rid of this extra consumption.
5. Start−up resistors: with low−voltage controllers, this is always the problem. How to combine a start−up time less than 3 s at low−line while consuming the least current on the mains at high line? You can always add an external bipolar transistor network to get rid of the start−up network, but why not taking advantage of the X2 discharge resistors presence to crank the controller? This is what is proposed in the following lines.
6. Output LED: needless to say that the addition of a LED in the adapter output can ruin all the efforts you put in saving the tens of mW! The best is, of course, to explain that the presence of the LED go against power saving initiatives and it would be better to abandon its implementation. In some cases, however, a light is needed and a solution has to be found to minimize its impact.

This application note gathers and details design tricks that will hopefully help you meeting stringent efficiency standards, while you enjoy working on this versatile controller called NCP1250!

Reducing the TL431 Bias Current

In association with JP Louvel

As we explained, the TL431 requires an operating bias current of at least 1 mA. To force the circulation of this current in the device, a well−known technique consists in adding a resistor in parallel with the optocoupler LED. As the LED forward voltage is nearly constant (≈1 V), you create a cheap current generator at the expense of an extra resistor. As shown in Figure 1, if a 1 kΩ resistor is installed across the LED, the extra injected current in the TL431 will simply be:

$$I_{bias} = \frac{V_f}{R_{bias}} \approx \frac{1}{1k} = 1\ \text{mA} \qquad \text{(eq. 1)}$$

Unfortunately, in no−load conditions, this extra current is still present. Coupled with a 19 V output and considering the operating feedback current (to maintain the primary−side controller in skip−cycle mode), a total power of around 25 mW is lost, reflected at higher value if we consider the poor efficiency at this amount of output power.

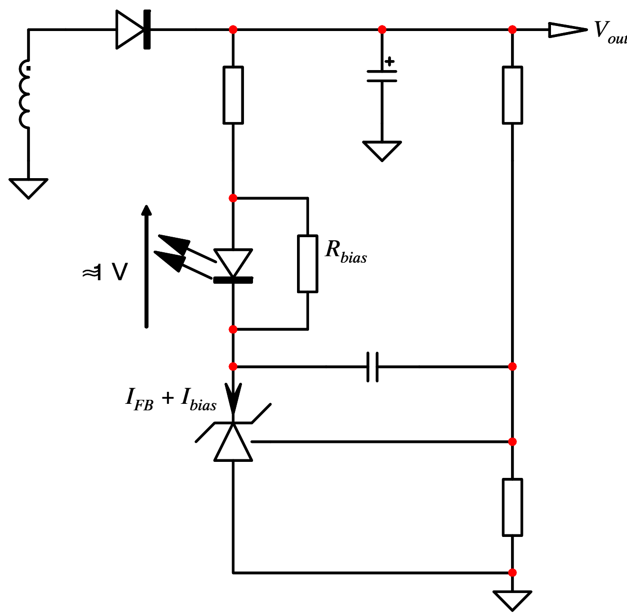


Figure 1. A Simple Resistor in Parallel with the Optocoupler LED is Enough to Force a Bias Current Into the TL431

There are several possibilities to reduce the bias current. However, as the TL431 operates with a minimum current, the biasing conditions must be restored as soon as the load comes back. Failure to do this will severely degrade the transient response. A simple means has recently been patented by ON Semiconductor and appears in Figure 2.

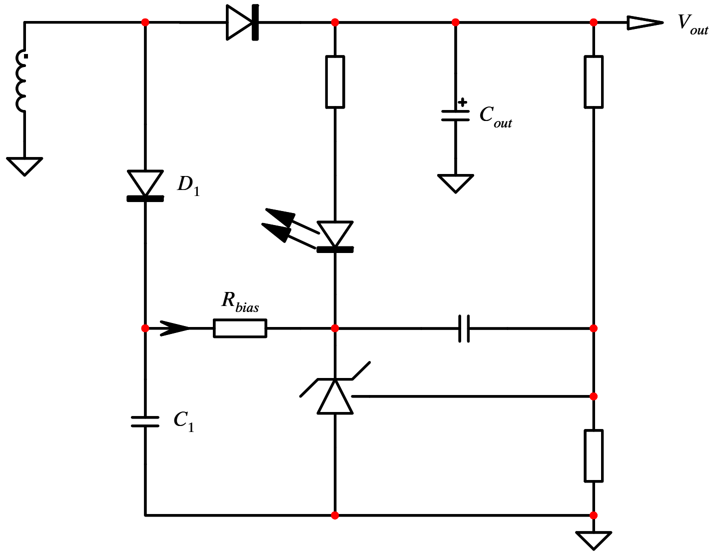


Figure 2. The Voltage across C_1 Collapses in Skip Cycle and the Bias Disappears on the TL431

In normal operating conditions, let's say nominal load, the voltage across C_1 equals that of C_{out}. R_{bias} is thus calculated to inject the needed extra bias current, 1 mA for instance (1 kΩ will do). When the load becomes lighter, the controller reduces its switching frequency down to 26 kHz. At this point, if the load current further goes down, the part enters skip cycle. In a no−load situation, as the distance between the pulses can be rather big, the amount of average charges brought to C_1 goes down and the ripple increases. At a certain point, the valley voltage drops and when the average voltage passes roughly below the output voltage minus 1 V (the LED drop), the bias current disappears, leaving the TL431 alone. At this point, the bias current is disconnected. When a transient load suddenly appears on the output, the bias current immediately rebuilds itself and provides the necessary bias to the TL431: the transient response does not suffer and remains identical to that delivered by Figure 1 solution. Figure 3 displays the transient response obtained with both solutions and confirms the validity of the proposed approach.

