AN6236

Application note

Overcurrent protection

Introduction

Current measurement has now become essential in any electronic system. One of the key points to prevent damage to an application is the ability to measure current variations very quickly and accurately. There are several ways to implement an overcurrent protection, with the oldest and simplest being the well-known fuse. New and more efficient solutions also exist.

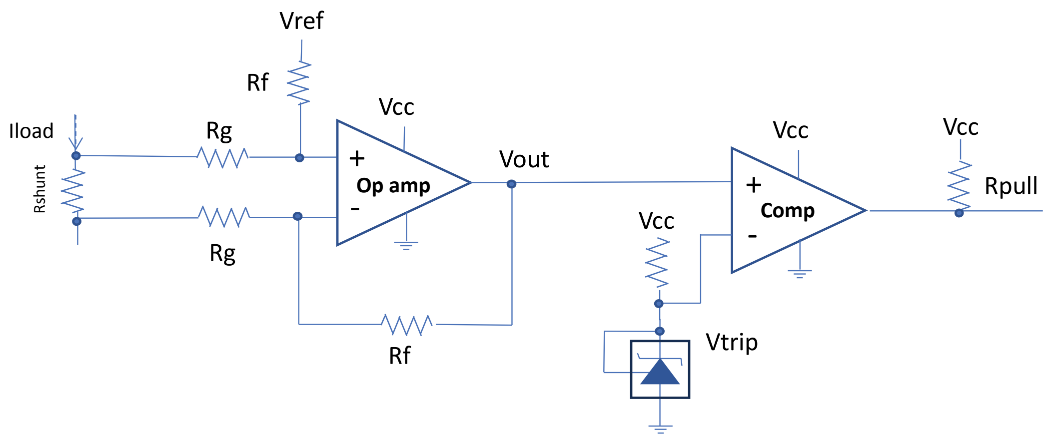
This application note focuses on overcurrent protection based on operational amplifiers or current sensing, which can measure and amplify a current flowing into a shunt resistance, and comparator that can trigger an event. It describes how to determine the expected response time to an overcurrent event and proposes different kinds of architecture with their pros and cons.

1 Overcurrent protection (OCP) function

- Operational amplifier and comparator

The simplest option is to build an OCP function with an op amp and a comparator available off the shelf. As presented in Figure 1, the first stage describes an op amp with four external gain resistors which can be considered current sensing.

Figure 1. **OCP function with an op amp and a comparator**



Its output voltage gives a linear answer to the current flowing into the shunt, following Equation 1.

$$Vout = Iload * Rshunt * \frac{Rf}{Rg} + Vref \quad (1)$$

The Vout voltage is then compared to a Vtrip voltage, which represents the triggering current Equation 2.

$$Vtrip = Itrigger * Rshunt * \frac{Rf}{Rg} + Vref - Overdrive \quad (2)$$

The Vref voltage allows shifting the output voltage, in order to avoid any saturation when no or extremely small current is flowing into the shunt resistor.

Overdrive is an important point to consider as it allows improving the response time of a comparator. The higher the overdrive, the faster the response time.

The Vtrip voltage can simply be realized with a divider bridge or with a reference voltage such as the TS432, for a more precise measurement.

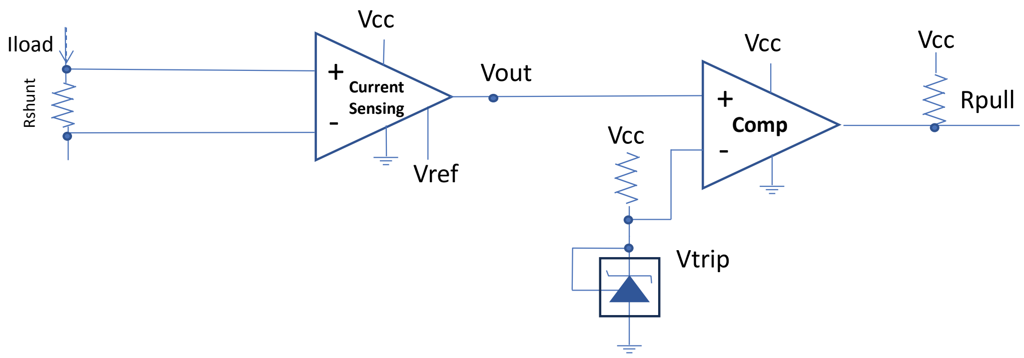
This solution is very flexible and can be very cost-effective by choosing standard products. To maximize cost reduction, a divider bridge can be used instead of the reference voltage to fix the triggering voltage of the comparator. Choosing a high-speed amplifier, a fast comparator, and accurate external components, allows achieving very efficient OCP function, with very good response time and precision. The main advantage of using an external reference voltage such as the TS432A is to increase the precision, as its accuracy is independent of the Vcc supply.

As this solution is op amp based, it is more commonly used in low-side current sensing applications. High-side can also be considered by using a rail-to-rail input op amp and having an input common-mode voltage Vicm ≤ Vcc.

- Current sensing and comparator

With the same approach as above, the op amp stage can be replaced with a current sensing amplifier as described in Figure 2. One of the main advantages is that the current sensing amplifier integrates the gain resistors, allowing better precision and PCB area saving.

Figure 2. OCP function with a current sensing and a comparator



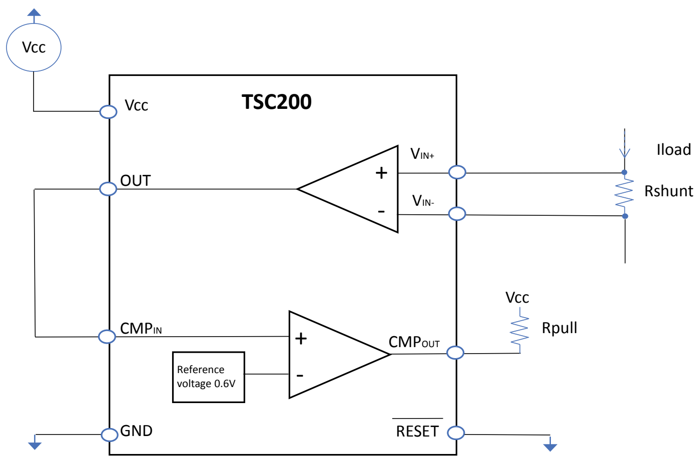
Another important point is that the current sensing amplifier can work with a common-mode voltage much higher than its power supply. This makes the device very useful for high-side current sensing applications, for example measuring the current in a 24 V rail while being supplied by 5 V. Depending on their Vicm range it can also be used for low-side current measurement.

- All-in-one

A third option is to use a dedicated device which integrates a current sensing amplifier and a comparator in the same package, such as the TSC200 as described in Figure 3. It is designed to measure current by amplifying the voltage across a shunt resistor connected to its input. Thanks to a bandgap reference (0.6 V) and a comparator which can latch the output, it acts as an overcurrent protection device.

TSC200 is a fixed-gain current sensing amplifier of 20 V/V. It provides an extended input common range from -16 V below negative supply voltage and up to 80 V, while the device can operate from 2.7 to 18 V. It can operate either as low-side or high-side current sensing.

Figure 3. OCP function with a dedicated device TSC200



This solution offers the great advantage of using extremely few external components allowing strong optimization of the PCB area and easier supply chain management. Integration of gain, reference voltage, and comparator allows for a better approach as well as for accurate measurement, as it is less temperature-dependent and has better mismatch.

