

## HOW TO HANDLE SHORT CIRCUIT CONDITIONS WITH ST'S ADVANCED PWM CONTROLLERS

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*Purpose of this note is to provide some advice on how to manage a short circuit condition for a switching converter controlled by ST's advanced PWM controllers L5991 and L5993.*

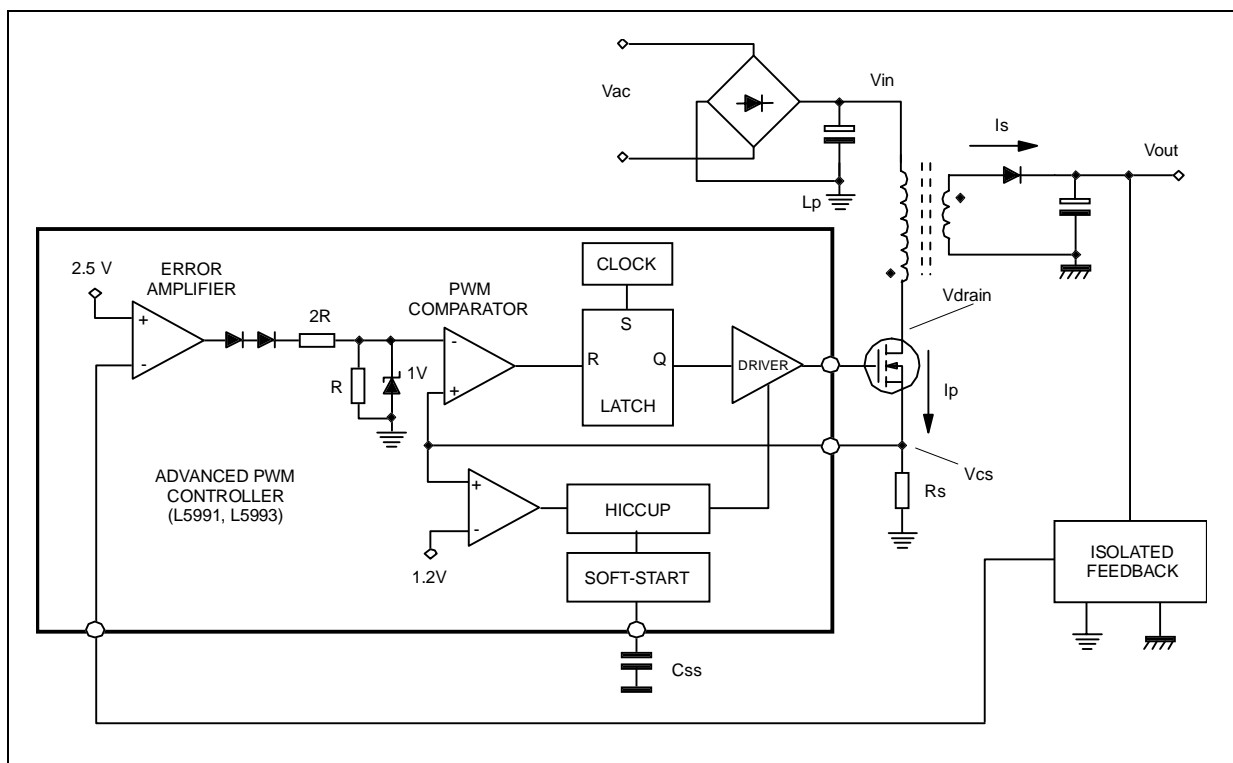
*After a quick review of merits and limits of the current limitation functionality embedded in these IC's, a simple circuit will be given that allows to fulfill designer's typical requirements on what the converter is supposed to do in case of overload or short circuit.*

### Current limitation schemes: merits and limits

ST's advanced PWM controllers L5991 and L5993 [1] [2] include a two-level overcurrent protection: the "pulse-by-pulse" (or "cycle-by-cycle") current limitation and the so-called "hiccup mode" operation.

Pulse-by-pulse current limitation ideally limits the peak primary current at  $1/R_s$  (see figure 1) when the converter is overloaded or there is a load short circuit. Even neglecting non-idealities, which worsen the scenario and must be accounted for in a safe design, pulse-by-pulse current limitation just prevents the system from losing control of the peak primary current. Nothing more.