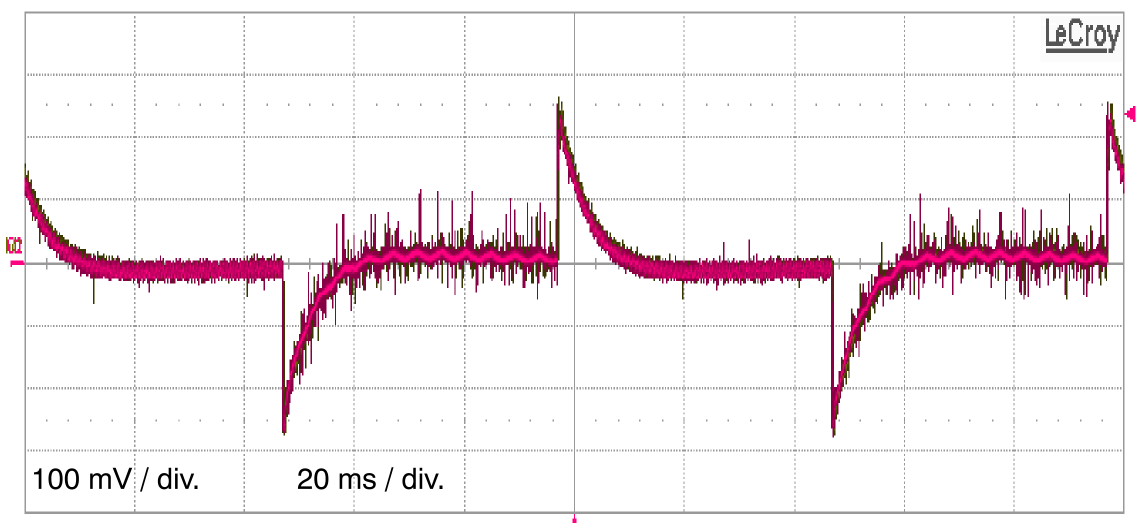


Figure 3. The Bias Suppression Circuitry Does Not Affect the Transient Response of the Adapter

We have performed no−load standby power measurements on a 65 W adapter with and without the bias suppression circuitry. With a fixed 1 mA bias, the consumption when supplied from a 230 V_{rms} source is around 110 mW. If you install the described embodiment, the input power drops to 87 mW, including the 2 MΩ discharge resistors: you pass the specs!

Please note that the circuitry is patented by ON Semiconductor but the company grants authorization to customers associating the circuit to one of its controlling device.

Connecting the LED in the Feedback Network

In some cases, an output LED is required by your end customer who wants to show that the adapter is powered and operational. Despite the efforts in selecting a high luminosity LED, the current necessary to enlighten it will cost you some input power loss. There is, however, a path that always requires some operating current, this is the feedback path. If you plan to keep injecting 1 mA in the TL431 as displayed in Figure 1, then the proposed solution appearing in **Figure 4** should interest you.

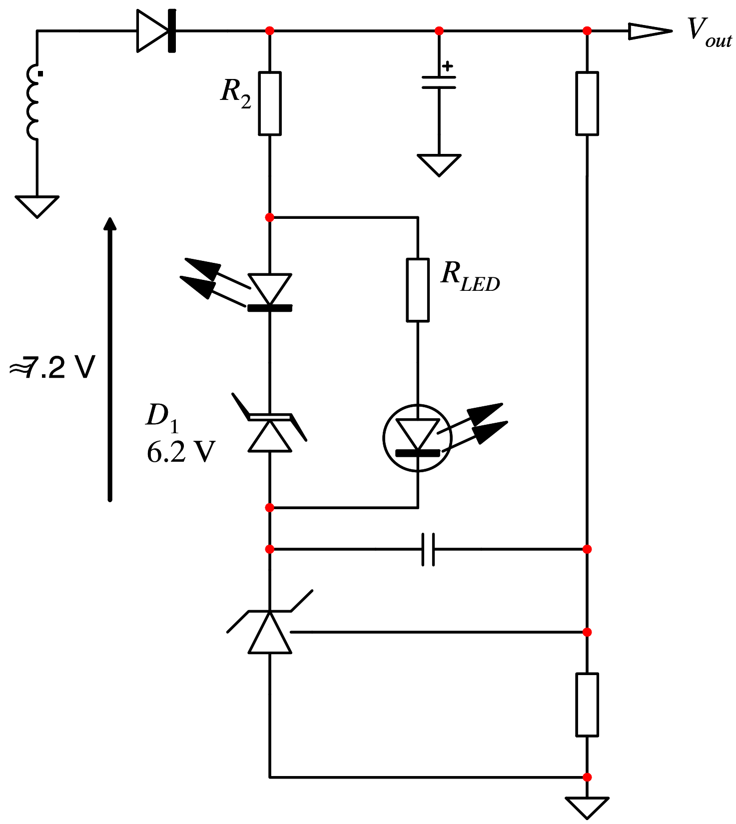


Figure 4. The Optocoupler LED Can Be Turned Into a Cheap Current Generator to Power the Output LED

