Application Note

A Self-Calibratable Current Detection Solution Based on MSPM0

TEXAS INSTRUMENTS

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ABSTRACT

This application note is created to build a description about a self-calibratable, high precision and low-cost current detection solution based on MSPM0 internal OPAs. This document describes the calibration method, paired with MSPM0 internal OPA usage. At the end, it also shows the test performance across different temperature.

Project collateral discussed in this application note can be downloaded from the following URL: https://www.ti.com/lit/zip/slaaee6.

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1 Introduction

The features of this solution are shown below:

- It supports bi-directional current detection.
- It can self-calibrate the voltage offset and GAIN setting of the OPA.
- It can do dynamic GAIN setting to reach high accuracy across low to high current.
- The detection error at room temperature can reach to ±0.25% at ±2A load.
- The current consumption can reach 8 μA with 1Hz sample rate.

As this solution is mostly developed for MSPM0 Gauge solution, the following section describes how to detect and test the discharge and charge current of a battery. For other current detection use cases, you can also take this note as a reference.

The solution is combined of three parts. All the software and hardware are reused by MSPM0 Gauge solution. You can find all of them at MSPM0 Current Detection Development package. You can also find the MCU code under the SDK (mspm0_sdk_xxx\examples\nortos\LP_MSPM0L1306\battery_gauge).

- It is the hardware board that is used to evaluate the current detection solution with MSPM0L130x integrated.

Figure 1-1. MSPM0 Current Detection Hardware Board

- It is the software project based on MSPM0L130x.

**Figure 1-2. MSPM0 Current Detection Software Project**

- It is the GUI. It is written by python, which is mostly used to read the test data from the board.
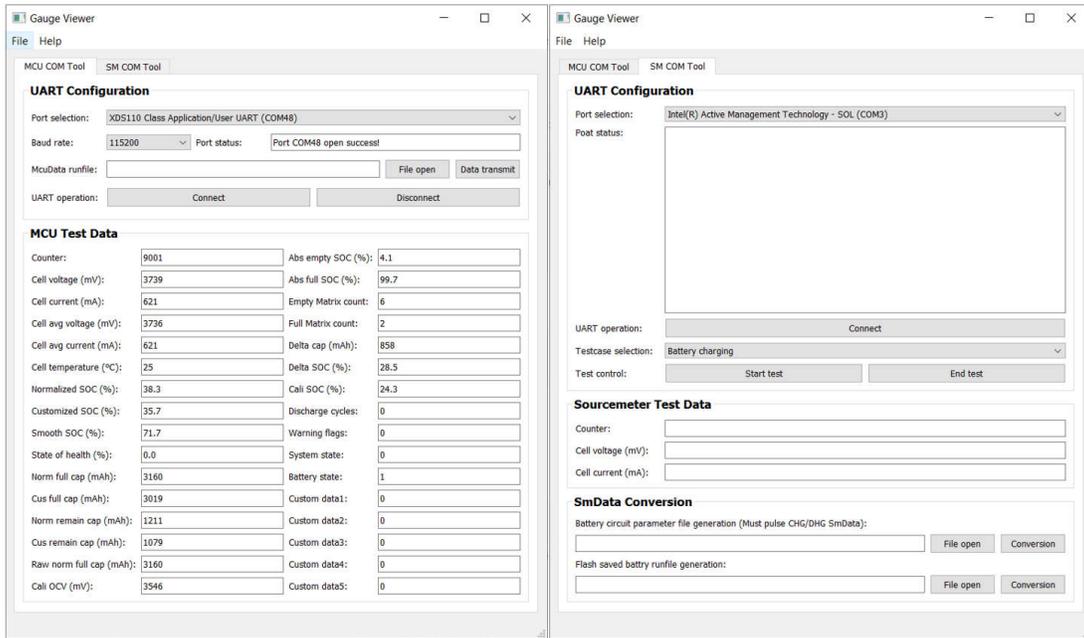
Figure 1-3. MSPM0 Gauge GUI Project

2 Hardware Introduction

Figure 2-1 is a hardware high-level block diagram, showing all the used pins by this demo. The solution demonstrates testing the current at ADC channel 13, the temperature at ADC channel5 and voltage at ADC channel1.
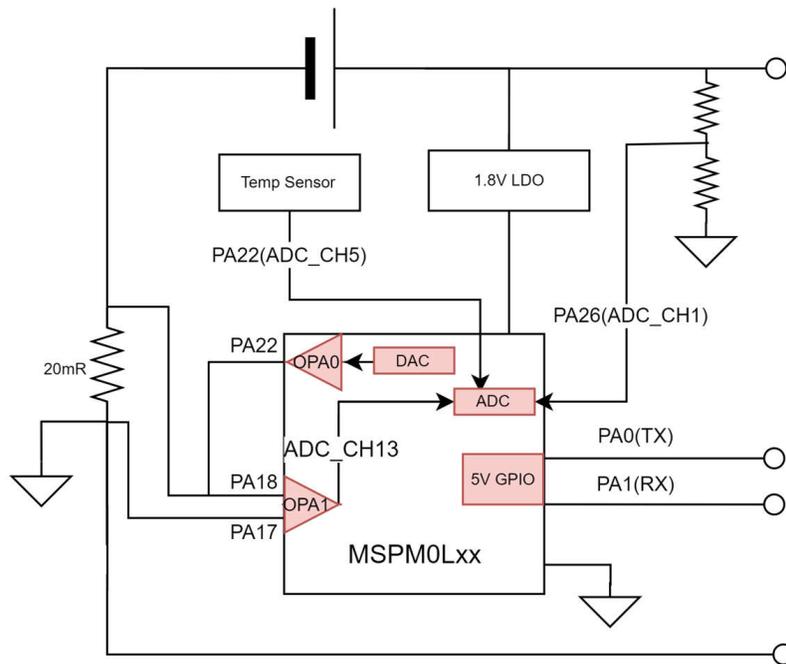
Figure 2-1. MSPM0 Gauge Board Block Diagram

This design can use these MSPM0 features:

- High precision 12-bit ADC for temperature, current and voltage detection.
- Integrated high precision chopper OPAs for current detection.
- 5 V tolerant open-drain I/O with UART or I2C function to communicate with masters under different power rails.
- Lower to 1.62 V working voltage to support single battery full voltage range.
- Lower to 1.1 µA STANDBY current with SRAM retention for battery application,

Here, a quick introduction for the hardware board is provided and how to use it. For the battery, you can insert it into the default socket or connect it to the backup battery supply input. Debug and UART COM port is used to connect to PC, which can download the code or communicate with the GUI.