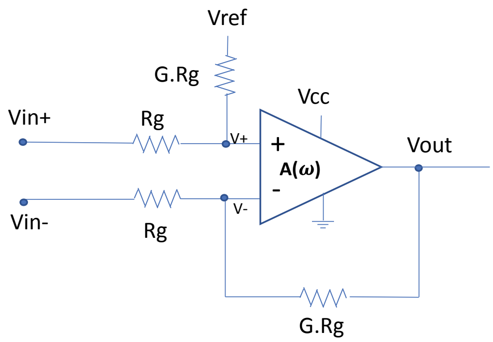
2 Response Time

One of the most critical points to prevent any damage to the application is the response time of such an OCP function to an event. Effectively, the faster the overcurrent is detected, the better the safety. The best choice seems to be a high-speed device. But when looking into a datasheet of an op amp or a current sensing amplifier, the response time is not clearly written. What is the most important parameter? An initial idea might be to consider the slew rate as a key parameter because it represents the maximum rate of variation that an amplifier can reproduce, but it is not the only parameter involved in the response time of the op amp.

Let us have a look at a mathematical approach to a differential op amp as described in Figure 4 in order to understand with equations how the output changes over time.

The gain is achieved thanks to four external resistances. An external Vref voltage can be applied to allow a bidirectional current sensing mode, or simply to avoid any output saturation when no current is flowing into the shunt, as the op amp is a single-supplied voltage.

Figure 4. Differential amplifier



A(ω) is the open-loop transfer function of the op amp and is defined at first order with Equation 3.

$$A\left(\omega\right) = \frac{AVD}{1 + \frac{j\omega}{\omega 0}} \tag{3}$$

AVD is the open-loop gain of the op amp at low frequency or DC level and the dominant pole of the op amp is given by $\omega 0 = \frac{2\pi.GBP}{AVD}$

Considering $(V_+ - V_-)^*A(\omega) = Vout$; we can write Equation 4.

$$G.Vdiff = Vout^*\left(\frac{1+G}{AVD} + 1\right) + \left(\frac{1+G}{2\pi GBP}\right)^* j\omega Vout - Vref \tag{4}$$

Considering $j\omega Vout = \frac{\partial Vout}{\partial t}$ and AVD >> 1

We can write the following differential equation (5).

$$\frac{\partial Vout}{\partial t} = -\frac{2\pi GBP}{1+G}{}^*Vout + \frac{(G.Vdiff + Vref).2\pi GBP}{1+G} \tag{5}$$

Considering a current step through the shunt, and thus a differential input voltage step from 0 V to Vdiff, we can solve this differential equation as described in Equation 6.

$$Vout(t) = G.Vdiff.\left(1 - e^{\frac{-2\pi GBP.t}{1+G}}\right) + Vref \tag{6}$$

Equation 6 describes the transfer function of a differential amplifier. And when we look deeper into, it is difficult to find any slew rate impact.

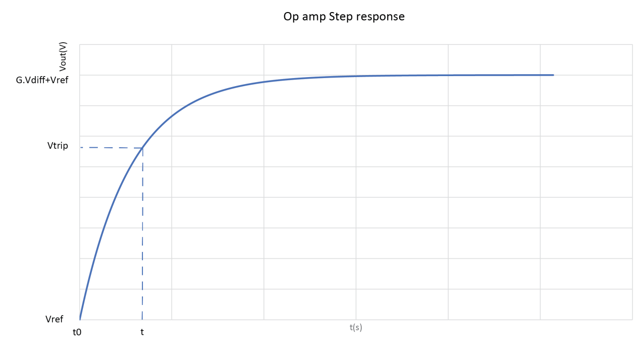
From Equation 6, we can deduce the timing needed (Equation 7) to reach a defined output voltage. In our case the output voltage we are interested in is a voltage Vtrip, which can trigger the comparator, as we want to consider an overcurrent happening in the shunt. So, the timing we are interested in is from Vout (t0) to Vtrip voltage. In this case, Vout (t0) = Vref

$$t = -\frac{1+G}{2\pi GBP} * ln\left(1 - \frac{Vtrip - Vref}{G.Vdiff}\right) \tag{7}$$

The first-order Equation 7 demonstrates that the response time of a differential amplifier is mainly dependent on the Gain Bandwidth Product (GBP), rather than slew rate. So the higher the GBP, the smaller the response time.

Figure 5 depicts the response time of an amplifier based on Equation 6.

Figure 5. Output op amp response to a step input



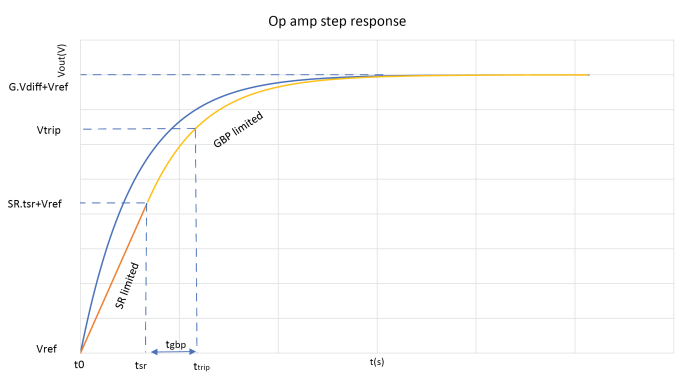
Slew rate limitation

But what about the slew rate? Does it really have no impact? The output of a differential voltage can indeed be limited by slew rate. Equation 6 is true only if the slew rate of the amplifier is faster than the rising slope of Figure 5. So, what happens in the case of slew rate limitation? In other words, what happens when the rising slope from t0 of Equation 6 is rising faster than the slew rate, as expressed by Equation 8?

$$\frac{\partial Vout(0)}{\partial t} = G.Vdiff.\frac{2\pi GBP}{1+G} > Slew\ Rate \tag{8}$$

As equations are sometimes difficult to understand, we can also illustrate the behavior with a graphic. The curve of Figure 6 shows the response time between two amplifiers.

Figure 6. Output op amp response to a step input with slew rate limitation



The blue curve represents the response time to a voltage input step with a high slew rate, and in this case only the GBP of the amplifier will limit the response time. The orange/yellow curve represents the response time of the same amplifier but with a lower slew rate. The amplifier first drives the output in slew rate mode until a t_{sr} time (orange curve). Then it starts following a similar shape to the one determined by Equation 6 and reaches the V_{trip} voltage after a time tgbp. At V_{trip} voltage (yellow curve), the comparator triggers for OCP. The blue curve (GBP limited) reaches V_{trip} point faster than the yellow curve (slew rate limited). So, in the case of on OCP application, it is important to choose an amplifier with firstly a high Gain Product Bandwidth (GBP) and then with a sufficient slew rate, to improve the overall response time.

In the case of slew rate limitation, it is also interesting to know and calculate what can be the theoretical response time of the amplifier to an input step voltage.

So, referring to the orange/yellow curve in Figure 6, the timing t_{trip} to reach an op amp output voltage V_{trip} can be defined by Equation 9.

$$t_{trip} = t_{sr} + t_{gbp} \tag{9}$$

t_{sr}: time duration when the op amp output is in slew rate mode.

t_{gbp}: the time from which slew rate mode is finished until the voltage of V_{trip} is reached.