By inserting a Zener diode in series with the optocoupler LED, we artificially expand the forward drop by the Zener voltage. If we consider a 1 V drop V_f for the optocoupler LED, adding a 6.2 V Zener in series with the device creates a total drop of around 7.2 V. This voltage stays almost constant regardless of the current circulating in the feedback line, except in no−load condition where it is slightly modulated by the skip operation. If we select a green LED (V_{LED} ≈ 2 V) and a 1 mA supply current, then the LED resistor is calculated as follows:

$$R_{LED} = \frac{V_f + V_Z - V_{LED}}{I_{LED}} = \frac{1 + 6.2 - 2}{\mathrm{lm}} = 5.2\ \mathrm{k\Omega} \qquad \text{(eq. 2)}$$

Experience shows that the circuit works well and the LED delivers an almost constant light. In no−load operation, the LED current weakens a little but after all, it is beneficial to the standby power. The small−signal response is not affected, unless the dynamic resistance of the Zener is really large (select a device capable to work with a small bias current). Make sure the output voltage and the choice of R_2 are compatible with the new minimum output voltage required to operate the loop. Before the addition of D_1, the minimum operating voltage was that of the TL431 (2.5 V) plus 1 V of the optocoupler LED. Adding a 6.2 V Zener diode brings this minimum operating voltage to 9.7 V.

You could also connect the green LED directly in series with R_2, after all, but the light would be permanently modulated by the feedback current. Not a real trouble, but some customers may not accept it.

Reducing the rms Output Current in Short−Circuit Conditions

After an idea from JP Louvel

The NCP1250 includes an internal timer. When the controller senses a demand for the maximum peak current setpoint (0.8 V/R_{sense} at zero OPP), a 100−ms timer starts to count down. If the condition disappears before the timer elapses, nothing happens and an internal reset occurs. On the opposite, if the timer reaches completion, all pulses are stopped and the auxiliary V_{cc} starts to come down. When it reaches VCC_{min}, the circuit enters sleep mode and reduces its total consumption below 15 μA. The start−up current now recharges the V_{cc} capacitor and lifts V_{cc} up towards the start−up level of 18 V. With a weak start−up current (to minimize losses), the operation can take several hundred of ms at the lowest input line. The part then re−starts and pulses for another 100 ms period in a recurrent manner. This is the so−called hiccup operation that appears in Figure 5.

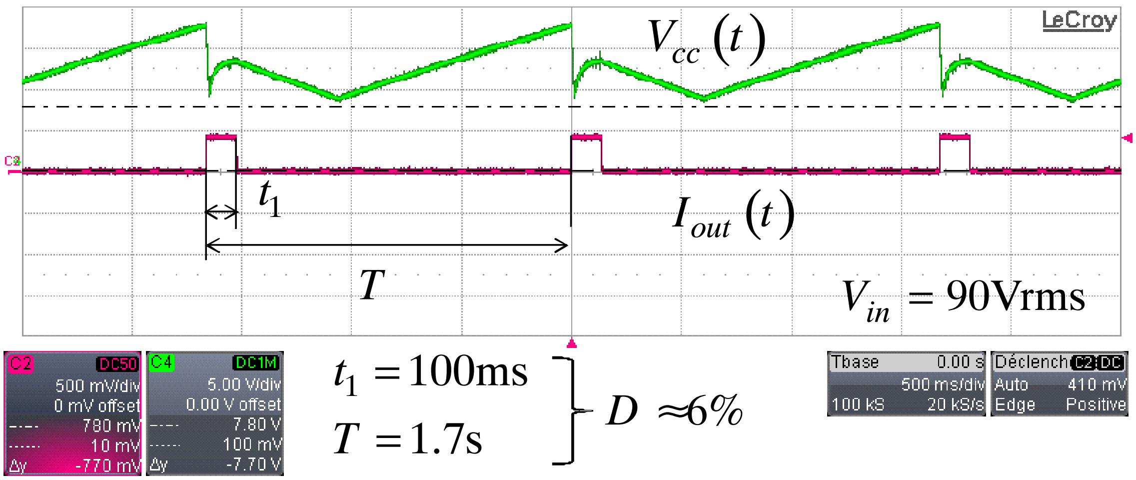


Figure 5. In Hiccup Mode, the Controller Tries to Re−start but Stops After the Timer Has Elapsed Since the Short−Circuit or the Overload is Still Present

As the circuit pulses for a 100 ms, a current circulates in the output cable and the load. It is a square wave as shown in Figure 6.