Note

Pay attention to the MCU power switch supply jumper. For downloading, connect VMCU to VEx, then the MCU are supplied with 3.3 V, which can ensure the voltage matching with the debugger. For evaluation, connect VMCU to VIn, then the MCU is supplied with 1.8 V LDO. It can ensure the best analog performance.
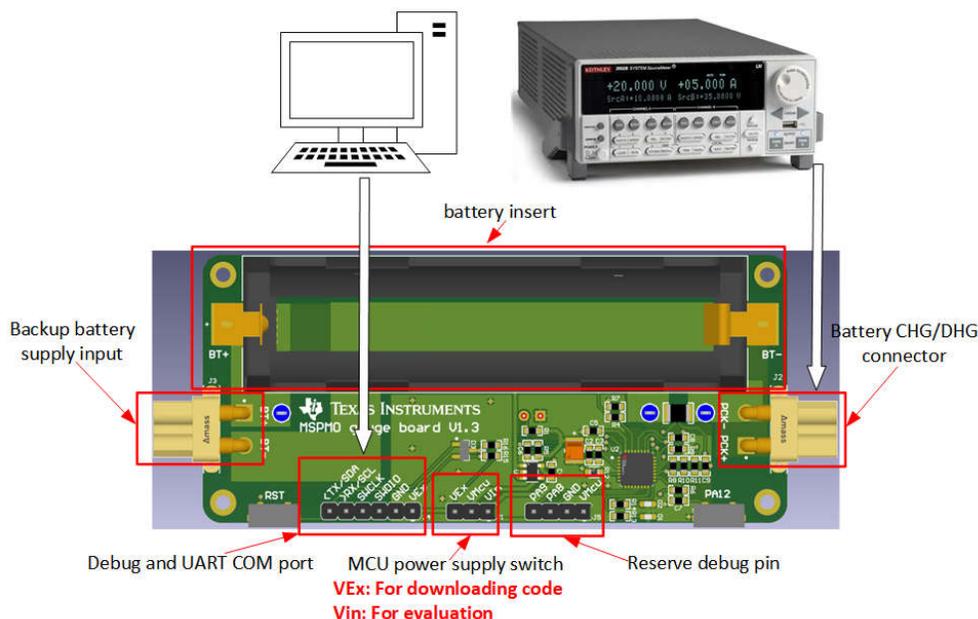
Figure 2-2. Gauge Board Instructions

3 Software Introduction

The software project is shown in Figure 3-1. The project and files related to the current detection has four parts. For other files, they are same for all the MSPM0 projects.
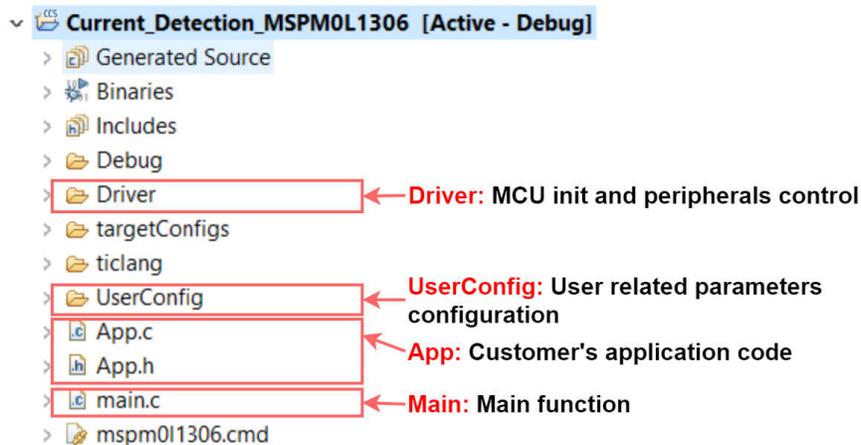
Figure 3-1. Software Project View

For UserConfig part, it is included in Section 6.2. For the driver part, it includes all the MPSM0 peripheral initialization and low-level function. Section 5 provides a description on how the current detection work. For the App part, it includes the high-level software function calling, which is the place for customers to customize their own functions. For the Main part, it includes the highest system function code.

4 Gauge GUI Introduction

In this demo, a Gauge GUI will be used to record MCU data, run battery test case, and do data conversion. This GUI has two pages. First is MCU COM Tool, used to communicate with MSPM0 and record the MCU transmitted battery running data. Second is SM COM Tool, used to communicate with the source meter, run battery test case and record the test data sent from the source meter.

The used function for current detection is to receive the battery running data from MCU, shown in the function 1 of Figure 4-1. The data is saved automatically in excel with a name "time-McuData.xlsx", after the battery test case settled in SM COM Tool is finished or you stop the test by pressing the "Disconnect" button.
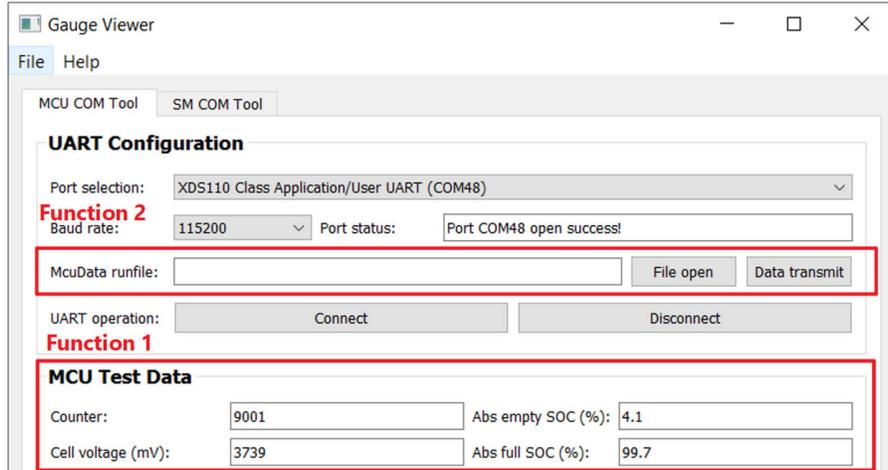
Figure 4-1. MCU COM Tool Function