Let us first express the t_{sr} time. The amplifier output is only limited by the slew rate,

V_{out} = SR.t + Vref until t_{sr}. Beyond this point, the amplifier output is mainly limited by the GBP as described by Equation 6.

So, from t_{sr} time the equation can be written as Equation 10. It is basically Equation 6 with an offset shift of SR * t_{sr}, as we consider that after this point, we are no longer is slew rate mode.

For $t \geq t_{sr}$

$$Vout(t) = (G.Vdiff - SR.t_{sr})*\left(1 - e^{\frac{-2\pi GBP.(t - t_{sr})}{1+G}}\right) + SR.t_{sr} + \mathrm{Vref} \tag{10}$$

t_{sr} time is the last point where the slope, or the derivative of Equation 10, is equal to the slew rate, and we can write Equation 11.

$$\frac{\partial Vout(t_{sr})}{\partial t} = \frac{2\pi GBP}{1+G}*(GVdiff - SR.t_{sr}) = \mathrm{SR} \tag{11}$$

t_{sr} can then be deduced as described by Equation 12.

$$t_{sr} = \frac{GVdiff}{SR} - \frac{1+G}{2\pi GBP} \tag{12}$$

t_{sr} cannot be negative, otherwise it means that there is no slew rate limitation, and we can consider t_{sr} = 0;

Considering Equation 10, the trip point is reached when Vout(t_{trip}) = V_{trip}, so it leads to Equation 13.

$$t_{trip} = t_{sr} - \frac{1+G}{2\pi GBP}*ln\left(1 - \frac{Vtrip - Vref - SR.t_{sr}}{G.Vdiff - SR.t_{sr}}\right) \tag{13}$$

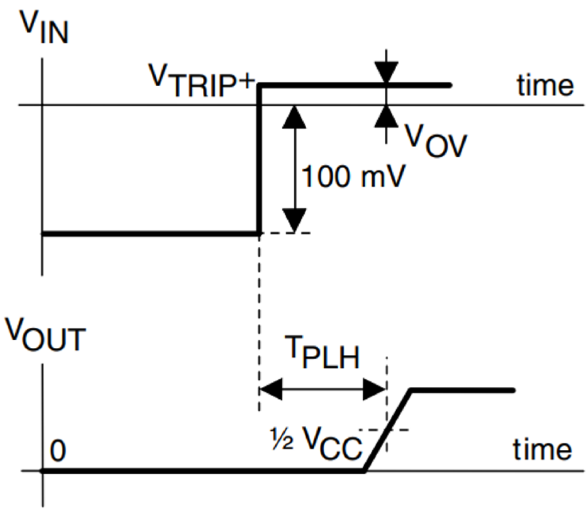
Comparator response time

Up to now, only the response time of an amplifier to an input voltage has been discussed. To complete the OCP function, the op amp output voltage must be compared to a reference voltage of a comparator. And this comparator has its own response time, which must be considered in the entire response time to a current event.

Propagation delay T_{PD} is one of the key parameters for many applications because it limits the maximum input frequency which can be processed. Voltage comparison of analog signals requires a minimum amount of time. T_{PD} is defined as the time difference between the moment the input signal crosses the reference voltage and the moment the output state changes (usually when the output signal crosses 50% of V_{CC}, if nothing is specified).

A graphical interpretation is shown in Figure 7. For T_{PLH} an input square signal from Vtrip+ -100 mV to Vtrip+ +V_{OV} is applied on the non-inverting input. V_{OV} is named overdrive.

Figure 7. T_{PLH} diagram



For more information about the comparator see Application Note AN4071: Introduction to comparators, their parameters and basic applications.

In addition to the response time of the amplifier, one should consider the response time of the comparator, which is mainly linked to the T_{PLH} and the overdrive.

The total response time after an overcurrent from 0 to Vdiff at shunt level can then be written as Equations 14 and 15.

$$t_{OCP} = t_{trip} + t_{PLH} \tag{14}$$

$$t_{OCP} = t_{sr} - \frac{1+G}{2\pi GBP} * ln\left(1 - \frac{Vtrip - Vref - SR.t_{sr}}{G.Vdiff - SR.t_{sr}}\right) + t_{PLH} \tag{15}$$

Note that if the amplifier response time is not limited by slew rate, $t_{sr} = 0$ (equation 7). If we want to consider an overdrive, we must replace Vtrip by Vtrip + Vov.

3 Precision

Detecting an overcurrent as fast as possible is a key element in a protection system. But detecting an overcurrent with accuracy is also very important to ensure the most efficient protection. To avoid affecting the current reading through the shunt, the current sensing amplifier and comparator must be as precise as possible. Unfortunately, as nothing is perfect, all active or even passive electronics add their own identity, or error, to a system. Several sources of error must be considered, and most of the time, most of them tend to change with temperature.

One of the principal sources of error is the input voltage offset generally called Vio in datasheets. The Vio results from an inherent mismatch of the input transistors. By carefully choosing a precision amplifier or precision current sensing, its impact can be limited.

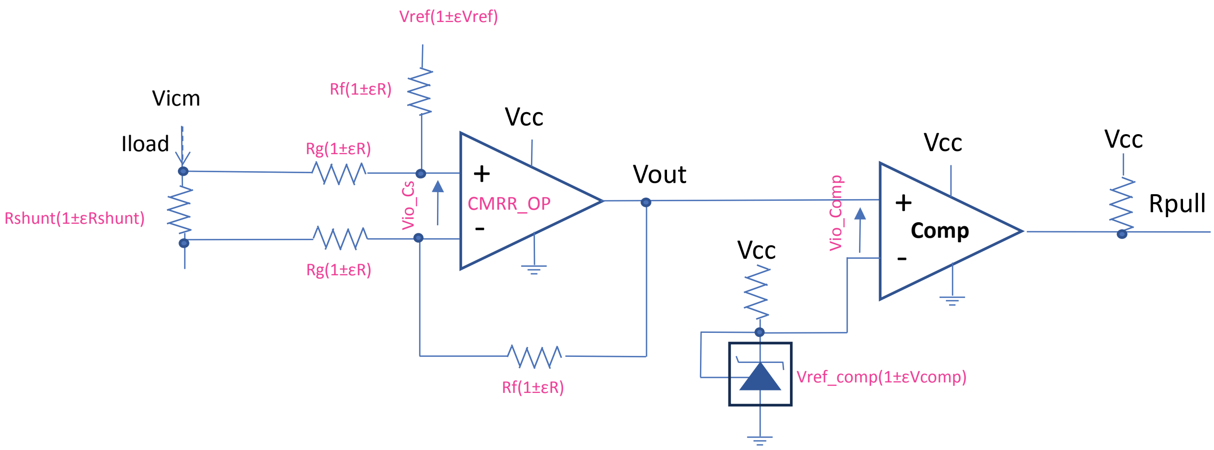
The common-mode rejection ratio ($CMRR_{Op\ amp}$) must also be considered in the error budget computation, as it expresses a variation of the Vio depending on the common-mode voltage applied.

And a $CMRR_{RES}$ which is linked to the mismatch of external resistors used in the case of op amp configuration as described in Figure 1. The mismatch of resistors also leads to a gain error.