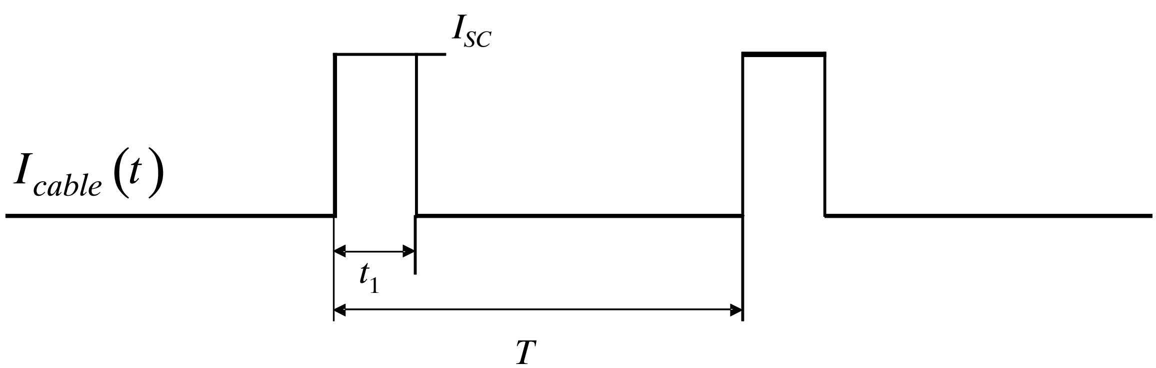


Figure 6. The Current Circulating in the Cable While the Load is Shorted Offers a Square Envelope

The Root Mean Square value of the signal envelope is given by:

$$I_{out,rms} = I_{SC}\frac{\sqrt{t_1}}{T} \qquad \text{(eq. 3)}$$

In the picture, the current peaks to 5 A. Applying Equation 3, the rms current amounts to:

$$I_{out,rms} = 5\sqrt{\frac{0.1}{1.7}} = 1.2\text{ A} \qquad \text{(eq. 4)}$$

Unfortunately, in high−line conditions, the recharge time of the V_{CC} capacitor is significantly shortened, naturally reducing the off−time duration and thus the recurrence T: the rms current in the cable while undergoing a short−circuit is increased.

One way to reduce the rms content is to act upon the secondary−side peak current. Alternatively, the off−time T can also be lengthened by an external means. The solution that appears in Figure 7 adopts this solution. The principle is simple: at start−up, in absence of auxiliary voltage, Q_1 is blocked and V_{cc} normally takes off thanks to the current delivered by the start−up network. When the part starts to pulse, the auxiliary voltage biases Q_1 that interrupts the charging current. As the part is already operating and owing to the diode D_3, it has no influence on the converter. When a fault occurs, the controller stops pulsing. Thanks to C_1, Q_1 is still biased and blocks the refueling of the V_{cc} capacitor: the charging current is now diverted to ground. The time constant involves R_3, R_2 and C_1. The division ratio must be selected so that the transistor remains biased until V_{aux} equals 8 V or less if long durations are expected. At this point, the voltage is below the controller UVLO and the circuit re−starts. By tweaking the time constants, you have a means to extend the off−time duration and nicely reduce the output cable rms current. We have captured some oscilloscope shots while this circuit was operating. The results appear in Figure 8.

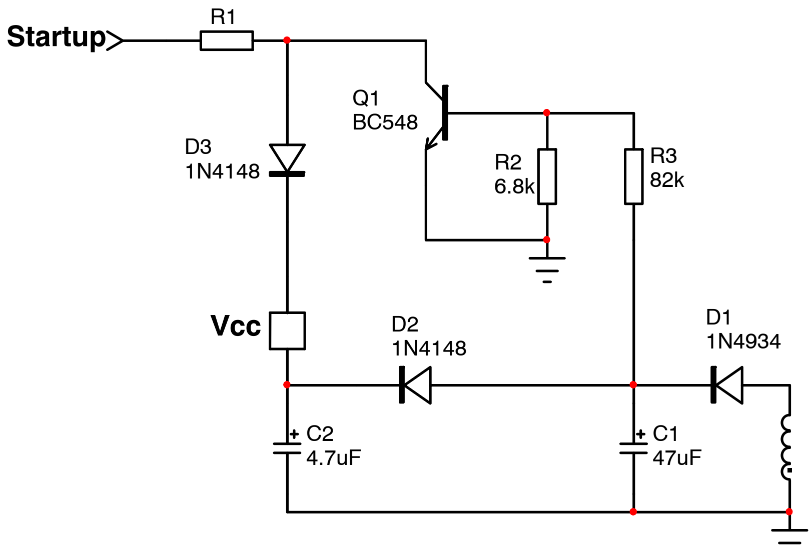


Figure 7. A Simple Bipolar Transistor Can Do the Job of Extending the Controller Off−Mode Period

In the figure, the duty−cycle has been reduced to 3.8%, bringing the rms current to a value equal to:

$$I_{out,rms} = 5\sqrt{\frac{0.1}{2.6}} = 980 \text{ mA} \qquad \text{(eq. 5)}$$

In high−line conditions, the repetition rate goes from 0.9 s (original circuit) to 1.6 s (when the bipolar transistor is added), bringing the duty−cycle from 11% down to 6%.