For the SM COM Tool, the used function for current detection is function 1, shown in Figure 4-2. It is used to control the source meter to run the battery testcase to test the current detection performance and show the data measured by source meter on the GUI. After the test is finished, the record data sent from source meter is saved in excel with a name "time-SmData.xlsx". If you want to recreate this part, for the software, you need to at least install NI_VISA. For hardware, you need to buy a USB to rs232 wire and a keithley 2602A source meter.
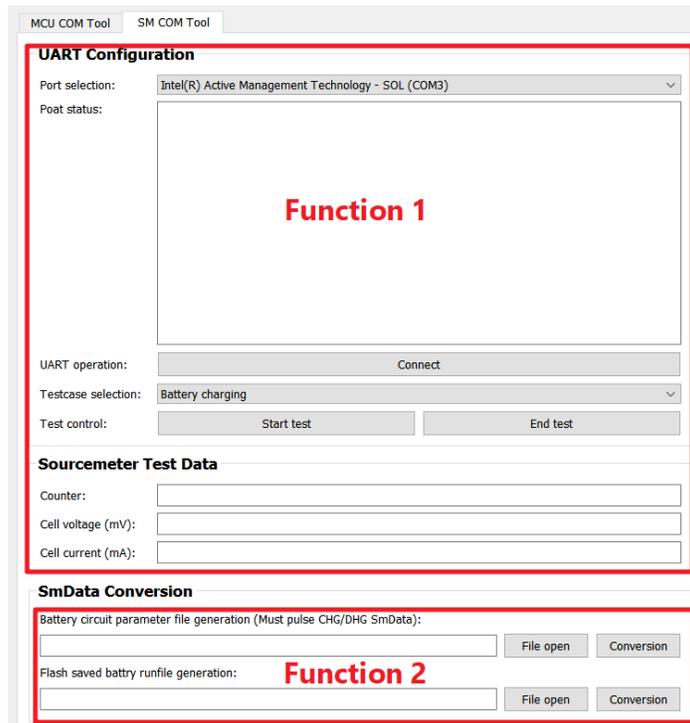
Figure 4-2. SM COM Tool Function

5 Current Detection and Calibration Method

Here is the current detection block diagram. We can see that this solution utilizes:

- Two high precision internal OPAs. One OPA used as a buffer for DAC. Another used to generate gain for the voltage across the shunter resistor.
- One internal DAC. Used to generate dynamic bias voltage for current detection paired with OPA0.
- Flexible internal gain, which is the key point to realize different detection range. If you increase the gain, then the detection range will be smaller, but its resolution will be increased as well.
- Internal SAR ADC. Used to calibrate and detect the voltage output from OPA0.
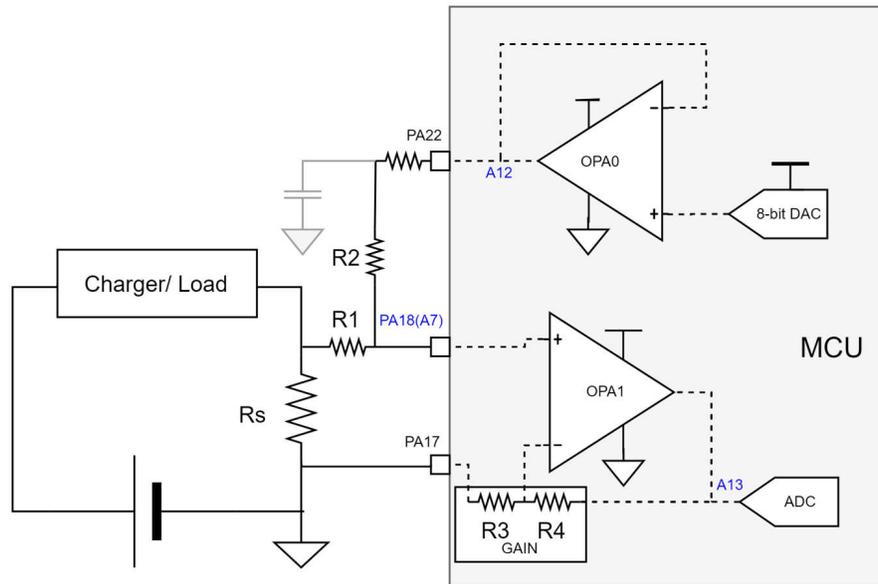
Figure 5-1. Current Detection Block Diagram

5.1 MSPM0 OPA Introduction

An introduction is provided about the MSPM0 OPA parameters to help you get a better understanding on how to get the best OPA performance and what is the OPA limitation.

5.1.1 OPA input and output limitation

OPA has some parameters related to operation limitations. If above the parameters, the linearity of OPA will get worst. That means the detection current is limited to a certain range.

The key parameters are common-mode voltage range and voltage output swing. Figure 5-2 shows the limitation value of MSPM0L130x, when set VDD as 1.8 V and RRI=1. For this device, to reach the typical open-loop voltage gain value on data sheet, output voltage should be controlled into a smaller voltage range (0.3 V ~ VDD-0.3 V) , compared with voltage output swing range. If above the range, the open-loop gain will be decreased. In this current detection demo, it doesn't take the open-loop gain into consideration which has smaller influence than other two parameters, especially under low gain setting.
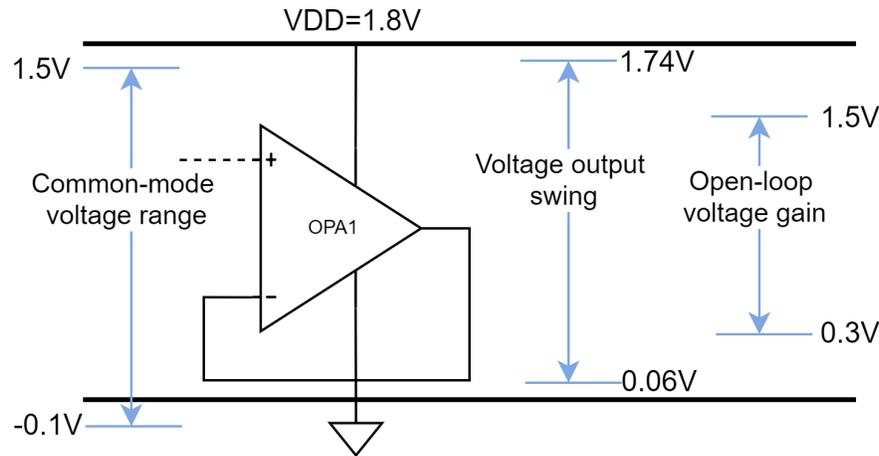
Figure 5-2. MSPM0 OPA Limitation

5.1.2 OPA Accuracy Influence

The OPA gain accuracy is influenced by three parameters:

- The first is the input bias current.
- The second is the input DC voltage offset.
- The third is the OPA noise.

As the MSPM0 OPA uses CMOS technology, the gate is physically isolated from the conduction channel to create an input that is truly high impedance. Its input bias current parameter is mostly combined of the leakage from ESD structures, protection diodes, and/or parasitic junctions. That means, input bias current can have positive or negative flow depending on the conditions. Unlike with bipolar amplifiers implementation, in this hardware design, that is why a matching resistor is not used at the OPA noninverting input. From calculation, the bias current influence to the OPA output is below than 1 mV, which also can be calibrated later.