Any other voltage source applied can add potential error to the whole measurement. For example, the Vref voltage fixes the output common-mode voltage of the amplifier, for a bidirectional current sensing operation. Similarly, the Vref_comp voltage fixes the Vtrip voltage of the comparator to trigger an overcurrent.

Figure 8 shows the sources of error to consider.

Figure 8. DC sources of error which affect the accuracy of the OCP



All these errors can be summed in a generic Equation 16. It gives a first-order, worst-case error at the comparator level, which must be adapted depending on the schematics used:

$$Vout\_error = (\pm I{*}Rshunt\ {*}(\varepsilon Rshunt \pm \varepsilon Gain) \pm \varepsilon CMRR_{RES} \pm Vio\_CS{*}\left(\frac{1+G}{G}\right) \pm \varepsilon CMRR_{Op\ amp}{*}\left(\frac{1+G}{G}\right)){*}G \pm Vref{*}\varepsilon Vref \pm Vio\_comp \pm Vref\_comp{*}\varepsilon Vref\_comp \quad (16)$$

Where:

- **I** is the current flowing into the shunt expressed in A.
- **Rshunt** is the shunt resistance value expressed in Ω.
- **εRshunt**: precision of the shunt resistance expressed in %.
- **εGain**: the gain error is linked to the accuracy of the resistance and can be expressed as 2 **εR** (the accuracy of the resistances).
- **$\varepsilon CMRR_{RES}$**: In the case of an op amp and external resistance used for the gain, it is important to note that a mismatch between resistors impacts the output voltage. This impact is measured by the CMRR_Res. Vicm can only be partially rejected if the resistors are not perfectly matched. The following CMRR_Res calculations highlight the impact of the resistor inaccuracies and in this case the following error must be used $\frac{\pm(Vicm - Vref)*4\varepsilon R}{1+\frac{Rf}{Rg}}$ with εR, the resistances accuracy.
- **Vio_Cs** is the input offset voltage of the first stage amplifier, either a current sensing, or an op amp expressed in V.
- **$\varepsilon CMRR_{Opamp}$**: express the Vio variation depending on the common-mode voltage of the application. In most datasheets, the Vio is given for a fixed common-mode voltage Vio_datasheet. This error must be considered if the common voltage of the application (Vicm) is different from the one defined in the datasheet and it is defined as $\frac{vicm - Vio\_datasheet}{CMRRop}$.
- **G** is the gain of the schematic. An internal gain for a current sensing amplifier, or a gain fixed by external resistance. In the case of an op amp, gain is given by $G = \frac{Rf}{Rg}$.
- **εVref** is the precision of the reference voltage used to fix the output common-mode voltage of the first stage (current sensing) expressed in %.
- **Vio_Comp** is the input offset voltage of the second stage, the comparator.
- **εVref_comp** is the precision of the reference voltage to fix the threshold voltage of the comparator expressed in %.

For deeper impact analysis of different sources of error, see Application Note AN4586: Signal conditioning, differential to single-ended amplification and AN5849: Expected output error for a current sensing amplifier.

4 Application example

Solution 1: Op amp and comparator

Although this kind of schematic is more dedicated to low-side current, Figure 9 describes an overcurrent function for high-side current sensing. It is realized with a fast precision 36 V op amp TSB711A, and a comparator TS391A. The goal of this OCP function is to detect a 40 mA overcurrent on a 24 V battery.

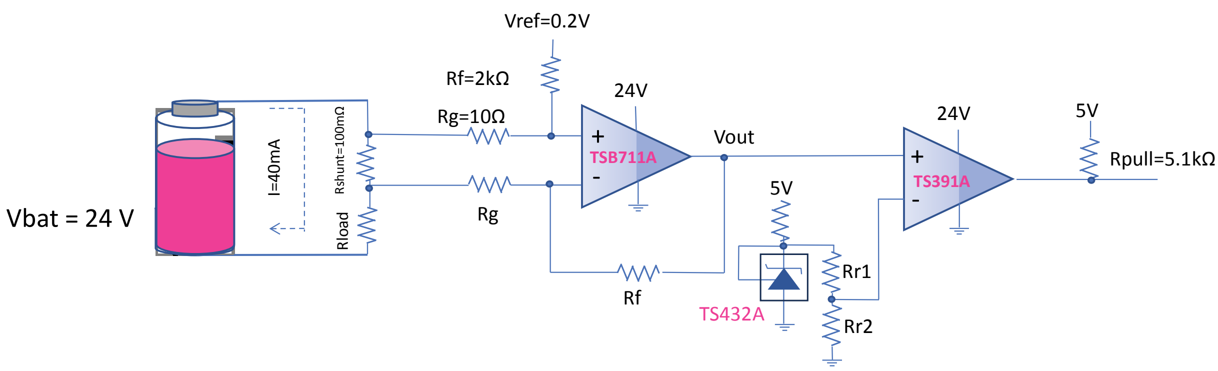
The gain of the first stage is set to 200 due to external resistances Rf and Rg. To optimize performance, the accuracy of all resistances (Rshunt included) is set at 0.1%.

To avoid any op amp output saturation when no current is flowing into the shunt resistance, a reference voltage Vref = 200 mV is applied on Rf resistance. This Vref voltage can fix the output common-mode voltage Vocm of the TSB711A.

To be sure to trigger the overcurrent event and to maximize speed of the comparator, an overdrive of 100 mV can be used (the larger the overdrive, the faster the response time). The 0.5% accuracy of the TS432A reference and 0.2% added due to external resistance Rr1 and Rr2 will not impact the triggering point of the comparator.

Note that due to the system response, the 100 mV overdrive would typically be reached asymptotically.

Figure 9. OCP function with a TSB711A op amp and a TS391A comparator



Note that, generally, the current sense information goes into the inverting pin of the comparator to benefit from a faster propagation delay as well as the possibility to perform OR gating at application level.

Vtrip point setup

The first step is to set the Vtrip voltage thanks to the TS432A reference voltage with accuracy of 0.5%, to fix the triggering of the overcurrent detection.

From Equation 2, considering a 100 mV overdrive, the theoretical Vtrip voltage should be set at

$$Vtrip = 40mA * 100m\Omega * \frac{2000}{10} + 0.2 - 0.1 = 0.9V$$

But the precision of the first stage must be considered to determine the error on the output voltage in the worst-case condition.

Equation 17 is based on Equation 16 and adapted to a differential amplifier with external gain.

In the TSB711A datasheet, the input offset max is defined for the whole common-mode voltage so the term $\frac{vicm - \text{Vio\_datasheet}}{CMRRop}$ can be neglected.

$$Vout\_error = \left[\pm I * Rshunt * \left(\epsilon Rshunt \pm 2\epsilon R \right) \pm \frac{(Vicm - Vref)*4\epsilon R}{1 + \frac{Rf}{Rg}} \pm Vio\_TSB711A \left(\frac{Rf + Rg}{Rf} \right) \right] \frac{Rf}{Rg} \pm Vref * \epsilon Vref \pm Vio\_TS391A \quad (17)$$

ε Rshunt = εR = εVref = 0.1%

Vio_TSB711A = 650 µV max at 25 °C

Vio_TS391A = 1.5 mV max at 25 °C

So, relatively to Vtrip, $Vout\_error = \pm 229mV$

In this case the maximum error that can be expected on the output is ±229 mV. It simply means that for a 40 mA current flowing into the shunt, the TSB711A output voltage can be in the range [0.77V:1.23V]. Note that this calculation approach is very pessimistic, as it combines all the maximum error on the same schematics. Nevertheless, to have an extremely safe design, let us consider that the minimum output op amp voltage can be 0.77 V instead of the expected 1 V. It represents an accuracy error of 23%.

