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Application Basics for the MSP430 14-Bit ADC

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Application Basics for the MSP430 14-Bit ADC

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ABSTRACT

This application report gives a detailed overview of several applications for the 14-bit analog-to-digital converter (ADC) of the MSP430 family. Proven software examples and basic circuitry are shown and explained. The 12-bit mode is also considered, when possible. The *References* section at the end of the report lists related application reports in the MSP430 14-bit ADC series.

1 Introduction

The application report *Architecture and Function of the MSP430 14-Bit ADC Application Report*[1] explained the architecture and function of the MSP430 14-bit analog-to-digital converter (ADC). The hardware (registers, current source, used reference, interrupt handling, clock generation) was explained in detail and typical ADC characteristics were shown.

Figure 1 shows the block diagram of the MSP430 14-bit ADC.

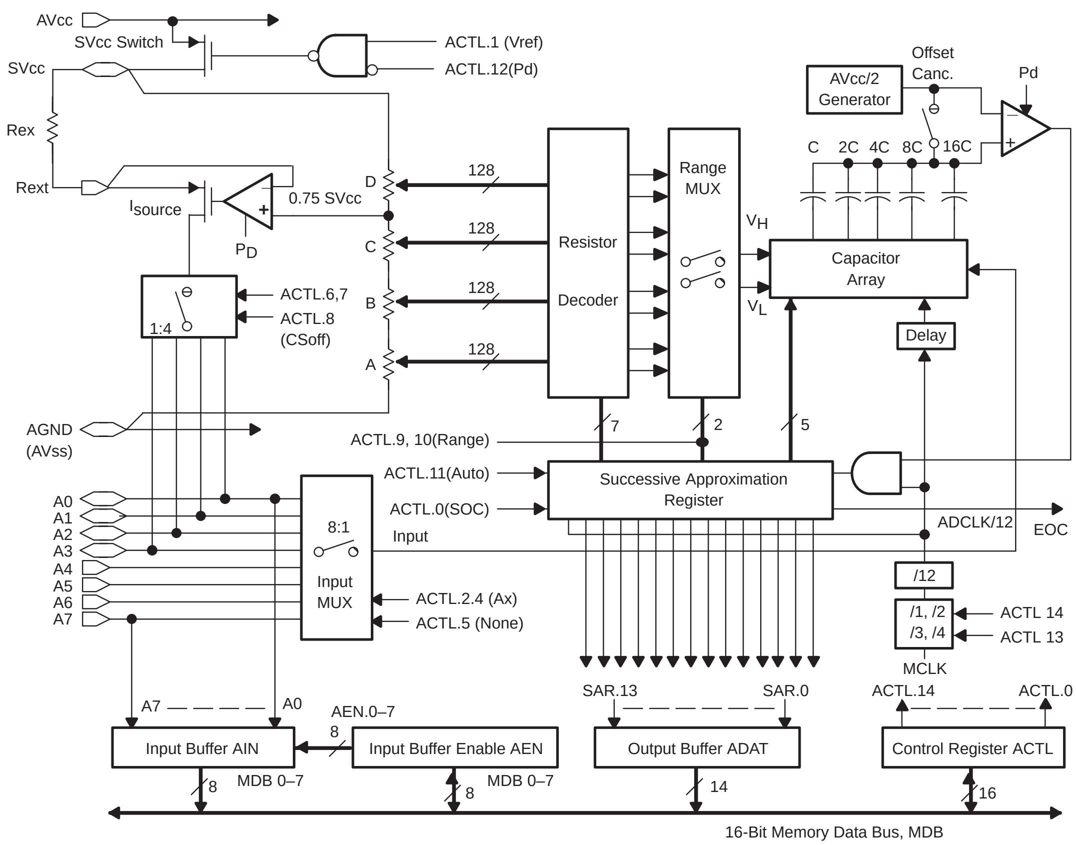

Figure 1. MSP430 14-Bit ADC Hardware

2 Applications

This application report shows several methods for connecting resistive sensors, bridge assemblies, and analog signals to the ADC. Solutions are given for the 12-bit and 14-bit conversions, with and without using the integrated current source. The equations shown result in voltages and resistances. To calculate the sensor values (pressure, current, temperature, light intensity a.s.o) normally with non-linear equations, refer to the following sections of Chapter 5 of the *MSP430 Application Report* [2]:

- Table Processing
- Temperature Calculations for Sensors
  - Table Processing for Sensor Calculations
  - Algorithms for Sensor Calculations
  - Coefficient Calculations for the Equations
- The Floating Point Package

2.1 Connection of Analog Signals and Sensors

Figure 2 shows possible methods for connecting analog signals to the ADC. The methods shown are valid for the 12-bit and 14-bit conversion modes:

1.	Current supply for resistive sensors	Rsens1 at analog input A0
2.	Voltage supply for resistive sensors	Rsens2 at analog input A1
3.	Direct connection of input signals	Vin at analog input A7
4.	Four-wire circuitry with current supply	Rsens3 at output A3 and inputs A4 and A5
5.	Reference diode with voltage supply	Dr1 at analog input A6
6.	Reference diode with current supply	Dr2 at analog input A2

The resistance of the wiring, Rwire, in the following equations may be neglected if it is low compared to the sensor resistance.

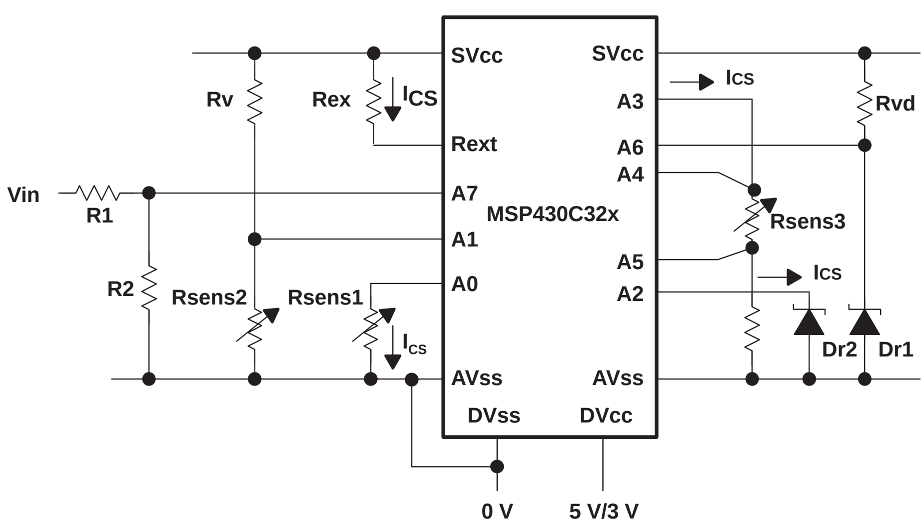

Figure 2. Possible Connections to the ADC

2.1.1 Current Supply for Sensors

The ADC formula for the resistor Rx in figure 3 (Rsens1 in Figure 2) which is fed from the current source is (14-bit conversion):

$$N = \frac{V_{A0}}{V_{REF}} \times 2^{14} = \frac{Ics \times (Rx + 2 \times Rwire)}{V_{REF}} \times 2^{14}$$

$$N = \frac{\frac{0.25 \times V_{REF}}{Rex} \times (Rx + 2 \times Rwire)}{Vref} \times 2^{14} = \frac{Rx + 2 \times Rwire}{Rex} \times 2^{12}$$

This leads to:

$$Rx = Rex \times \frac{N}{2^{12}} - 2 \times Rwire$$

For the 12-bit conversion the formula is:

$$N = \frac{V_{A0} - n \times 0.25 \times V_{REF}}{V_{REF}} \times 2^{14} = \left(\frac{Rx + 2 \times Rwire}{Rex} - n\right) \times 2^{12}$$

This leads to:

$$Rx = Rex \times \left(\frac{N}{2^{12}} + n\right) - 2 \times Rwire$$

Where:			
	N	ADC conversion result for resistor Rx	
	Rx	Sensor resistance	[Ω]
	Rex	Current source resistance (defines Ics)	[Ω]
	Rwire	Wiring resistance (one direction only)	[Ω]
	V_{REF}	Voltage at terminal SVcc (internal or external reference)	[V]
	V_{A0}	Voltage at the analog input A0	[V]
	n	Range number (0,1,2,3 for ranges A,B,C,D)	
	Ics	Current generated by the current source	[A]