For input voltage offset, chopper function is more helpful to reduce its influence. After enabling the chopper function, the max of input offset voltage goes from 3.5 mv to 0.5 mV, shown in the chapter 7.17 of MSPM0L1306 datasheet. The chopper function realizes this result by modulating the input signal and demodulating the output signal. For the input signal, it will keep the same after modulation and demodulation. For input voltage offset, it starts to take effect after signal modulation, and will do demodulation with signal together. As a result, chopper function changes the DC voltage input offset to an AC voltage output offset and do no effect to the input signal.

For the OPA on MSPM0, it has an ADC assisted chopping mode. After enabling ADC hardware oversampling, the chopper frequency is controlled by ADC sampling frequency (see Figure 5-3). As the hardware sampling times is an even number, the sampling times at the positive voltage output offset and negative voltage output offset will be the same. After the final averaging, the AC voltage offset can be removed through the digital filter, which means you don't need a hardware filter anymore! Because chopper technology cannot truly remove the voltage offset, when you calculate the output voltage range, you need to take the voltage offset into consideration, besides of the voltage output swing range limitation. As the voltage offset calibration shown in the current detection demo, the voltage offset parameter will mostly affect the detection current range.
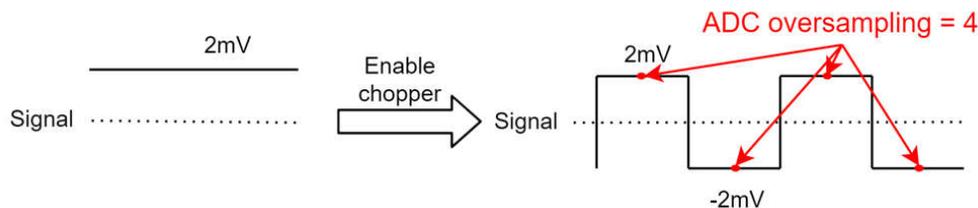

Figure 5-3. MSPM0 Chopper Function

The OPA noise is the key parameter that mostly affects the current detection performance. However, after enabling the chopper function, you can efficiently reduce part of the noise (1/f noise), because the chopping shifts the base-band signal to the chopping frequency beyond the input stage's 1/f region. Paired with software averaging filter, you can control the total AC noise to an acceptable level.

5.2 Current Detection Method

In this part, equations will be used to show how to do the current detection with dynamic GAIN setting by changing the resistor value of R3 and R4 in Figure 5-1.

Equation 1 shows the OPA1 output voltage. I is the current to be tested. Voffset includes the influence of voltage offset and bias current.

$$V_o = \left(\frac{V_{DAC} + OPA0 - IR_S}{R_1 + R_2} * R_1 + IR_S \right) \left(1 + \frac{R_4}{R_3} \right) + V_{offset} = \left(\frac{R_1}{R_1 + R_2} V_{DAC} + OPA0 + \frac{R_2}{R_1 + R_2} IR_S \right) \frac{R_3 + R_4}{R_3} + V_{offset} \tag{1}$$

The OPA1 output voltage is recorded as a reference for the current calculation, when I = 0. Here Vref is used to represent it in Equation 2.

$$V_{ref} = \frac{R_1}{R_1 + R_2} \frac{R_3 + R_4}{R_3} V_{DAC} + OPA0 + V_{offset} \tag{2}$$

When I != 0, the gap between OPA1 output voltage and Vref will be as shown in Equation 3.

$$V_o - V_{ref} = \frac{R_2}{R_1 + R_2} \frac{R_3 + R_4}{R_3} IR_S \tag{3}$$

After that subtraction, you can remove the Voffset influence and the current can be calculated using Equation 4.

$$I = \frac{V_o - V_{ref}}{R_S} \frac{R_3}{R_4 + R_3} \frac{R_1 + R_2}{R_2} \tag{4}$$

The thing to consider when changing the gain is the Common Voltage(Vc), as changing the gain setting can cause the structure unsymmetrical. If you take Vc into consideration, the OPA1 output voltage can be as shown in Equation 5.

$$V_o = \left(\frac{V_{DAC} + OPA0 - IR_S - V_c}{R_1 + R_2} * R_1 + IR_S + V_c - V_c \right) \left(1 + \frac{R_4}{R_3} \right) + V_{offset} + V_c = \left(\frac{R_1}{R_1 + R_2} V_{DAC} + OPA0 + \frac{R_2}{R_1 + R_2} IR_S \right) \frac{R_3 + R_4}{R_3} + V_{offset} + \left(\frac{R_1}{R_1 + R_2} * \frac{R_3 + R_4}{R_3} - 1 \right) * V_c \tag{5}$$

To control the structure mismatch influence, you need to control the voltage level of Vc. That is why you need to assign the MCU GND to be closed to the Rs on the PCB layout.

5.3 Current Calibration Method

This section shows how to use ADC to calibrate the circuit in order to get a more accurate result.

To realize this calibration, you need to make some changes to the peripheral setting. For these working structures and used ADC channels, see Figure 5-4. For the total calibration, it is suggested to do it under 0 loading/charger. Otherwise, it influences the GAIN factor calibration accuracy. For the condition when the current cannot ensure to be 0 after the MCU power on, see Section 8.3 for a further description. In the following part, every calibration detail following the real code calibration flow is shown.
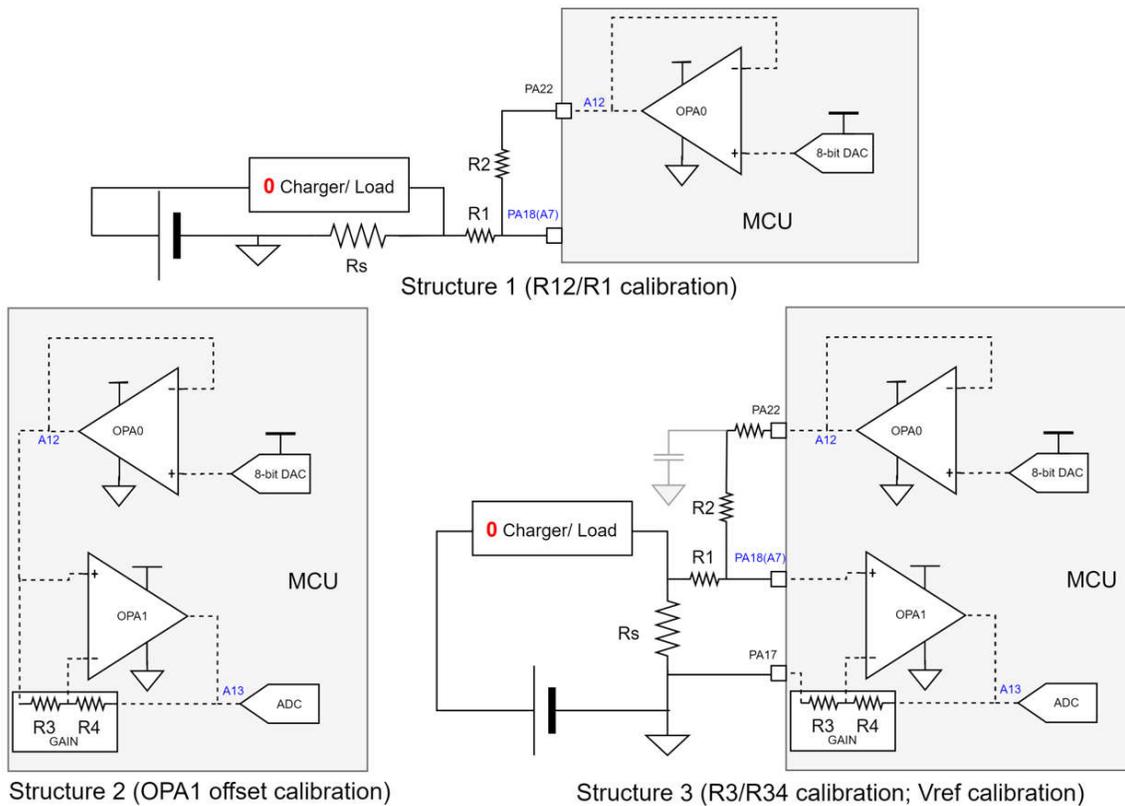
Figure 5-4. Current Calibration Structure