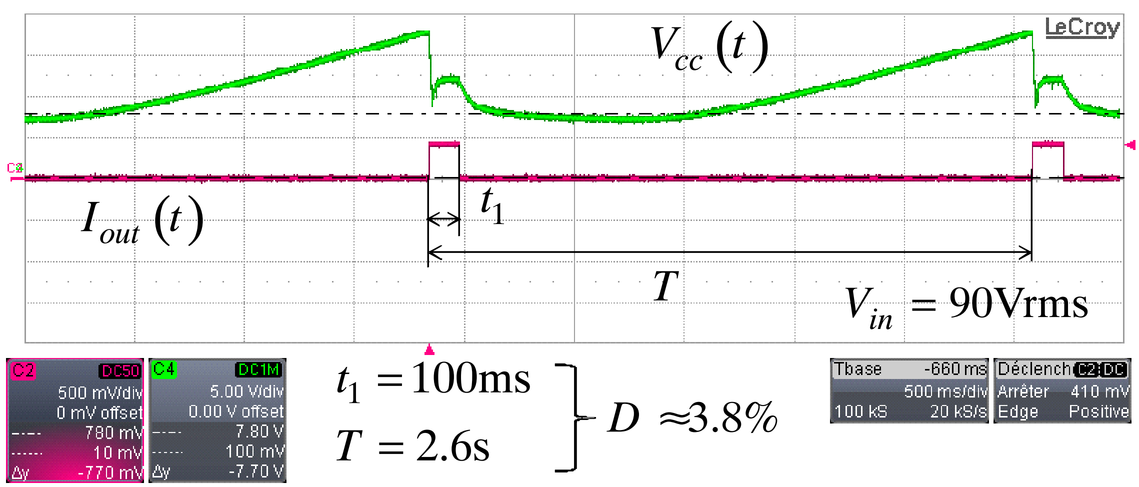


Figure 8. A Simple Bipolar Transistor Can Do the Job of Extending the Controller Off−Mode Period

Discharging the X2 Capacitors

As recommended by the IEC−950 safety standard, the voltage across the male terminals of the power cord, when unplugged, must decrease at a pace where it reaches a non−lethal voltage in less than 1 s. To meet this goal, so−called discharge resistors are installed across the capacitor and ensure its natural discharge when the power cord is unplugged. The standard does not define an ending voltage but simply specifies the equivalent *RC* time constant that at maximum must exist: 1 s. If we have a 0.47 μF capacitor, a resistor of 2 MΩ will fit the bill. After 1 s, if the capacitor is charged to 330 V dc (230 V_{rms} in), its voltage will reach:

$$V_c(t) = V_{dc} e^{-t/RC} = \frac{330}{e} = 121\ \mathrm{V} \qquad \text{(eq. 6)}$$

or 36% of its initial charge. Unfortunately, when biased at 230 V_{rms}, these resistors will dissipate a permanent power of ≈26 mW, plaguing your low−standby power design. Since these resistors bother us, why not disconnecting them in normal operation? We could then put them back across the capacitor only when the plug is disconnected. This is a solution proposed in Figure 9. In this drawing, D_5 and D_6 route the high−voltage input signal to a differentiator made of R_1 and C_9. When a positive slope is sensed by this network, it positively biases transistor Q_1: Q_2 is permanently maintained in a blocked state as its gate−source voltage refuses to take off. When the user pulls the plug, the voltage across the X2 capacitor no longer changes polarity: the capacitor electrical state freezes, charged in a certain configuration. As the voltage transition has left, Q_1 no longer receives "reset" pulses and C_{10} can now be charged through R_3. When the voltage reaches the enhancement level for Q_2, it turns on and brings the diodes cathode to ground via the 1−MΩ resistor. If resistors R_1 and R_3 are selected sufficiently high, their impact on standby power is negligible. C_{10} is selected to offer a certain immunity to input voltage dropouts and can be adjusted if necessary. C_9 must be a 1 kV type of capacitor.

This circuit has been assembled and tested on a real board. The resulting waveforms appear in Figures 10 and 11. Despite a low input voltage, the monitoring circuit still works fine and maintains Q_2 blocked. At high line, the charging current directly derived from the bulk capacitor reduces C_{10} charging time and induces a small sawtooth across Q_2 gate−source. However, the level is low enough to avoid false tripping of the circuit. When the converter is un−plugged from the wall outlet, the voltage no longer swings up and down on D_6D_5 cathodes and Q_1 blocks. C_{10} charges until Q_2 conducts and provides a discharge path to the X2 capacitor. The 600 V MOSFET is a 1 A type but a TO92 version can also work since the current involved in the discharge process is weak.