The TS432A reference voltage gives a stable 1.24 V over temperature, independently of its power supply. To have a proper overdrive at comparator level, it should be set 100 mV below the minimum expected output voltage, so at 0.67 V. However, considering the accuracy of the voltage reference and the resistors (0.7%), we need to fix Vtrip to (1-0.7%) *0.67 = 665 mV. So, by selecting the resistance value from the E48 standard, the Vtrip voltage is set to Vtrip = 0.664 V. The divider Rr1 and Rr2 divider bridge can be set as follows:

Rr1 = 1.3 kΩ

Rr2 = 1.5 kΩ

So, we ensure a 100 mV minimum overdrive for 40 mA, but in typical condition the OCP would switch for $\frac{0.664 - 0.2}{200*100m} = 23mA.$

So, this case is not really realistic but only serves as a base for comparison with a current sensing solution.

Response time considering the worst case conditions

Following Equation 6, it is important to note that the output common-mode voltage Vocm, fixed thanks to Vref voltage, is also impacted by the imprecision of the op amp and the external components. It is necessary to take this into consideration, as the response time of the amplifier is directly linked to the output voltage step. The Vocm error can be expressed as Equation 18 when no current is flowing into the shunt:

$$Vocm\_error = \pm Vio\_TSB711A * \left(1 + \frac{Rf}{Rg} \right) \pm Vref * \epsilon Vref + \frac{(Vicm - Vref)*4\epsilon R}{1 + \frac{Rf}{Rg}} * \frac{Rf}{Rg} \pm Vio\_Comp \quad (18)$$

Vocm_error = ±227 mV

For I = 0 A, the Vocm voltage can be in the range of [-27 mV: 427 mV]. The worst case to consider is when Vocm = -27 mV instead of 200 mV, as it represents the worst-case scenario to reach the trip voltage.

Vtrip and Vocm voltage being determined, we can then calculate the response time of this OCP schematics. The AC performances of the TSB711A are:

- GBP = 4 MHz min
- Slew rate = 2 MV/s min

t_{sr} time duration given by Equation 12, for a 0 to 40 mA step, is negative, meaning the TSB711A is not limited by its slew rate but only by its GBP, so the total response time to an overcurrent t_{OCP} can be written as Equation 15 by considering t_{sr} = 0 and considering the required overdrive.

$$t_{OCP} = t_{sr} - \frac{1 + G}{2\pi GBP} * ln\left(1 - \frac{Vtrip + Vovd - Vocm - SR \, . \, t_{sr}}{G \, . \, Vdiff - SR \, . \, t_{sr}} \right) + t_{PLH}$$

To have the worst-case scenario, we must consider Vtrip max and Vocm min.

So, with Vtrip = 0.664 * (1+0.7%) = 0.669 V, Vocm = -27 mV, Vdiff = 4 mV, G = 200 and t_{PLH} = 500 ns Vovd = 50 mV

$$t_{OCP} = \; = 22\mu s$$

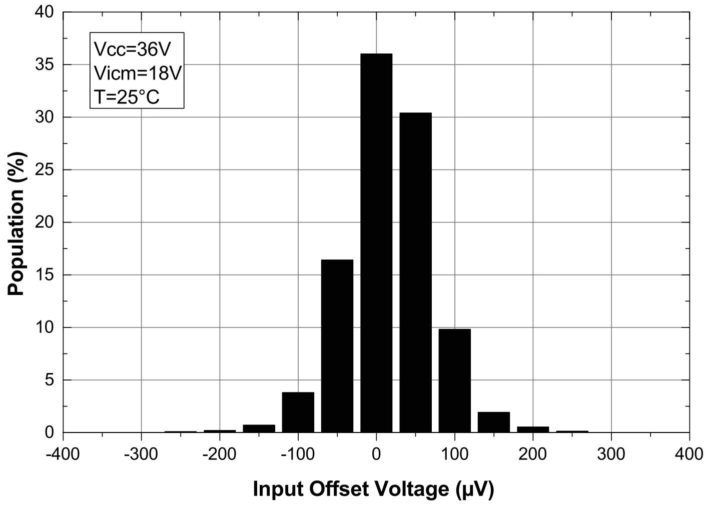
The worst-case timing (considering the sum of all the maximum values) necessary to detect a current step of 40 mA into a shunt of 100 mΩ is 22 µs. Again, this is a very pessimistic approach as the chance to get all the maximum parameter values is extremely low, if not nil.

Note that we have designed the application to ensure an overdrive of 100 mV in the worst-case scenario. But as it may be reached over a long period of time, we only consider going 50 mV above the trip point for the above calculation with a 500 ns propagation delay.

Response time for a typical application

As we can see in this application example, one of the main limitations is the input voltage offset Vio in terms of precision. Let us look at the distribution in Figure 10 of the TSB711A Vio, where we can see that the majority of the population will be in the range of ±100 µV.

Figure 10. Input offset voltage distribution at V_{CC} = 36 V TSB711A



The voltage trip point of the comparator is kept at 0.664 V to be sure to trigger an overcurrent whatever the TSB711A. But for the calculation of a typical overcurrent detection, we consider typical parameter as described below (for Vocm error we only consider the error related to the input-offset voltage).

GBP = 6 MHz typ, SR = 3 V/µs typ, Vio = 100 µV typ, Vocm = 0.18 V, Vdiff = 4 mV, G = 200 and t_{PLH} = 500 ns

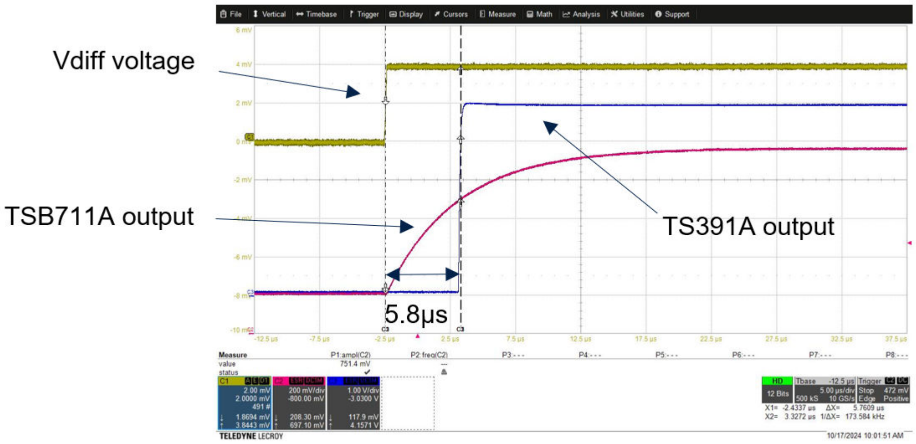
In this typical case and considering Equation 15, the expected response time without considering errors besides Vio is

$$t_{OCP} = \; = 6.4\mu s$$

not considering the overdrive of the comparator.