5.3.1 (R1+R2)/R2 calibration

For more information, see structure1 in Figure 5-4. ADC channel 12 and ADC channel 7 is used to test the voltage divided by R1 and R2. You will also increase the DAC output to make OPA1 output closed to the ADC full range, to increase the accuracy. In order to reduce the resistance influence of GPIO trace, choose to test the voltage at OPA0 internal ADC trace directly, instead of the GPIO output (PA22/A4). Equation 6 is the calibration equation based on ADC detected numbers:

$$\frac{R_1 + R_2}{R_2} = \frac{N_{A12}}{N_{A12} - N_{A7}} \tag{6}$$

In this current detection demo, as the VDD=1.8 V and the selected internal reference is 1.4 V. In order to get a high changeable OPA1 voltage output range with GAIN32 and GAIN16 setting paired with DAC change, we choose to make (R1+R2)/R1 = 8.

5.3.2 OPA1 Voffset calibration

The reason to do Voffset calibration is that you want to reduce its influence when calibrating R3/(R4+R3). For more information, see structure 2 in Figure 5-4. You will change the signal connection on the OPA1 input, from GND(MSEL) and PA18(PSEL) to OPA0 output (MSEL and PSEL). By comparing the voltage at OPA0 and OPA1 output, you know the OPA1 output offset. In this demo, two different GAIN settings are used, you need to do two offset calibrations as well. Equation 7 shows the equation based on ADC detected numbers.

$$N_{Voffset} = N_{A13} - N_{A12} \tag{7}$$

Although this adds some gain error as the R3 value will be a little different with difference structures, the key reason why we change the structure is that the Voffset error is acceptable as the output voltage offset is small after choopering, and a DAC offset is needed as the voltage offset can be bidirectional.

5.3.3 R3/(R4+R3) calibration

For more information, see structure 3 in Figure 5-4. The structure of R3/(R4+R3) calibration is the same with the final current test. The difference lies on that the DAC output is increased to make the OPA1 output to be closed to ADC full range in order to get higher accuracy. Instead of testing the voltage at PA18, use the calibrated (R1+R2)/R1 value and the detection result at A12 to calculate the PA18 voltage. This can help increase the resolution and reduce the DAC noise, compared with directly connecting OPA0 to PSEL MUX. For this demo, you also need to do two different R3/(R4+R3) calibration. Equation 8 shows the equation based on ADC detected numbers.

$$\frac{R_3}{R_4 + R_3} = \frac{\frac{R_1}{R_1 + R_2}*N_{A12}}{N_{A13} - N_{Voffset}} \tag{8}$$

5.3.4 Vref calibration

For the Vref calibration, you control the OPA1 output voltage to be closed to ½ ADC reference voltage and then save this value. For more information, see structure 3 in Figure 5-4. Peripheral setting is the same with the final current detection structure. For this demo, two GAIN settings are used and two Vref calibration results are needed to be saved.

6 Solution Evaluation Steps

6.1 Step1: Hardware Preparation

If you want to evaluate this total solution, you need to make the gauge board first.

In order to do the test and evaluate the current detection performance, you also need to prepare a source meter or other battery test machines to control the battery charge and discharge.

6.2 Step2: Evaluation

Before you start, there are some settings related to evaluation in UserConfig.h.

```
UserConfig.h ×
6 //#define DEBUG          //Transmit calculated data to COM port
7 //****************************Algorithm data output mode selection*****************
8 //#define OUTPUT_MODE     (NO_OUTPUT)
9 #define OUTPUT_MODE     (UART_OUTPUT)
10
11 //*****************************Customized configuration settings*****************
12 #define VCELL_FACTOR                                        (3.2)
13 #define SHUNT_RESISTOR_VALUE_R                              (0.02)
14 #define NTC_PAIR_RESISTOR_VALUE                             (10000)
15 #define ADC_AVERAGE_TIMES                                   (2)
16 #define ADC_SAMPLE_TIME                                     (10)
17 #define CURT_CALI_TEMP_THD                                   (5)
```

Figure 6-1. Mode Setting

If you enable the definition of "DEBUG", you can see the current calibration data output from the UART1.

For different output modes, "UART_OUTPUT" means enabling data output through UART1. Then you can observe the running parameters on the GUI. "NO_OUTPUT" means terminating the UART data output.

For customized configuration settings, you may need to make change for "VCELL_FACTOR", "SHUNT_RESISTOR_VALUE_R" and "NTC_PAIR_RESISTOR_VALUE" according to your hardware setup.

The hardware test setup is shown in Figure 6-2. You need the MSPM0 Gauge board and a real battery for test. You can insert the battery into the socket and charge or discharge it on J2. You can also connect a power supply to J3 directly.

The detection data (Vcell, Icell, Tcell) comes from the MSPM0 analog peripherals. Then it sends the data to the GUI through UART1 on J4. The GUI can help to record the battery running data for further analysis.