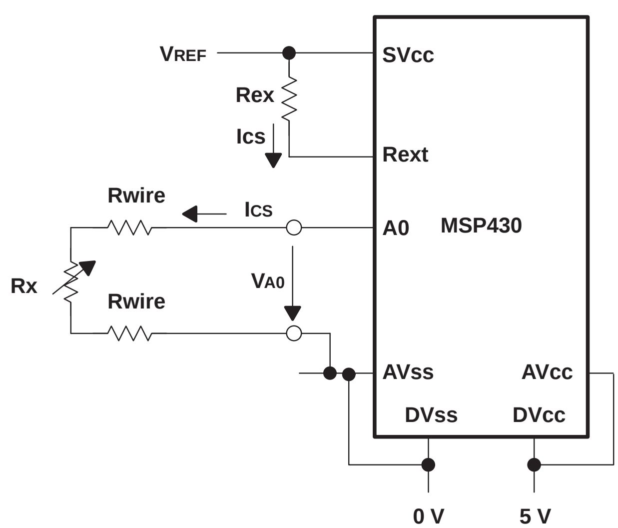

Figure 3. Current Supply for the Sensor Rx

If the resistance of the wires may be neglected (Rx >> Rwire) then the above formulas simplify to (14-bit conversion):

$$N = \frac{Rx}{Rex} \times 2^{12} \qquad Rx = Rex \times \frac{N}{2^{12}}$$

For the 12-bit conversion the formulas become:

$$N = \left(\frac{Rx}{Rex} - n\right) \times 2^{12} \qquad Rx = Rex \times \left(\frac{N}{2^{12}} + n\right)$$

2.1.2 Voltage Supply for Sensors

The ADC formula for the resistor Rx in figure 4 (Rsens2 in Figure 2) which is connected to Vref through the series resistor Rv is (14–bit conversion):

$$N = \frac{V_{A1}}{V_{REF}} \times 2^{14} = \frac{Rx + 2 \times Rwire}{Rv + Rx + 2 \times Rwire} \times 2^{14} \rightarrow Rx = Rv \times \frac{N}{2^{14} - N} - 2 \times Rwire$$

For the 12-bit conversion the formula is:

$$N = \left(\frac{V_{A1}}{V_{REF}} - n \times 0.25\right) \times 2^{14} = \left(\frac{Rx + 2 \times Rwire}{Rv + Rx + 2 \times Rwire} - n \times 0.25\right) \times 2^{14}$$

This leads to:

$$Rx = Rv \times \frac{1}{\frac{2^{14}}{N + n \times 2^{12}} - 1} - 2 \times Rwire$$

Where:	Rv	Resistance of the series resistor	[Ω]
	V_{A1}	Voltage at the analog input A1	[V]

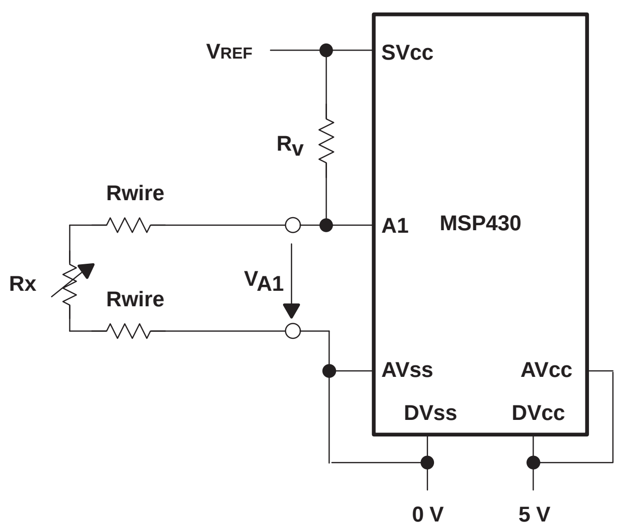

Figure 4. Voltage Supply for the Sensor Rx

If the resistance of the wires can be neglected (Rx >> Rwire), the above formulas simplify for the 14-bit conversion to:

$$N = \frac{Rx}{Rv + Rx} \times 2^{14} \rightarrow Rx = Rv \times \frac{N}{2^{14} - N}$$