We can now realize a test bench to verify it. Figure 11 is a scope probe of the schematics described in Figure 9 with a Vtrip = 0.664 V.

Figure 11. **Typical response time to an overcurrent with a TSB711A (gain = 200) and TS391A**



The bench evaluation of such a solution shows a total response time of 5.8 µs, to detect a 40 mA over current.

Nevertheless, there is an error of roughly 42% in the precision of the OCP. To improve this, and to optimize the accuracy of overcurrent detection, several solutions can be considered:

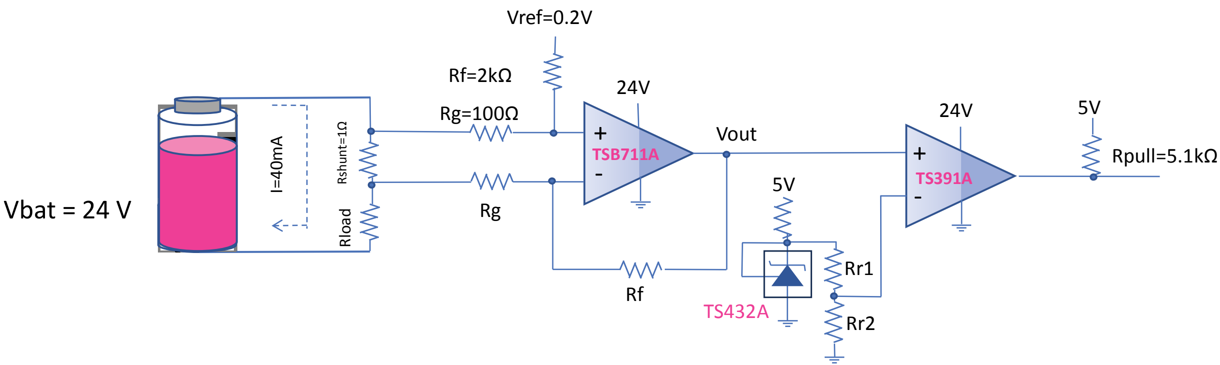
- Choose an amplifier with a better input offset.
- Minimize the overdrive for comparator triggering, which will add some response time.
- It is also possible to calibrate the first stage current sensing with zero current.

Impact of input differential voltage on response time

As seen with Equation 17, the response time of such overcurrent protection is principally linked to the GBP and the SR of the chosen op amp. But it also depends on the application gain, so the differential input voltage has a strong impact. Indeed, if it is ten times higher, the input-offset voltage has a ten times lower impact.

Figure 12 depicts the same application as the one described by Figure 9, but this time using a gain of 20 and a shunt of 1 Ω, giving a differential voltage of 40 mV for a 40 mA step current.

Figure 12. **OCP function with an op amp and a comparator and gain of 20**



Doing the same as in the previous case, Vtrip can be set to 786 mV.

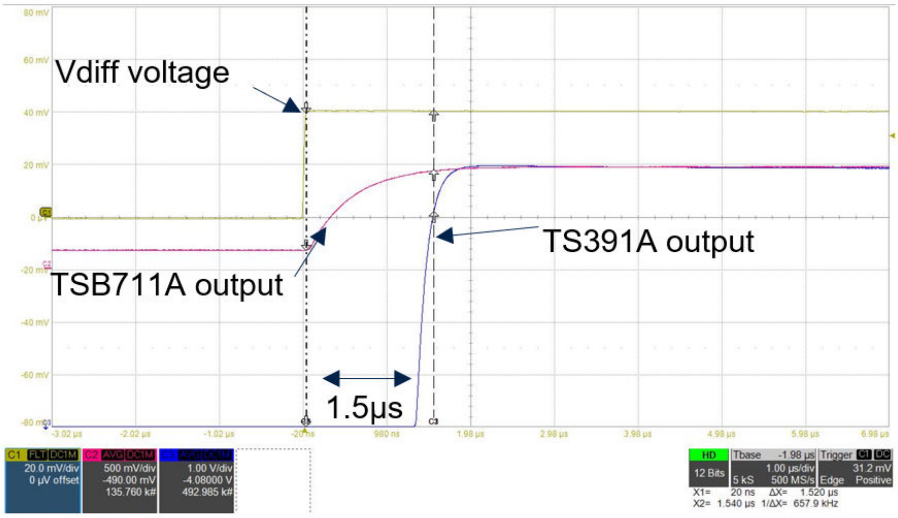
In this typical case and considering Equation 17, the expected response time is:

$$t_{OCP} = \; = 1.4\mu s$$

With GBP = 6 MHz typ, SR = 3 V/µs typ, Vio = 100 µV typ, Vocm = 0.198 V, Vdiff = 40 mV, G = 20 and t_{PLH} = 500 ns

Let us now realize a test bench to verify it. Figure 13 is a screenshot based on the schematics described in Figure 12 with a Vtrip = 0.786 V.

Figure 13. **Typical response time to an overcurrent with a TSB711A (G = 20) and TS391A**

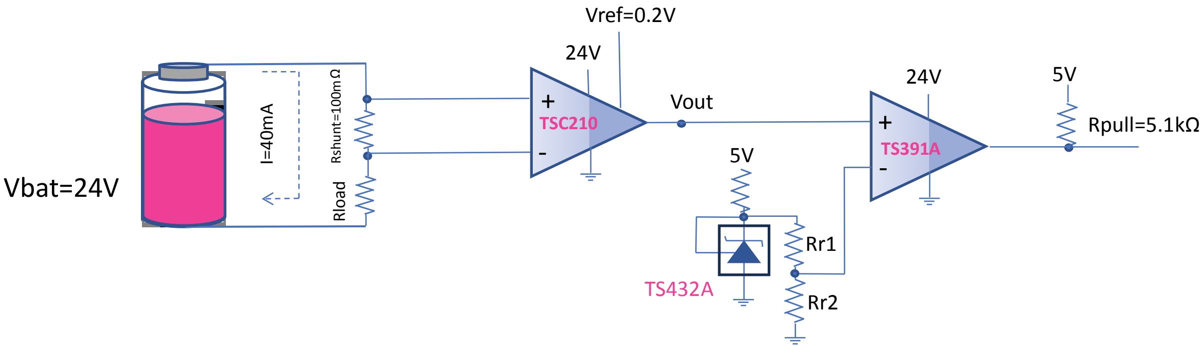


In this use case, the response time to a 40 mA step current has been divided by 5 on bench, just by changing the shunt and the gain of the schematics. So, reducing the gain is a solution to improve the reaction time to an overcurrent event (at the cost of a higher power dissipation in the shunt).

Solution 2 High-side current sensing and comparator

Let us try to choose a first stage amplifier with a better input-offset voltage than the TSB711A, and to simplify the architecture, by using a high-side bidirectional precision current sensing with an integrated gain of 200, the TSC210. The TSC210 is a zero drift current sense amplifier that can sense current via a shunt resistor over a wide range of common-mode voltages from -0.3 V to +26 V, whatever the supply voltage is. Figure 14 describes the schematic of such an OCP. It is realized with a high-side current sensing TSC210 (gain 200) and a fast comparator TS391A. As before, the goal of this OCP function is to detect a 40 mA overcurrent on a 24 V battery.