**Figure 1. PWM control and current limiting in L5991 and L5993.**



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The output current is not limited and may cause PCB traces burning. Converter's power throughput declines only slightly. Furthermore, power losses get concentrated on some specific parts, e.g. the output rectifier(s), which may overheat and go towards destruction if not rated for this extreme condition and if heatsinking capability is not generously provided.

When pulse-by-pulse current limitation cannot maintain the peak primary current under control, hiccup mode protection takes over and halts the resulting current runaway. This happens when the switch ON-time needed to control the peak primary current is lower than the minimum value  $T_{ONmin}$  that the controller can generate as a result of its internal delays. In L5991 and L5993  $T_{ONmin}$  can be estimated as high as 200 ns, typical value. More details on this "runaway condition" can be found in the appendix.

With reference again to figure 1, in case of current runaway the voltage on the current sense pin will go over 1V. As it reaches 1.2V a comparator triggers this special functionality [1], [2].

Unlike pulse-by-pulse limitation, hiccup mode operation keeps output current and power throughput very low. It would be desirable, then, to rely on hiccup operation to conveniently handle a load short circuit. However, it is quite common that even a dead short at the converter's output cannot activate hiccup protection.

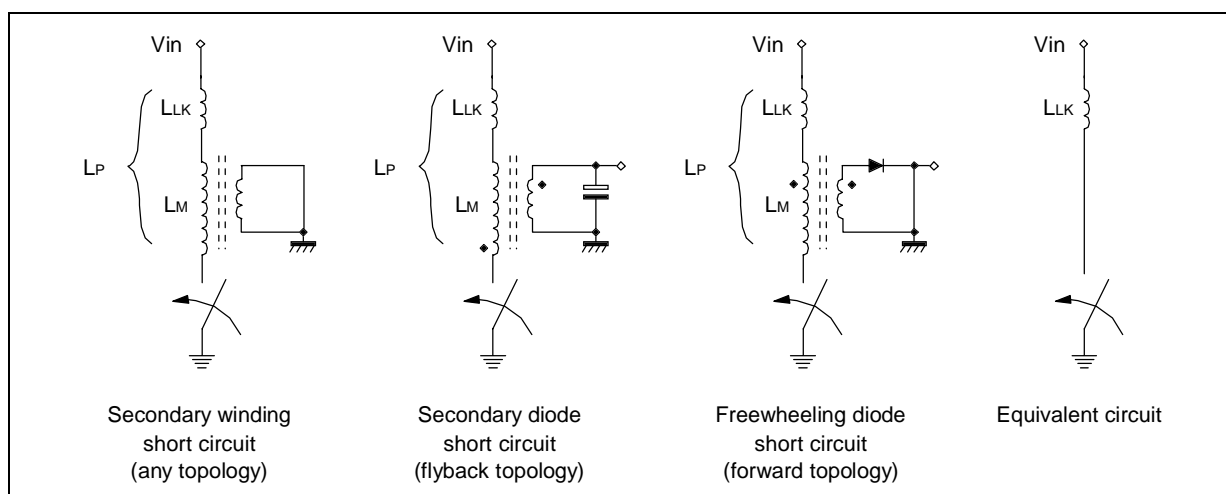
Referring to the appendix for a detailed analysis, high input voltage, high regulated output voltage and high switching frequency favour current runaway and, therefore, hiccup operation to be tripped further to a load short circuit. In many applications, such as AC-DC adapters or PC's silver boxes, where the output voltage is not so high and the switching frequency is quite low (below 100 kHz) it is extremely likely that the runaway condition will never be met and the hiccup protection will never be activated. Furthermore, in these applications the load is usually connected through long cables, then the total short circuit resistance is not low and the  $T_{ON}$  needed to control peak current needs not be so short.

In other applications, such as monitor SMPS, the runaway condition is likely to be met for the highest output voltages, especially at high input voltage and if they are synchronized at high frequency.

Actually, hiccup protection serves two different purposes. The first one is related to the capacity of the L5991 and L5993 to be operated at a very high frequency (up to and beyond 500 kHz): to guarantee a safe handling of a load short circuit under all conditions of operating frequency, hiccup protection must be added to pulse-by-pulse limitation. The second purpose is to cover some fault conditions that pulse-by-pulse limitation cannot handle. These are basically:

1. short circuit of the secondary winding (regardless of the topology);
2. secondary catch diode short circuit (in flyback converters);
3. secondary freewheeling diode short circuit (in forward converters).