For the 12-bit conversion the formula becomes:

$$N = \left(\frac{Rx}{Rv + Rx} - n \times 0.25\right) \times 2^{14} \;\rightarrow\; Rx = Rv \times \frac{1}{\dfrac{2^{14}}{N + n \times 2^{12}} - 1}$$

2.1.3 Four-Wire Sensors Circuit

Four-wire circuits eliminate errors due to the voltage drop caused by the connection lines (Rwire) to the sensor. Instead of two lines, four are used—two for the measurement current, and two for the sensor voltages. The two sensor lines do not carry current; the current at the analog inputs is in the nanoamp range, so no voltage drop falsifies the measured values. The four-wire circuit is used with a heat volume counter shown in the section *Heat Volume Counter* of Chapter 4 of the *MSP430 Application Report*.[2]

Figure 5 shows the four-wire circuit with its current supply.

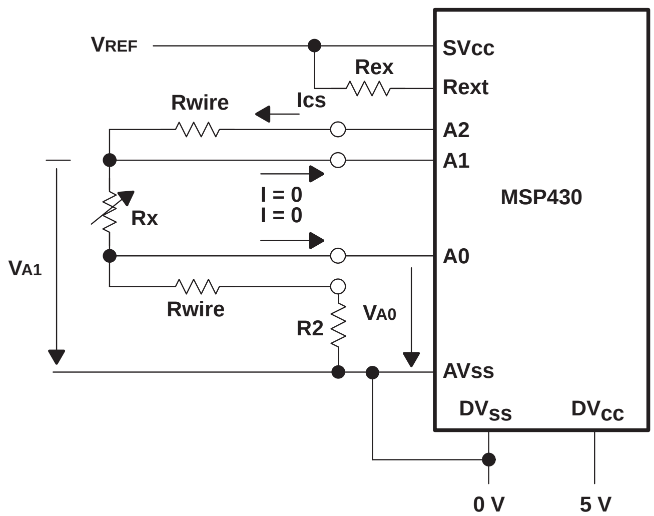

Figure 5. Four-Wire Circuit With Current Supply

The difference ΔN of the two measurement results for the analog inputs A1 and A0 is:

$$\Delta N = (V_{A1} - V_{A0}) \times \frac{2^{14}}{V_{REF}} = Ics \times ((Rx + Rwire + R2) - (Rwire + R2)) \times \frac{2^{14}}{V_{REF}}$$

$$\Delta N = \frac{0.25 \times V_{REF}}{Rex} \times Rx \times \frac{2^{14}}{V_{REF}} = \frac{Rx}{Rex} \times 2^{12}$$

This gives for Rx:

$$Rx = Rex \times \frac{\Delta N}{2^{12}}$$

Where: ΔN Difference of the two ADC results (here NA1 – NA0)

As the two final equations for ΔN and Rx show, the influence of the Rwire resistances disappears completely.

NOTE:

The two formulas above are valid for 14-bit and 12-bit conversions. If the 12-bit ADC results are measured in different ADC ranges, then the 12-bit results need a correction (the missing two MSBs—13th and 14th bits—of the ADC results must be added):

Range A: 0   Range B: 1000h   Range C: 2000h   Range D: not possible

Resistor R2 is necessary, because the ADC cannot measure down to AVss (0 V) due to saturation effects. R2 may be quite small; it is only necessary to get above the saturation voltage—normally less than 30 ADC steps.