Figure 14. **OCP function with a high side current sensing and a comparator**



Vtrip Voltage setup

As in the previous example, the first step is to estimate the maximum possible error in the output of the TSC210 current sensing. It is necessary to define the comparator reference voltage Vtrip.

Equation 19 is based on Equation 16 and adapted to a current sensing with integrated gain.

$$Vout\_error = (\pm I*Rshunt*(\varepsilon Rshunt \pm \varepsilon Gain) \pm Vio\_CS \pm \varepsilon CMRR_{Opamp})*G \pm Vref*\varepsilon Vref \pm Vio\_comp \quad (19)$$

εgain = 1%

εRshunt = εR = εVref = 0.1%

Vio_TSC210 = 35 µV max at 25 °C

Vio_TS391A = 1.5 mV max at 25 °C

$$\varepsilon \mathrm{CMRR}_{\mathrm{Op\ amp}} = \frac{24V - 12V}{10^{105dB/20}}$$

$$Vout\_error = \pm 31\, mV$$

In this case, the maximum error is largely improved compared to the TSB711A. It simply means that for a 40 mA current flowing into the shunt, the TSC210 output voltage can be in the range [0.97V:1.03V].

Note that this calculation approach is very pessimistic, as it combines all the maximum errors on the same schematics. Nevertheless, to have an extremely safe design, let us consider that the minimum output op amp voltage can be 0.97 V instead of the expected 1 V .

To be sure to trigger the overcurrent event and to maximize speed of the comparator, the Vref_comp voltage can be set at 0.97 V - 100 mV for a large overdrive, so set at 0.87 V. Thanks to the 0.5% accuracy of the TS432A reference and 0.2% due to external resistance Rr1 and Rr2, the trip voltage error is up to 6 mV.

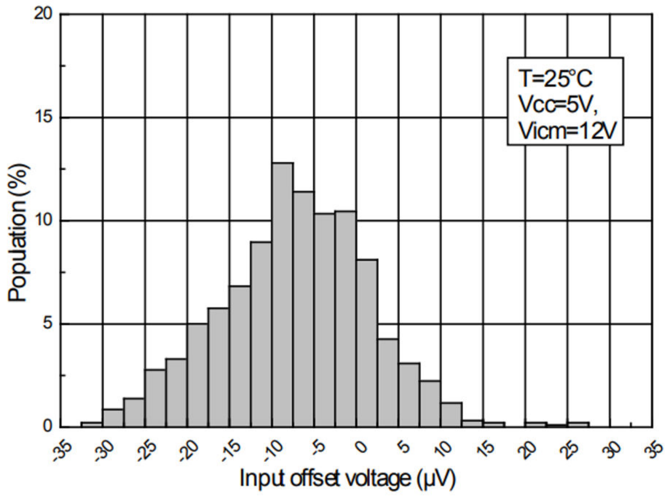
The TS432A reference voltage gives a stable 1.24 V over temperature, independently of its power supply. To select the resistance value from the E48 standard the Vtrip voltage is set to Vtrip = 0.86 V. The divider Rr1 and Rr2 divider bridge can be set as follows:

Rr1 = 442 Ω

Rr2 = 1000 Ω

In the distribution in Figure 15 of the TSC210 input offset voltage Vio, we can see that the majority of the population will be in the range of ±10 µV.

Figure 15. Input offset voltage distribution at V_{CC} = 5 V TSC210



This offset plus its variation due to CMRR only contributes to 15.5 mV of Vout error.

Output common-mode voltage error

Following Equation 6, it is important to note that the output common-mode voltage Vocm, fixed thanks to Vref voltage, is also impacted by the imprecision of the op amp and the external components. It is necessary to take this into consideration as the response time of the amplifier is directly linked to the output voltage step. The Vocm error can be expressed as Equation 20 when no current is flowing into the shunt:

$$Vocm\_error = \pm Vio\_TC210 * Gain \pm Vref * \varepsilon Vref \pm \varepsilon CMRR_{Op\ amp} * G \pm Vio\_comp \quad (20)$$

Vocm_error = ± 22 mV

The Vocm voltage can typically be in the range of [178 mV: 222 mV]. Vocm =178 mV is considered instead of 200 mV.

Response time of the system to an overcurrent

The AC performances of the TSC210 are:

GBP = 5 MHz

Slew rate = 0.2 MV/s min

t_{sr}: time duration given by Equation 12 is negative, meaning the TSC210 is not limited by its slew rate but only by its GBP, so Equation 17 can be used by considering t_{sr} = 0

$$t_{OCP} = t_{sr} - \frac{1+G}{2\pi GBP} * ln\left(1 - \frac{Vtrip + Vovd - Vocm - SR.t_{sr}}{G.Vdiff - SR.t_{sr}}\right) + t_{PLH}$$

With Vtrip = 0.86 V, Vocm = 0.178 V, Vdiff = 4 mV, G = 200 and t_{PLH} = 500 ns, Vovd = 50 mV

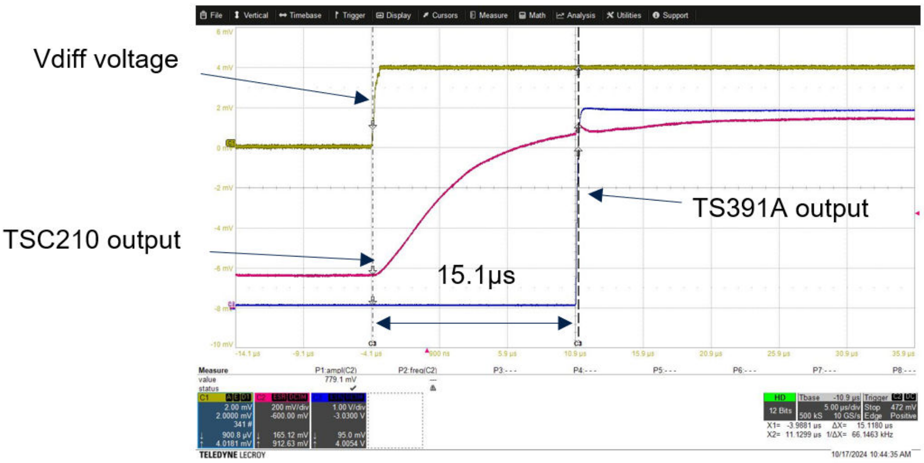
In this typical case, the expected response time to an over current of 40 mA is:

$$t_{OCP} = 16.3 \mu s$$

Test bench measurement of OCP realized with a TSC210 current sensing and TS391A comparator

Figure 16 is a scope probe of the schematics described in Figure 14 with Vtrip = 0.86 V.

Figure 16. **Typical response time to an overcurrent with a TSC210 and TS391A**



The bench evaluation of such a solution shows a total response time of 15.1 µs, to detect a 40 mA overcurrent. As before, the typical application result fits with the equation.