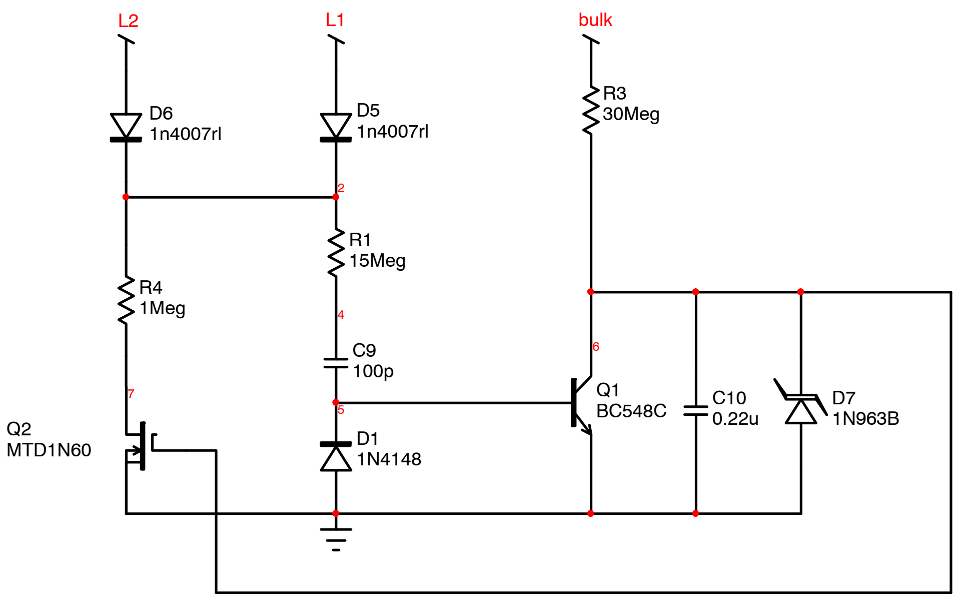
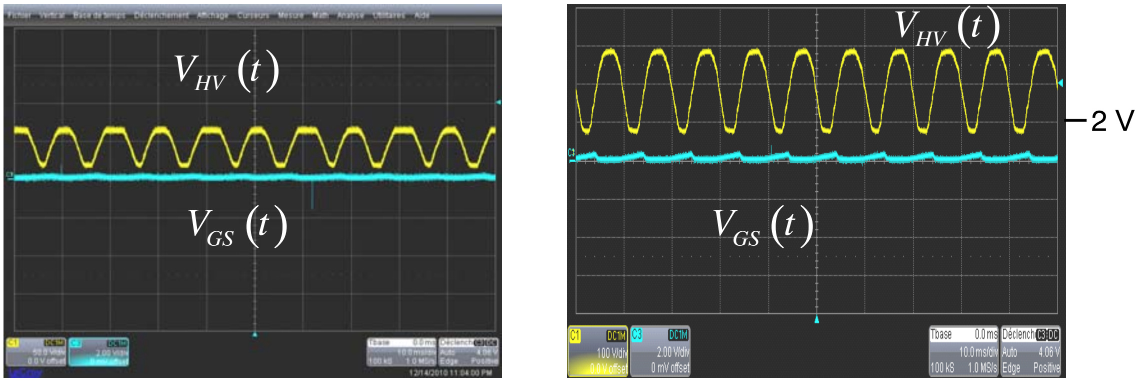
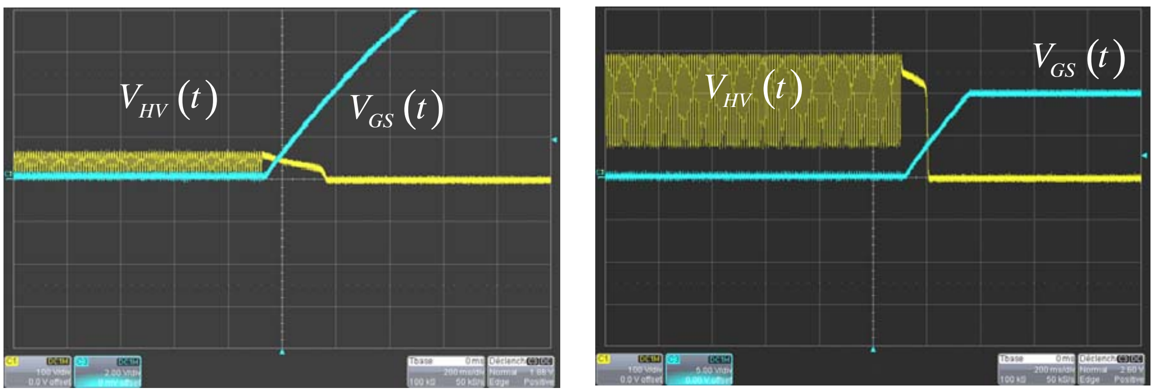


Figure 9. A Bipolar is Periodically Activated by the Input Signal Slope and Keeps the MOSFET Off in Normal Operation



$V_{in,rms} = 45V$ $V_{in,rms} = 230V$

Figure 10. As Long as the Mains is Present, The V_{GS} Voltage of Q_2 is Kept Low Enough to Block It



$V_{in,rms} = 45V$ $V_{in,rms} = 230V$

Figure 11. When the Mains Disappears, the Voltage on Q2 Gate−Source Terminals Quickly Rises Up and Triggers the X2 Capacitor Discharge

An overview of the complete circuit appears in Figure 12.