The software to measure ΔN is shown next. The hardware of Figure 5 is used:

```
; Measure upper leg of Rx at input A1 and store ADC value.
; The Current Source is connected to A2
;
   MOV    #RNGAUTO+CSA2+A1+VREF,&ACTL   ; Define ADC
   CALL   #MEASR                        ; Upper leg voltage of Rx (A1)
   MOV    &ADAT,R5                      ; Store A1 value in R5
;
; Measure lower leg of Rx at input A0. Current Src to A2
;
   MOV    #RNGAUTO+CSA2+A0+VREF,&ACTL   ; Define ADC
   CALL   #MEASR                        ; Lower leg voltage of Rx (A0)
;
; The difference delta N of the 2 measurements is proportional
; to the value Rx: Rx = Rext x deltaN x 2∧–12
;
   SUB    &ADAT,R5                      ; R5 contains delta N
   ...                                  ; Calculate Rx
```

2.1.4 Connection of Bridge Assemblies

Bridge assembly sensors are best known for pressure measurement. The voltage difference (Vp – Vm) between the two bridge legs changes with the pressure to be measured. For clarity, the temperature measurement circuitry that is normally necessary is not included.

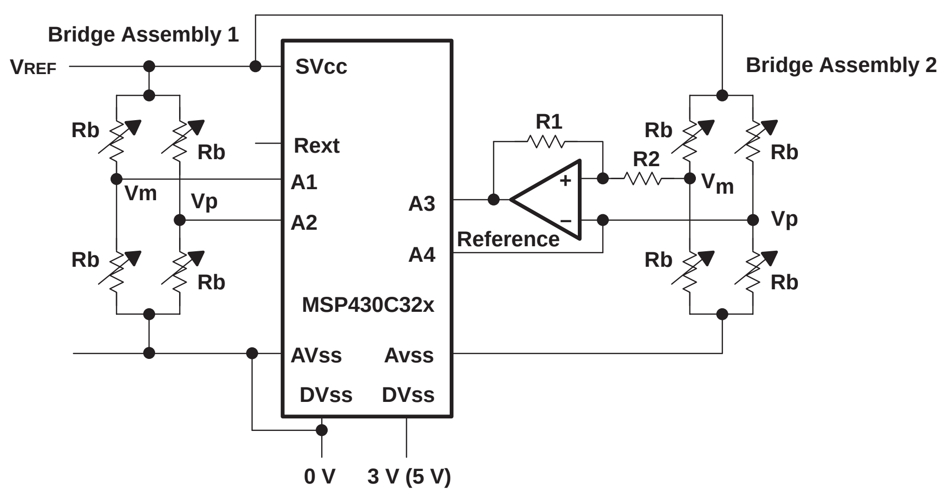

Figure 6. Connection of Bridge Assemblies

On the left side of Figure 6, a bridge assembly creates a voltage difference large enough to be measured by the ADC with appropriate resolution. The measurement result is the difference of the two ADC results measured at the A1 and A2 analog inputs.

$$\Delta N = \frac{V_{A2} - V_{A1}}{V_{REF}} \times 2^{14} \rightarrow \Delta V = \Delta N \times V_{REF} \times 2^{-14} = V_{REF} \times \frac{\Delta Rb}{Rb}$$

Where:

ΔN	Difference of the two ADC results (here N_{A2}–N_{A1})	
VAx	Voltage at the ADC input Ax measured to AVss	[V]
ΔV	Difference of the two bridge leg voltages (here V_{A2}–V_{A1})	[V]
ΔRb	Change of a single bridge resistance due to measured value	[Ω]
Rb	Nominal value of a single bridge resistor	[Ω]

With the above equations, the interesting bridge output value $\Delta Rb/Rb$ becomes:

$$\frac{\Delta Rb}{Rb} = \Delta N \times 2^{-14}$$

If the difference of the two measurements is too small to be used, an operational amplifier may be used as shown on the right of Figure 6. Here the possibility to measure the reference voltage (one of the two bridge legs) is shown too: analog input A4 measures the reference that can be used for a better result together with the input A3.

The voltage difference ΔV between the analog inputs A3 and A4 is:

$$\Delta V = V_{A3} - V_{A4} = (v \times (Vp - Vm) + Vp) - Vp = v \times (Vp - Vm)$$

$$\Delta V = \frac{R1}{R2} \times (Vp - Vm) = \frac{R1}{R2} \times \frac{V_{REF}}{2 \times Rb}((Rb + \Delta Rb) - (Rb - \Delta Rb)) = \frac{R1}{R2} \times V_{REF} \times \frac{\Delta Rb}{Rb}$$