In order to ensure the voltage of programming pins mach with the debug port of MUC, when programming the MSPM0, use a jumper to connect VMCU and VEx on J1. As the UART1 port supports up to 5 V input and output voltage, When evaluating, use a jumper to connect VMCU and VIn on J1.
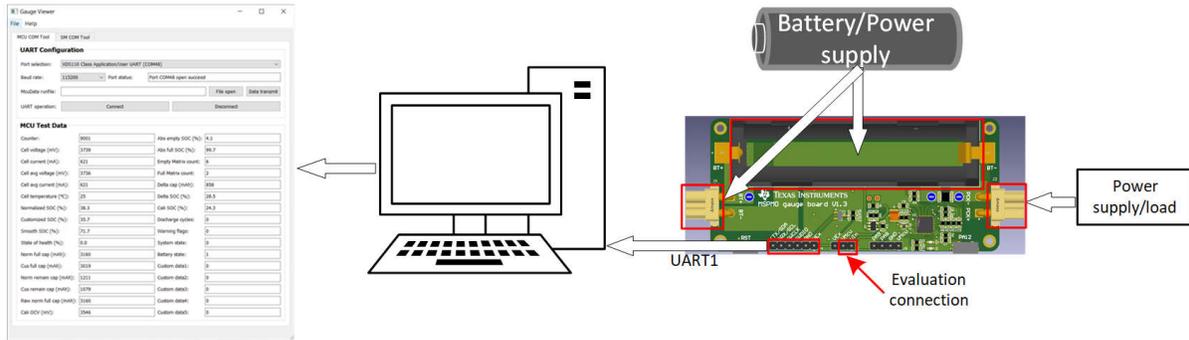

Figure 6-2. Hardware Setup and Connection

7 MSPM0 Gauge Solution Test Results

7.1 Calibration Test Result

After you enable the "DEBUG", you can output the calibration data through UART1. This data can help to evaluate the calibration error. The data shown in this part are all from one MSPM0L1306 with 1.8 V supply and comes from a 12-bit ADC with 1.4 V internal reference under 25°C.

Figure 7-1 shows the test data of (R1+R2)/R1. You can see that the value fluctuation is ±0.13%.
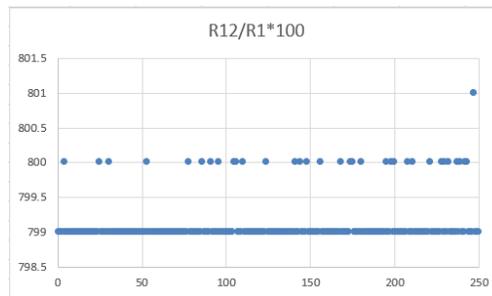
Figure 7-1. Tested (R1+R2)/R1 Under 25C

Figure 7-2 shows the voltage output offset with different gain. All of them are below than 0.7 mV.
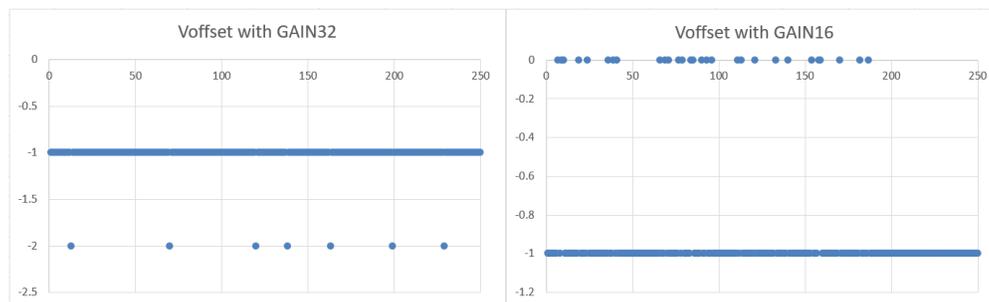
Figure 7-2. Tested OPA1 Offset Under 25°C

Figure 7-3 shows the test data of (R3+R4)/R3. You can see that the value fluctuation under GAIN32 is ±0.11%. For GAIN16, it is ±0.13%.
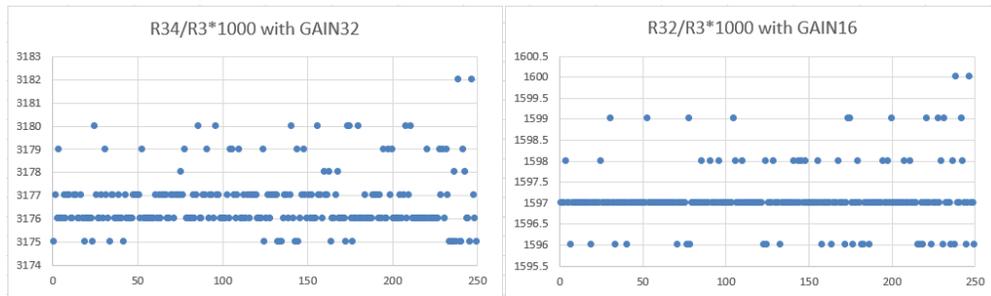

Figure 7-3. Tested (R3+R4)/R3 Under 25°C

To evaluate the calibration accuracy, compare the calculated Vref through the Equation 2 and the tested Vref through ADC channel 13, when the current is 0. Pay attention that the tested data is the raw data from 12-Bit ADC with 1.4 V reference. The Vref is about 0.7 V, close to ½ ADC reference.

You can see that the tested Vref almost only have 1 or 2 ADC count change in two GAIN settings. If you treat the tested Vref as the accurate value for the GAIN32 setting, the max calculated Vref error is about 0.2%. For GAIN16 setting, it is about 0.15%.

As seen from these results, you can say that the self-calibration structure is very stable and has a very high accuracy.
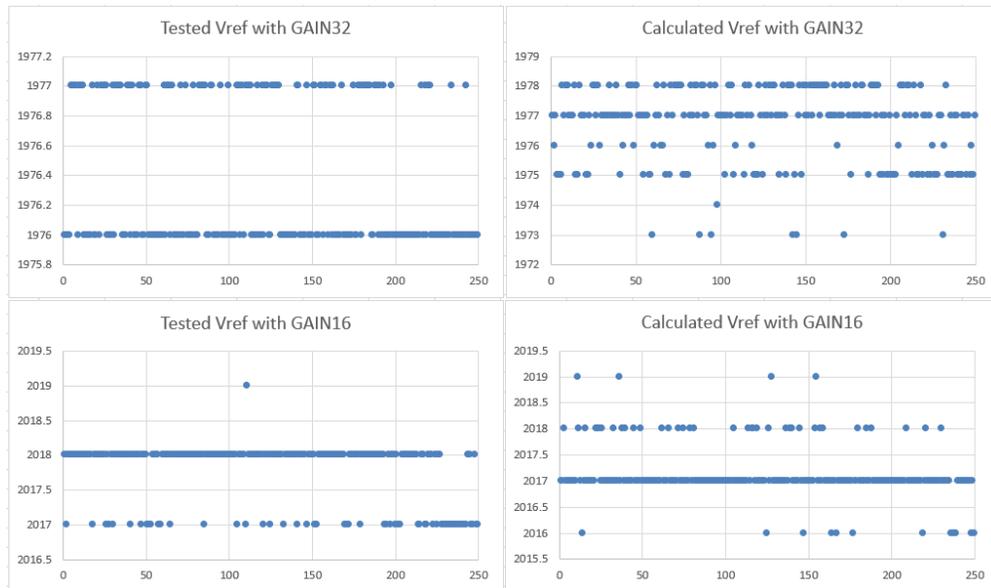
**Figure 7-4. Tested and Calculated Vref Under 25°C**

7.2 Current Detection Result

This section shows the current detection test performance under different temperature of MSPM0, using the source meter at the same time. In this current detection demo, the used shunter resistor is 20 mR. The GAIN setting changes from GAIN32 to GAIN16 when the current reaches about 1A, and it changes from GAIN16 to GAIN32 when the current reaches about 0.7A.