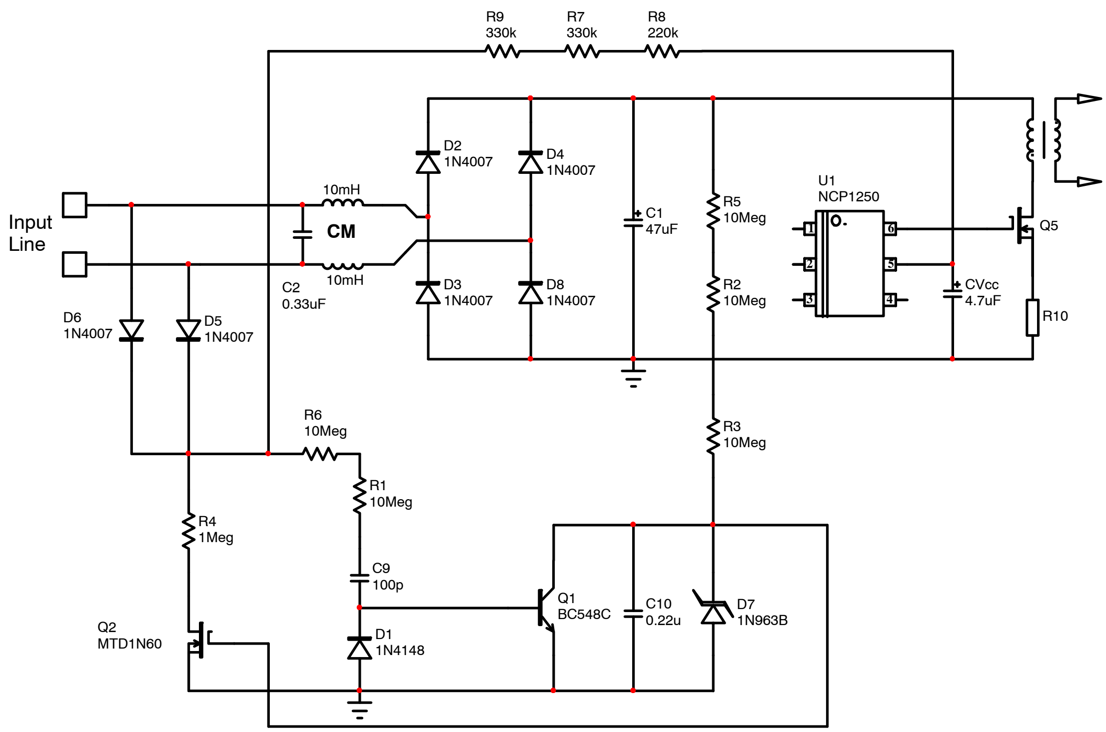


Figure 12. The Discharge Circuit Easily Inserts Into an Existing Design

Start−up Resistors Used as a X2−Discharge Network

In the above circuitry, we suppressed the discharge resistors and replaced them by an active circuit. Unfortunately, we still need a start−up network to crank the controller. This is the resistors string made of R_9, R_7 and R_8. An idea is to combine the discharge path and the V_{cc} capacitor charging current. If we consider a 2 MΩ network, a solution is to wire it as recommended by Figure 12.

In this particular configuration, the charging current is simply:

$$I_{chg} = \frac{V_{peak}}{\pi\left[\left(R_1 + R_2\right) \parallel \left(R_3 + R_4\right)\right]} \qquad \text{(eq. 7)}$$

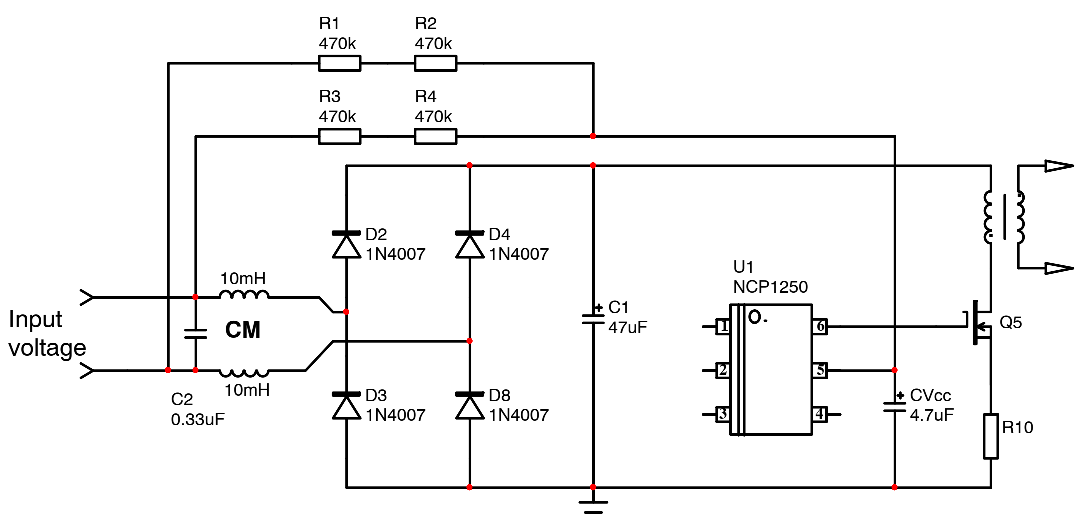


Figure 13. By Joining the X2 Resistors Terminals to the NCP1250 VCC Pin, it Naturally Combines the Discharge and Start−up Functions

For a 85 V_{rms} input voltage and neglecting the V_{cc} level, this charging current amounts to:

$$I_{chg} = \frac{V_{peak}}{\pi\left[\left(R_1 + R_2\right) \parallel \left(R_3 + R_4\right)\right]} = \frac{85 \times \sqrt{2}}{3.14 \times 470k} = 81\ \mu A \quad \text{(eq. 8)}$$

Suppose we have a V_{CC} capacitor of 4.7 µF. In that case, considering a start−up level of 20 V max. and assuming a constant charging current, the start−up time would be:

$$t_{start-up} = \frac{VCC_{on}C_{VCC}}{I_{chg} - ICC1} = \frac{20 \times 4.7\mu}{81\mu - 15\mu} = 1.4s \quad \text{(eq. 9)}$$

Simulation gives 1.7 s as it accounts for the current reduction as V_{CC} rises up.