The same voltage difference ΔV described with the ADC equation is:

$$\Delta V = V_{A3} - V_{A4} = \left(\frac{N_{A3}}{2^{14}} - \frac{N_{A4}}{2^{14}}\right) \times V_{REF} = (N_{A3} - N_{A4}) \times \frac{V_{REF}}{2^{14}}$$

Combining the two equations above delivers the interesting two equations:

$$\Delta N = v \times \frac{\Delta Rb}{Rb} \times 2^{14} = \frac{R1}{R2} \times \frac{\Delta Rb}{Rb} \times 2^{14}$$

For the bridge output value $\Delta Rb/Rb$, the following equation is used: the value $\Delta Rb/Rb$ is necessary for the final calculation of the measured item, e.g., pressure $p = f(\Delta Rb/Rb)$:

$$\frac{\Delta Rb}{Rb} = \frac{\Delta N}{v \times 2^{14}} = \frac{R2}{R1} \times \frac{N_{A3} - N_{A4}}{2^{14}}$$

Where:

ΔN	Difference of the two ADC results (here NA3–NA4)	
ΔV	Voltage difference of analog inputs A3 And A4 (VA3–VA4)	[V]
v	Amplification of the operational amplifier: v=R1/R2	
Vp	Voltage of the bridge leg connected to the noninverting input	[V]
Vm	Voltage of the bridge leg connected to the inverting input	[V]

If the reference input (analog input A4 in Figure 6) is not implemented, then the difference of two measurements at the amplifier output (analog input A3 in Figure 6) is used. The voltage difference ΔV between two measurements is:

$$\Delta V = V_{A31} - V_{A30} = (v + 0.5) \times Vref \times \frac{\Delta Rb1 - \Delta Rb0}{Rb}$$

The same voltage difference ΔV described with the ADC equation is:

$$\Delta V = V_{A31} - V_{A30} = \left(\frac{N_{A31}}{2^{14}} - \frac{N_{A30}}{2^{14}}\right) \times V_{REF} = (N_{A31} - N_{A30}) \times \frac{V_{REF}}{2^{14}}$$

The two equations above deliver the equation for ΔN, e.g., the ADC value representing the difference of two weights:

$$\Delta N = N_{A31} - N_{A30} = (v + 0.5) \times \frac{\Delta Rb1 - \Delta Rb0}{Rb} \times 2^{14} = \left(\frac{R1}{R2} + 0.5\right) \times \frac{(\Delta Rb1 - \Delta Rb0)}{Rb} \times 2^{14}$$

And for the difference of the two bridge output values that represent for example a weight difference. The value $\Delta Rb/Rb$ is used for the final calculation of the measured item, e.g., the weight G = f(ΔRb/Rb):

$$\frac{\Delta Rb}{Rb} = \frac{\Delta Rb1 - \Delta Rb0}{Rb} = \frac{N_{A31} - N_{A30}}{(v + 0.5) \times 2^{14}} = \frac{N_{A31} - N_{A30}}{\left(\frac{R1}{R2} + 0.5\right) \times 2^{14}}$$

Where:	ΔN	Difference of the two ADC results (here N_{A31}–N_{A30})	
	N_{A30}	ADC result of the 1st measurement, e.g., the zero point of the bridge	
	N_{A31}	ADC result of the 2nd measurement, e.g., a weight measurement	
	ΔV	Voltage difference of two analog measurements (VA31–VA30)	[V]
	V_{A30}	Voltage at the analog input A3, e.g., for the zero point of bridge	[V]
	V_{A31}	Voltage at the analog input A3, e.g., a weight measurement	[V]
	v	Amplification of the operational amplifier: v=R1/R2	
	$\Delta Rb0$	Resistor deviation (Rb0–Rb) of the 1st measurement	[Ω]
	$\Delta Rb1$	Resistor deviation (Rb1–Rb) of the 2nd measurement	[Ω]
	ΔRb	Resistor difference (Rb1–Rb0)	
	Rb	Nominal value of a single bridge resistance	[Ω]

2.1.5 Reference Measurements

The simplest way to get a reference voltage is to use the supply voltage of the MSP430. If this is not possible, and a stable reference voltage is needed, e.g. for voltage measurements, then a reference diode can be used. Figure 7 shows two ways to connect a reference diode to the MSP430:

- The reference diode Dr1 is fed via the series resistor Rvd
- The reference diode Dr2 is fed by the current source of the MSP430

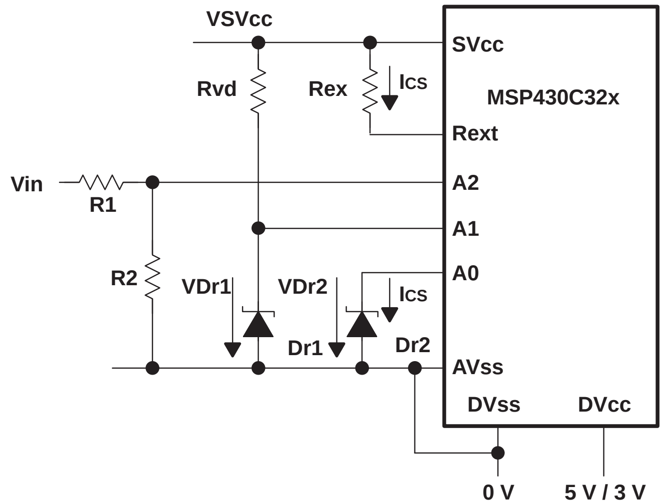

Figure 7. Connecting Reference Elements

If the external voltage Vin shown in Figure 7 is to be measured, then the following equations may be used. For reference purposes the voltage VDr is used, not the *unknown* supply voltage Vsvcc:

$$Vin = \frac{R1 + R2}{R2} \times \frac{Nin}{2^{14}} \times Vsvcc$$

The unknown voltage Vsvcc is fixed by the measurement of the reference voltage VDr:

$$V_{Dr} = \frac{N_{Dr}}{2^{14}} \times Vsvcc \rightarrow Vsvcc = \frac{2^{14}}{N_{Dr}} \times V_{Dr}$$

This leads to:

$$Vin = \frac{R1 + R2}{R2} \times \frac{Nin}{N_{Dr}} \times V_{Dr}$$

Where:	*Vin*	Input voltage to be measured	[V]
	Vsvcc	Supply voltage at terminal SVcc	[V]
	V_{Dr}	Voltage of the reference diode Dr	[V]
	Nin	ADC measurement result for the input voltage Vin	
	N_{DR}	ADC measurement result for the reference voltage V_{Dr}	
	R1, *R2*	Voltage divider for input voltage Vin	[Ω]

If the supply voltage Vsvcc is overlaid by hum (mains driven supply), then the referencing method shown above gives much better results if the reference diode Dr is measured twice—once before the input voltage Vin (N_{Dr0}), and once afterwards (N_{Dr1}). The two ADC results, N_{Dr0} and N_{Dr1}, are used as follows:

$$Vin = \frac{R1 + R2}{R2} \times \frac{2 \times Nin}{N_{Dr0} + N_{Dr1}} \times V_{Dr}$$

The calculation above uses the mean value of the measured values of the voltage Vsvcc (linear correction).

2.2 14-Bit Analog-to-Digital Conversion With Signed Signals

The MSP430 ADC measures unsigned signals from Vref, the voltage applied to the terminal SVcc (internal or external), to AVss. If signed measurements are necessary then a virtual zero point has to be provided. Signals above this zero point are treated as positive signals; signals below it are treated as negative ones. Four possibilities for a virtual zero point are shown in this chapter:

- Virtual ground IC: The zero point is provided by a special IC
- Split power supply: The zero point is provided by two power supplies
- Current source: The zero point is provided by the current source and a drop resistor
- Resistor divider: The zero point is provided by a resistor divider

The signal source is connected to the virtual zero point with its reference potential (first two solutions) or to the AVss potential (last two solutions).