**Figure 2. Fault conditions that activate hiccup mode operation.**



All of these conditions reflect a short circuit to the transformer's primary side while the switch is ON. The primary current rate of rise is then limited only by the leakage inductance of the transformer, which is few percent of the primary magnetizing inductance (see fig. 4). This means that the current slope will be 30-50 times higher than in normal operation, thus pulse-by-pulse limitation has no chance to control the current.

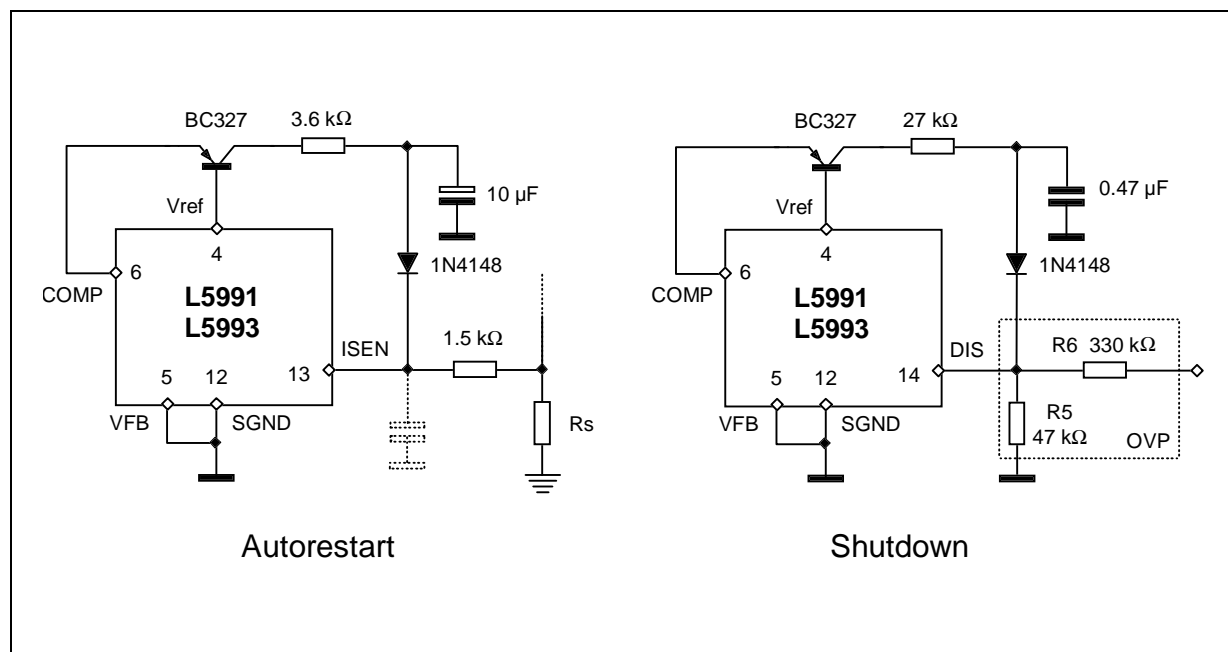
### Overcoming pulse-by-pulse issues

With the aim of reducing power consumption during an overload or a load short circuit, it is possible to select one of the following two basic strategies:

1. Making the converter enter hiccup mode operation as a fault is detected and letting it go back into normal operation automatically as the fault is removed (autorestart).
2. Shutting down the converter as a fault is detected and keeping it off until it is disconnected from the mains. Only after a power-off / power-on cycle the converter is enabled to restart.

Both of these strategies can be easily implemented with ST's advanced family of PWM controllers, thanks to the numerous functions available. Figure 3 shows how this can be done with the same basic circuit comprising only four cheap external parts.

**Figure 3. Different strategies for low dissipation during load short circuit**



When in current limitation because of an overload or a load short circuit, the output voltage drops below its regulated value and the output of the error amplifier (pin 6, COMP) saturates high ( $\geq 6V$ ) in the attempt of recovering voltage regulation. This is used for turning on the small-signal PNP (its base is tied at 5V) and getting a current available for triggering a protection functionality. However, a delay of several ms is needed to prevent the function from being activated at start-up or as a result of a load surge. This is simply done with an RC cell. Finally a diode is needed to decouple the circuit from the others connected to the same pin, during normal operation.