In the displayed configuration, only one resistive branch is involved for the discharge process. Therefore, if needed, resistors values can be further increased to reduce the standby power consumption. For instance, if we use 1.5 MΩ resistors, the charging current is reduced to roughly 50 µA and the simulated start−up time reaches 2 s with the same 4.7 µF capacitor.

Always make sure the injected current in the V_{CC} pin stays above 60 µA (design margin) at the lowest input line. The 2 x 1 Meg resistors are a recommended solution since they increase the injected average current via a dual-wave connection.

A Simple Brown−Out Circuit

A brown−out protection circuit is a means to protect the power supply against a low−voltage operation. However, thanks to the overload protection, a power supply fed from a low−voltage source is likely to protect itself by entering a hiccup mode. If this operation does not bother the vast majority of loads, e.g. notebooks, some applications cannot accept a hiccup mode. Printers fall within this category as a hiccup operation could lock their micro−controller in an undetermined mode. As a result, most of printer manufacturers forbid hiccup and requires clean single−shot on and off sequences. To cope with this requirement, a simple brown−out circuit has to be installed on the NCP1250.

Unlike a classical brown−out circuitry defining a valid input voltage range, our circuit blocks the auto−recovery operation in presence of a regulation fault at a low input voltage. When the input voltage is too low, the converter cannot start−up and is actively blocked by a transistor. As soon the input voltage grows and reaches the level you have selected, the controller pulses and delivers power. At this point, the controller is self−supplied and its start−up current is no longer diverted to ground: full auto−recovery is available. If the input voltage goes down and the output is maintained, fine, no fault is detected and the converter keeps operating. If for any reason the load increases and exceeds the converter capability, the time−delayed fault protection activates and stops the pulses, initiating a cycle of auto−recovery. However, if this cycle happens at a low input voltage, outside the level for which the converter is authorized to start−up, the start−up current is diverted to ground and nothing happens: the converter is latched and there is no hiccup. If the input voltage now increases back and crosses the start−up threshold, the controller starts−up normally and the load is fed. This ensures a clean start−up at full load and a clean cutoff as the input voltage diminishes. Figure 13 presents the adopted solution where two simple bipolar transistors are enough to do the job.

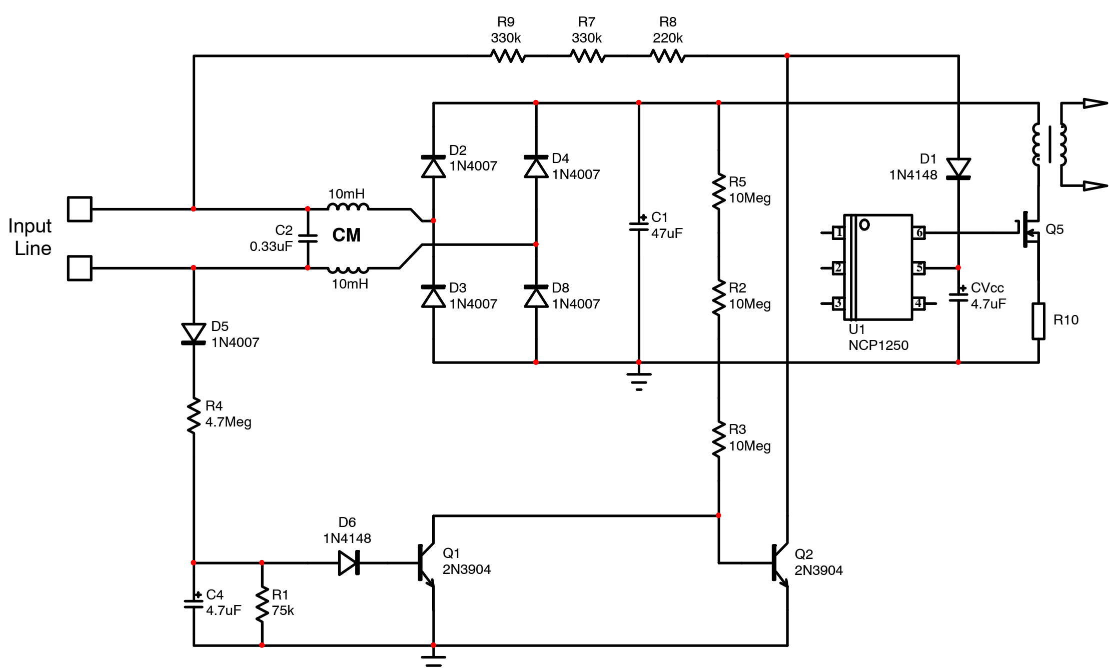


Figure 14. Two Bipolar Transistors Are Enough to Implement a Brown−out Protection

This circuit has been added to the NCP1250 65 W demonstration board and has proved to work ok. With the component values put in the schematic, the converter starts−up at V_{in} = 78 V_{rms} and I_{out} = 3.3 A. For a 3 A load, it stops working for V_{in} = 65 V_{rms}, going down to 41 V_{rms} for I_{out} = 2 A.

Conclusion

This application note describes how a 6−pin controller housed in a TSOP6 package expands its capabilities when adding a few components around it. The proposed ideas are just a few examples our application engineering thought about when dealing with day−to−day customers problems. As more application ideas are tested and documented, we will update this application note to make it an evolving document.

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