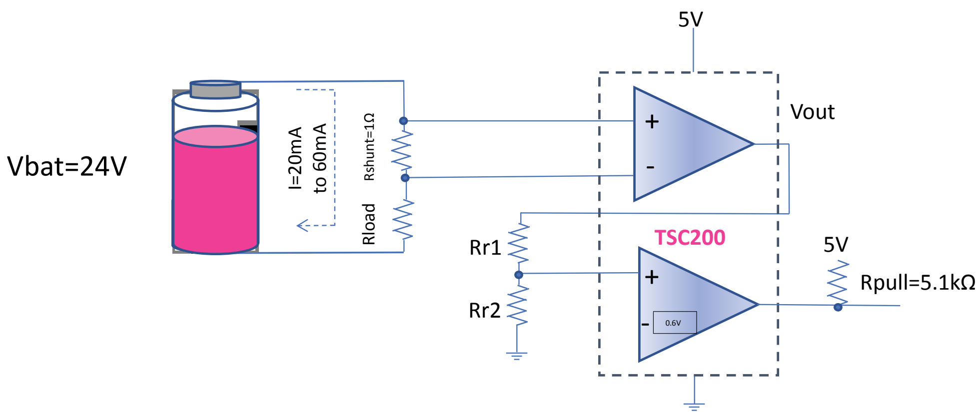
By using the TSC210, a typical error of 17.5% can be expected in the accuracy of overcurrent detection. Note that 12.5% of the error is related to the overdrive of 100 mV. The TSC210 offers other advantages compared to solution 1, as it is space- and BOM-saving thanks to the internal gain. Nevertheless, solution 2 is a bit slower in terms of the response time to an overcurrent event compared to solution 1.

Solution 3 All in one OCP function

In this last example, we consider a high-side current sensing with integrated comparator in one package to maximize PCB area and drastically reduce external components. The TSC200 can be considered for overcurrent protection. The internal topology used in the TSC200 is different from a classic instrumentation amplifier approach. So, the timing defined by Equation 17 cannot be applied for such an architecture.

This topology determines a different operation functional mode for the TSC200, and it is a function of two main variables: The differential input voltage Vsense (resulting from a current flowing into the shunt resistance) and the input common-mode voltage Vicm, relative to the supply voltage Vcc. To work properly and ensure the parameters in the datasheet, a minimum input voltage of 20 mV is required. The TSC200 offers a gain of 20. So, the approach is slightly different from the previous solutions 1 and 2. In this case, the 40 mA step current will be from 20 mA to 60 mA through a shunt of 1 Ω the OCP threshold is fixed at 60 mA. Figure 17 depicts the application schematic using the TSC200 for overcurrent protection.

Figure 17. OCP function with a high side current sensing and an integrated comparator



Vtrip voltage setting

As in the previous solution, let us first determine the comparator reference voltage Vtrip. It is important to estimate the maximum possible error in the output of the current sensing TSC200.

Equation 21 is based on Equation 16 and adapted to a current sensing with integrated gain without considering the offset of the comparator:

$$Vout\_error = (\pm I \cdot Rshunt \cdot (\varepsilon Rshunt \pm \varepsilon Gain) \pm Vio\_CS \pm \varepsilon CMRR_{CS}) \cdot G \tag{21}$$

εRshunt = 0.1%

εGain = 1%

Vio_TSC200_CS = 2.5 mV max at 25 °C

$$\varepsilon \mathrm{CMRR}_{CS} = \frac{24V - 12V}{10^{100dB/20}}$$

$$Vout\_error = \pm 66\, mV$$

In this case, the maximum error that can be expected in the output is ± 66 mV. It simply means that for 60 mA into the shunt, the TSC200_CS output voltage can be in the range [1.13V:1.27V]. Note that this calculation approach is very pessimistic, as it combines all the maximum errors on the same schematics. Nevertheless, to have an extremely safe design let us consider that the minimum output op amp voltage can be 1.13 V instead of the expected 1.2 V.

In the case of TSC200, the reference voltage for the triggering point is already integrated and is fixed at 0.6 V. So, considering a maximum offset of 20 mV at comparator level, and 0.1% resistors, we should have $\frac{1.13}{Rr1 + Rr2} Rr2 \geq 720mV \cdot \left(1 + 0.2\%\right)$.

To obtain the fastest response time of the comparator, an overdrive of 100 mV is considered. Following the E48 standard, the divider bridge on the output can then be set with:

Rr1 = 562 Ω

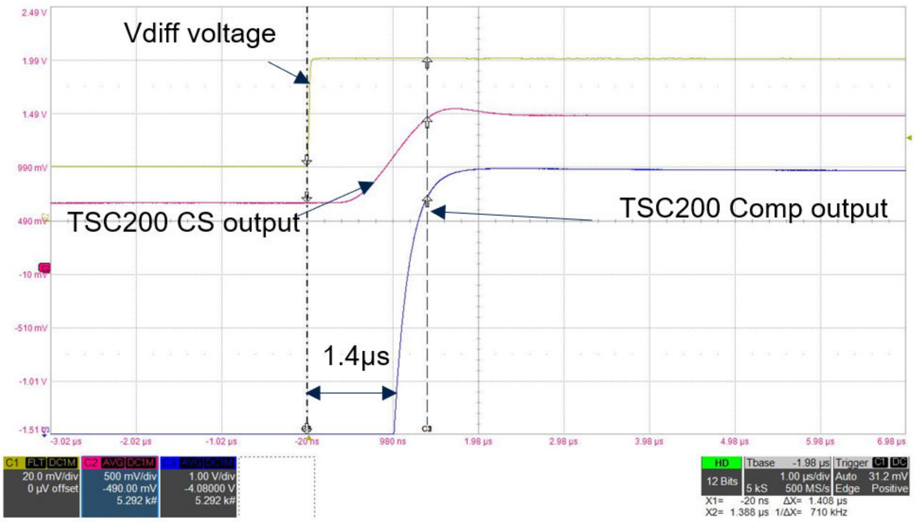
Rr2 = 1000 Ω

Rr1 and Rr2 are chosen with 0.1% precision, so it will not impact the triggering point of the comparator.

Test bench with a TSC200

Figure 15 represents a typical response time to a 40mA current step (from 20 mA to 60 mA).

Figure 18. **Typical response time to an overcurrent with a TSC200**



The response time is 1.4 µs for an input current step of 40 mA. And this response time is similar to solution 1 using an op amp TSB711A with a gain of 20 a comparator TS391A and external components. Nevertheless, the main advantage of this solution is the simplicity of implementation, space-saving area and cost.

5 Conclusion

Monitoring current is essential in many kinds of application. On top of that, being able to detect a potential issue such as an overcurrent event accurately and as fast as possible, helps improve the robustness of the application.

There are many overcurrent detection solutions and all of them have their own advantages depending on the application constraint.

In this application, three different architectures of overcurrent protection based on op amp or current sensing have been studied. Key parameters to choose a device to build an OCP function have been described, as well as equations and bench results.

Nevertheless, it can be difficult to compare them because the configurations are not the same. The following table summarizes the pros and cons of each of the solutions studied above.

Table 1. Pros and cons solution

	Pros	Cons
Solution 1: Op amp (TSB711A) + Comp (TS391A)	Flexibility: Gain can be set to fit the application Response Time Bidirectional mode	Larger BOM More suitable for low-side current sensing Accuracy of detection
Solution 2: Current sensing (TSC210) + Comp (TS391A)	Very accurate detection of overcurrent events Low-/high-side configuration Bidirectional mode Few external components	Slightly longer response time
Solution 3: Current sensing (TSC200)	Only one device so BOM and PCB area saving Response time Low-/high-side configuration	Monodirectional mode Less efficient at measuring 0 current

Revision history

Table 2. **Document revision history**

Date	Version	Changes
03-Apr-2025	1	Initial release.

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