Hiccup is activated by pulling the current sense pin (ISEN, 13) above 1.2V. Error amplifier source capability with high saturation voltage is about 1 mA, thus the resistor between the current sense resistor and ISEN pin is increased from 1 to 1.5 kΩ. The 3.6 kΩ collector resistor and the 10 μF capacitor form a delay cell of about 15 ms.

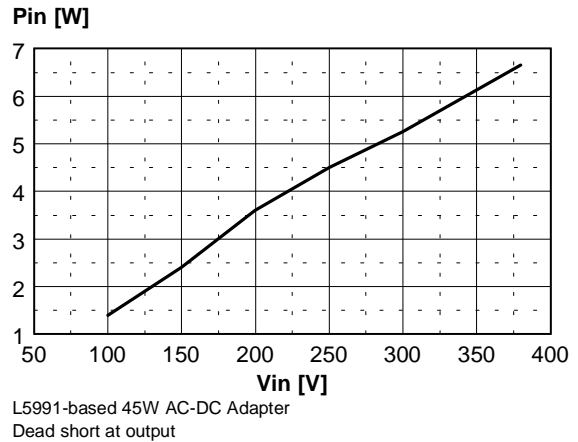
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This circuit has been implemented in a 45W, wide range mains AC-DC Adapter with the L5991 [3] and the results are shown in fig. 4.

Shutdown will be activated by injecting the current delivered by the PNP into pin 14 (DIS) so that the voltage at the pin exceeds 2.5V. The resistor divider R5-R6 refers to the circuit described in [3] and is part of the latched OVP circuit. In case of a different value of R5, the collector resistor and the delay capacitor will be changed consequently.

Besides increasing safety, the advantage of each of these techniques is that heatsinking of the power parts needs to be designed neglecting anomalous operating conditions (overload or short circuit), with a considerable saving of size and cost.

**Figure 4. Input power vs. input voltage during hiccup mode operation**



### Conclusions

Pulse-by-pulse and hiccup mode current limitation are the ways ST's advanced PWM controllers L5991 and L5993 handle overload and short circuit conditions. It has been shown that these two functions cover two different fault categories: pulse-by-pulse does not allow the power throughput to rise out of control in case of a load failure whereas hiccup usually protects against short circuits inside the converter itself (transformer, diodes) where pulse-by-pulse is not effective.

Unlike hiccup, which dramatically reduces power throughput, pulse-by-pulse limitation little does about that, thus forcing designers to oversize heatsinking of the critical parts for the sake of safety.

A simple and versatile circuit has been presented that helps overcome this problem and can be used to achieve two common protection schemes, autorestart and shutdown, resulting in very low power dissipation.

As an appendix, a calculation method that helps evaluate whether a load short circuit can activate hiccup mode protection or not in a given application, is provided.

### REFERENCES

- [1] "L5991/A Primary Controller with Standby" Datasheet
- [2] "L5993 Constant Power Controller" Datasheet
- [3] "45W AC-DC Adapter with Standby Function" (AN1134)

## APPENDIX

**Will hiccup mode be activated or not ?**

Basically, hiccup mode will be activated if the control is not able to guarantee a steady-state operation for the converter under a given short circuit condition.

Like in normal operation, under overload or short circuit conditions a steady-state operation requires volt-second balance across the storage element of magnetic energy: the transformer in a flyback converter and the output inductor in a forward. Otherwise, current runaway will occur: energy rise inside the magnetics during MOSFET's ON-time exceeds energy decay during OFF-time, thus current will raise with no control saturating the magnetics and leading to converter destruction.

The condition for the current runaway to occur (runaway condition) will be derived for both flyback and forward topology. Please refer to the end of the appendix for symbol explanation.