To reach higher current detection resolution, it is okay to use large resistor value, but on the GAIN side, this solution is limited by the max internal GAIN (32x).

7.2.1 Test Under 25°C

Figure 7-5 shows the current test pattern (left) and the detection error (right) under 25°C. You can find that this solution can control the error within 5 mA in the range of ±2000 mA. The error percentage is below than ±0.25% at ±2000 mA.
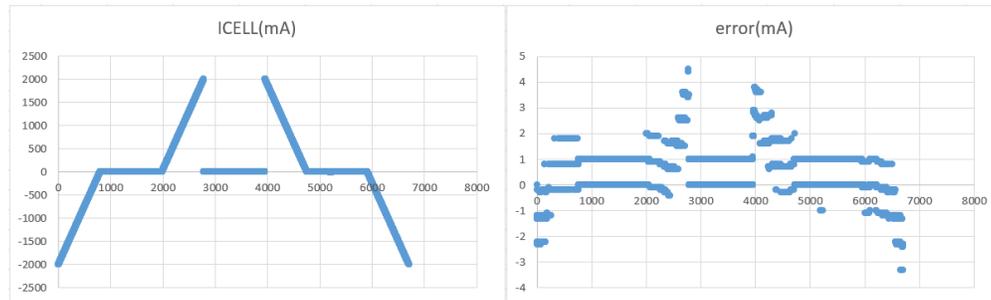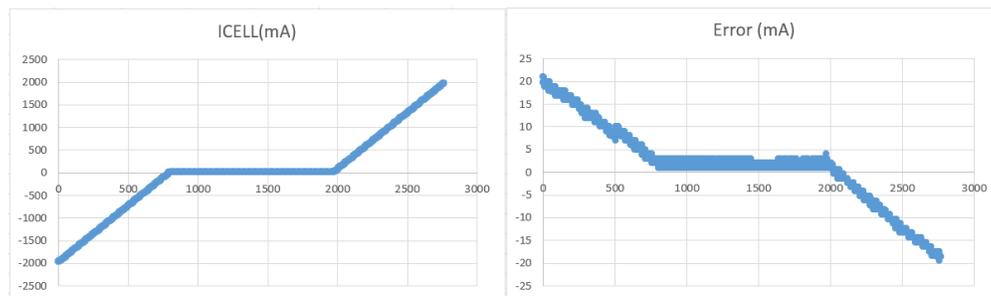

Figure 7-5. Current Test Pattern and its Error Under 25°C

7.2.2 Test Under 0°C

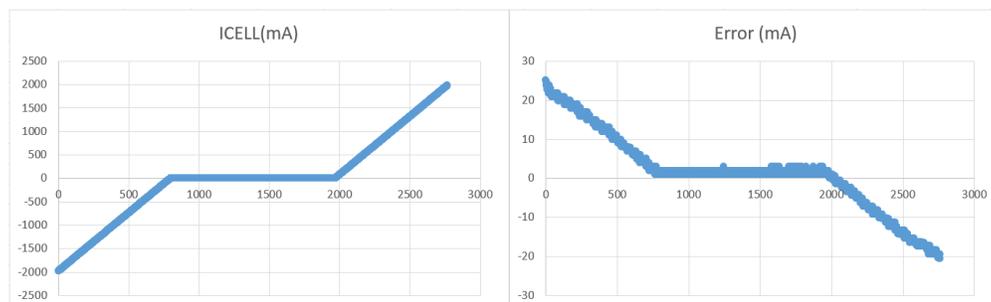Figure 7-6 shows the current test pattern (left) and the detection error (tight) under 0°C. However, the GAIN factor is gotten at 25°C. That is why you can find a 4 mA offset when the current is 0. After you transmit this 4 mA current into the output voltage shift and compare it with Vref, you can evaluate the GAIN factor shift is about 0.4% (About 2.56 mV output voltage shift with Vref=700 mV).

Under this condition, you can find that this solution can control the error within 22 mA in the range of ±2000 mA. The error percentage is below than ±1.1% at ±2000 mA.

Figure 7-6. Current Test Under 0°C With 25°C Calibration

See what happens when you do GAIN factor calibration under 0°C. You can find that offset current is reduced to 3 mA. However, the error reaches to 25 mA in the range of ±2000 mA. The error is below than ±1.3% at ±2000 mA.

Figure 7-7. Current Test Under 0°C With 0°C Calibration

You will see that the current detection error is increased when the GAIN factor is gotten under 0°C. Considering the GAIN factor shift, you would assume the increase error lies on the shunter resistor value shift.

7.2.3 Test Under 50°C

Figure 7-8 shows the current test pattern (left) and the detection error (right) under 50°C and the GAIN factor is gotten at 25°C. In this high temperature, the offset current is same with the one at room temperature. You would assume there is almost no GAIN factor shift.

Under this condition, you can find that this solution can control the error within 11 mA in the range of ±2000 mA. The error is below than ±0.6% at ±2000 mA.
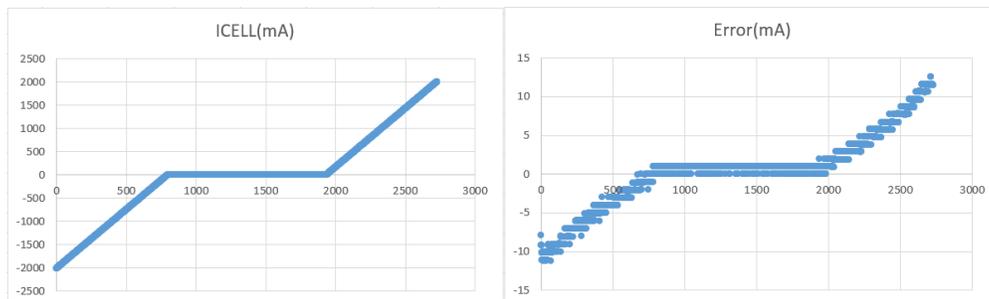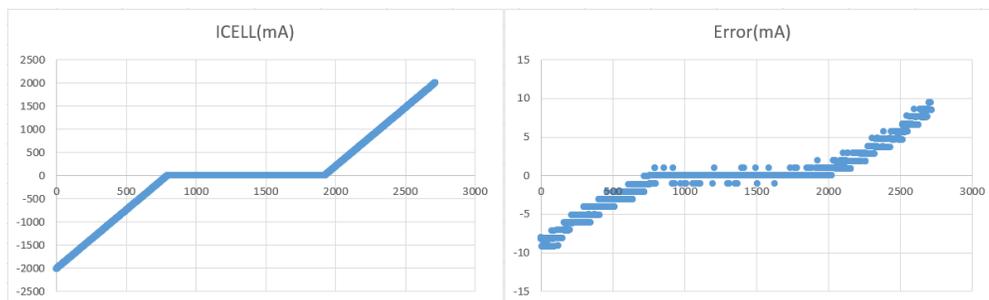

Figure 7-8. Current Test Under 50°C With 25°C Calibration

See what happens when you do GAIN factor calibration under 50°C. You can find that offset current is little reduced. The error reaches to 9.5 mA in the range of ±2000 mA. The error percentage is below than ±0.5% at ±2000 mA.

Figure 7-9. Current Test Under 50°C With 50°C Calibration

7.2.4 Conclusion

Through the test shown above, it is known that the calculated GAIN factor only has about 0.2% error/shift from Section 7.1. Therefore, the following is assumed:

- The increase error with temperature change mostly lies on the shunter resistor value shift.
- The solution can supply a high accuracy current detection solution with lower than 1% error for full temperature range, if the shunter resistor temperature shift can be carefully handled.

7.3 Current Consumption Test

Figure 7-10 shows the current test result, under NO_OUTPUT mode, and after removing the tantalum capacitor, the temperature sensor, and voltage divider resistors. You just want to show the MCU power consumption of detection part. If customer wants to reach the same performance with full hardware setup, use the MOSFET or bipolar to control the voltage supply to resistor dividers.

EnergyTrace™ Profile	
Name	Live
∨ System	
Time	35 sec
Energy	0.786 mJ
∨ Power	
Mean	0.0252 mW
Min	0.0069 mW
Max	5.6453 mW
∨ Voltage	
Mean	3.3000 V
∨ Current	
Mean	0.0076 mA
Min	0.0021 mA
Max	1.7107 mA

Figure 7-10. Current Consumption

8 Solution Summery and Improvement Direction

Seen from the tests above, it is shown that this solution can ensure a high accuracy across a large current detection range to meet some common application's requirement like BMS **without any factory calibration**.

However, this solution still has some error source that cannot be auto calibrated.

8.1 Shunter Resistor

Shunter resistor's accuracy across temperature strongly affects the current detection accuracy. To reduce this error, it is suggested to add an NTC to test the temperature of the shunter resistor and then do temperature compensation.

8.2 ADC and its Reference

ADC DC error source (INL, DNL, GAIN, Offset) affects the calibration accuracy. However, the Offset error has the largest influence, especially for low-voltage detection. For MSPM0L130x, the max offset error is -3.5mV, when using internal reference. If the used reference is 1.4 V, about 9 LSB error is caused from ADC offset. To reduce this error influence, you need to do an additional ADC offset calibration.

ADC reference error also takes the same influence like the GAIN error. For MSPM0L130x, the max reference error is ±1.5%. To reduce this error influence, you can do an additional ADC reference calibration, or use an external reference.

In a word, all the ADC error cannot be calibrated automatically and the cost is added.

8.3 Runtime Calibration

As this demo will only do calibration at first power up, if the temperature changes, the GAIN factor and Voffset are also influenced. If you want to realize runtime calibration, the power-on current cannot be 0, especially for the MCU on system side. The structure in Figure 8-1 can be used.
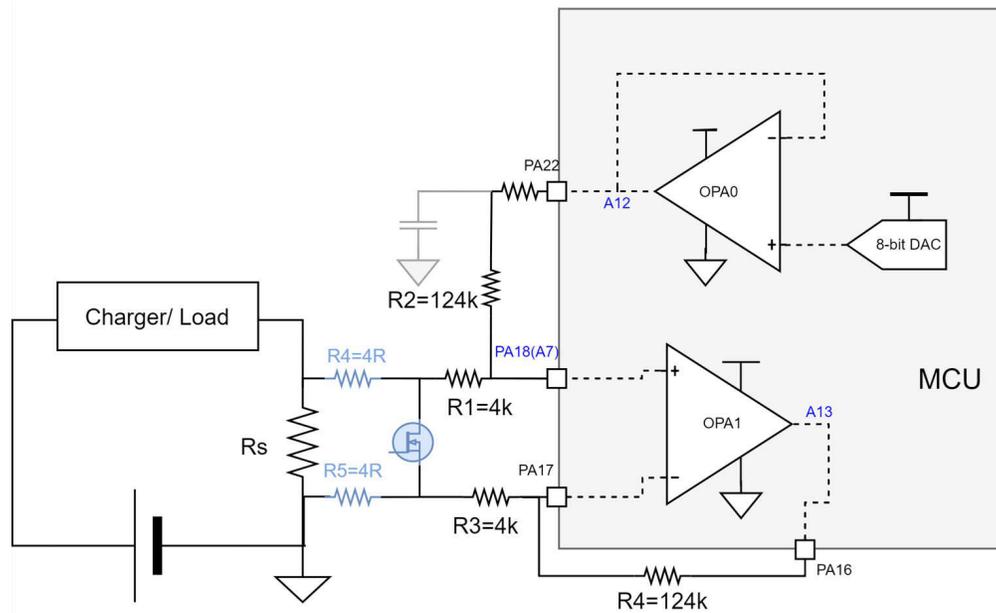
Figure 8-1. Runtime Calibration Hardware Structure

In this structure, during calibration, the MOSFET should be enabled. The current across Rs are divided by R4 and R5, which are treated as the common voltage input into the amplifier. From Equation 5, you know that the mismatch between R1, R2, R3 and R4 causes the common voltage to have a large influence on the OPA output. So, you should make sure that R2=R4 and R3=R1. As internal gain has larger gain error (2.7% with GAIN32), use external gain instead.

For Voffset calibration, it can use the Voffset calibrated with internal gain to simulate the real Voffset. For R3/(R3+R4) calibration, it can be done by detection the voltage on ADC channel 7 and ADC channel 13. For (R1+R2)/R2 calibration, you can use a default value at the beginning and do the calibration when the load current is low and the error influence can be ignored. It is not suggested that you use another ADC to detect the voltage across Rs in the calibration to make up the error cause from the load current. First, is that this detection voltage will be positive only when in discharge mode. Second, the tested result is affected by ADC offset error.

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