*Flyback topology.* The balance condition (assuming the system is working in continuous current conduction, which is extremely likely under overload condition and absolutely certain in short circuit) is:

$$V_{in} \cdot T_{ON} = V_R \cdot (T_{sw} - T_{ON}) \quad (A1).$$

Solving eqn. (A1) for  $T_{ON}$  yields the switch ON-time. This is applicable during both normal operation and pulse-by-pulse current limitation:

$$T_{ON} = \frac{V_R}{V_{in} + V_R} \cdot T_{sw} \quad (A2).$$

It is worthwhile reminding that the reflected voltage  $V_R$  is given by:

$$V_R = n \cdot (V_o + V_F) \quad (A3).$$

Whereas in normal operation  $V_o$  is regulated and, therefore,  $V_R$  is constant, in case of overload or short circuit  $V_o$  drops and so does  $V_R$ . Their value will be dependent on the total resistance of the secondary circuit, including diode, secondary winding wire, PCB tracks, and the short circuit resistance. Obviously, the more severe the overload, the lower the total resistance, and the lower  $V_o$ .

Eqn. A2 shows that, as the overload gets closer to a short circuit,  $T_{ON}$  gets shorter and shorter. If the time resulting from (A2) is lower than  $T_{ONmin}$ , the shortest switch ON-time that the controller can generate (200 ns typ. in the L5991 and L5993), then the controller will not be able to maintain the volt-second balance and the current will begin to rise with no control.

Combining equations (A2) and (A3), the runaway condition, that is the condition for hiccup operation to be invoked, is:

$$\frac{n \cdot (V_o + V_F)}{V_{in} + n \cdot (V_o + V_F)} \cdot T_{sw} \leq T_{ONmin} \quad (A4)$$

It is possible to estimate the left side of eqn. A4 assuming  $V_o = 0$  (ideal short circuit). In real-world condition  $V_o$  will be greater than zero thus  $T_{ON}$  will be in this way underestimated.

Ideally, eqn. (A4) becomes:

$$\frac{n \cdot V_F}{V_{in} + n \cdot V_F} \cdot T_{sw} \leq T_{ONmin} \quad (A5)$$

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In this equation it is reasonable to assume  $V_F = 1V$  for a Schottky diode and  $V_F = 1.5 V$  for a p-n type.

The inspection of (A5) shows that the conditions that favour current runaway are a high input voltage  $V_{in}$ , a high switching frequency (i.e. a shorter  $T_{sw}$ ) and a high regulated output voltage: in fact, being  $V_R$  fixed by MOSFET breakdown issues (typically around 100V), the higher  $V_o$ , the lower  $n$ .

*Forward topology.* The balance condition (continuous current conduction is in use even in normal operation) for the output inductor is:

$$\left(\frac{V_{in}}{n} - V_F - V_o\right) \cdot T_{ON} = (V_o + V_F) \cdot (T_{sw} - T_{ON}) \quad (A6)$$

Solving eqn. (A6) for  $T_{ON}$  yields the switch ON-time:

$$T_{ON} = n \cdot \frac{V_o + V_F}{V_{in}} \cdot T_{sw} \quad (A7).$$

Considering equation (A7), the runaway condition for a forward converter is:

$$n \cdot \frac{V_o + V_F}{V_{in}} \cdot T_{sw} \leq T_{ONmin} \quad (A8)$$

Ideally, with  $V_o = 0$ , this equation becomes:

$$\frac{n \cdot V_F}{V_{in}} \cdot T_{sw} \leq T_{ONmin} \quad (A9)$$

where it can be assumed again  $V_F = 1V$  for a Schottky diode and  $V_F = 1.5 V$  for a p-n type.

Like in flyback topology, the conditions that favour current runaway are a high input voltage, a high switching frequency and a high regulated output voltage. However, comparing equations A5 and A9 under the same conditions of  $n$ ,  $V_F$ ,  $V_{in}$  and  $T_{sw}$ , it is possible to note that forward converter is less prone than flyback to current runaway. This is reinforced by the fact, not visible in the above equations, that in a forward converter the output inductor resistance will always be part of the short-circuit current loop.

### Symbol list

Symbol	Description
$V_{in}$	Converter's DC input voltage.
$V_o$	Converter's output voltage.
$T_{ON}$	Power switch ON-time.
$T_{ONmin}$	Minimum switch ON-time manageable by the PWM controller IC.
$V_R$	In flyback, secondary voltage reflected back to the primary during secondary rectifier conduction.
$T_{sw}$	Switching period, equal to the reciprocal of the converter's switching frequency.
$n$	Transformer's primary-to-secondary turn ratio.
$V_F$	Forward drop across secondary rectifiers (catch and freewheeling diode)

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