2.2.1 Virtual Ground IC

With the *phase splitter* TLE2426, a common zero point is provided which lies exactly in the middle of the voltage between the Vref and the AVss potential. The reference voltage Vref may be internal (AVcc) or external. All signed input voltages are connected to this virtual ground with their reference potential. The virtual ground voltage (at analog input A0 in Figure 8) is measured after regular time intervals, and the measured ADC value is stored and subtracted from the measured analog input signal V1 (here at input A1). This results in a signed, offset corrected ADC value for the signal at the analog input A1. The virtual ground method is used with some electronic electricity meters shown in the *Electricity Meters* section of Chapter 4 of the *MSP430 Application Report*.[2]

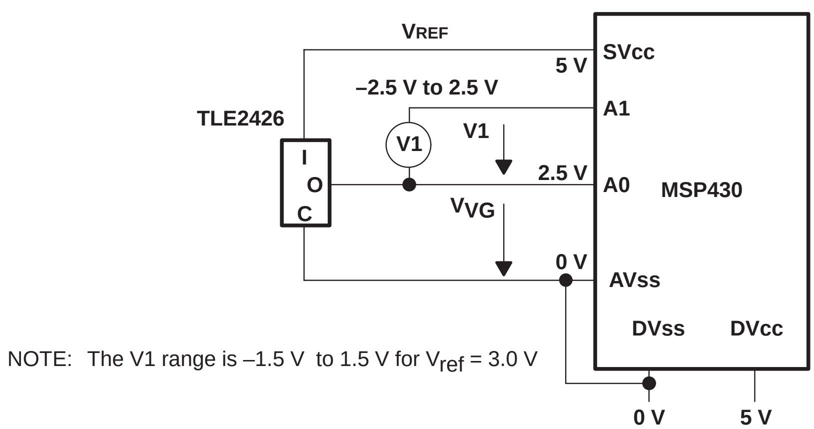

Figure 8. Virtual Ground IC for Signed Voltage Measurement

The formula for the difference of the ADC results ΔN is:

$$\Delta N = (N_{A1} - N_{A0}) = \frac{V1 + Vvg}{V_{REF}} \times 2^{14} - \frac{Vvg}{V_{REF}} \times 2^{14} = \frac{V1}{V_{REF}} \times 2^{14}$$

This leads to the formula for V1:

$$V1 = V_{REF} \times \frac{\Delta N}{2^{14}}$$

Where:	$V1$	Voltage to be measured	[V]
	ΔN	Difference of the two ADC results (here $N_{A1} - N_{A0}$)	
	V_{REF}	Voltage at the SVcc terminal measured against AVss terminal	[V]
	Vvg	Voltage at the A0 terminal ($0.5 \times$ Vref)	[V]

EXAMPLE: The virtual ground voltage at A0 is measured and stored in location VIRTGR (register or RAM). The value of VIRTGR is subtracted from the ADC value measured at input A1; this gives the signed, offset corrected value for the input signal at the A1 input. The measurement subroutine MEASR shown in section 4.1 is used.

```
VIRTGR .EQU   R6            ; Virtual Ground ADC value
;
; Measure virtual ground voltage at input A0 and store value
; for reference. MCLK = 3MHz: divide MCLK by 2
;
       MOV    #ADCLK2+RNGAUTO+CSOFF+A0+VREF,&ACTL
       CALL   #MEASR        ; Measure A0 (virtual ground)
       MOV    &ADAT,VIRTGR  ; Store result: 14–bit value
       ...;

; Measure analog input signal V1 (0 ...03FFFh) and compute
; a signed, offset corrected value for V1 (0E000h ...01FFFh)
;
       MOV    #ADCLK2+RNGAUTO+CSOFF+A1+VREF,&ACTL
       CALL   #MEASR        ; Measure A1 (input voltage V1)
       MOV    &ADAT,R5      ; Read ADC value for V1
       SUB    VIRTGR,R5     ; R5 contains signed delta N
    ...                     ; V1 = Vref x deltaN x 2^–14
```

2.2.2 Split Power Supply

With two power supplies, for example with 2.5 V and –2.5 V, a potential in the middle of the MSP430 ADC range can be created. Figure 9 shows this arrangement. All signed input voltages are connected to this voltage with their reference potential (0 V). The mid range voltage (at analog input A0) is measured after regular time intervals and the measured ADC value is stored and subtracted from the measured signal (here at analog input A1). This gives a signed, offset corrected result for the analog input A1. The split power supply method is used with some of the electronic electricity meters shown in the *Electricity Meters* section of Chapter 4 of the *MSP430 Application Report*.[2]

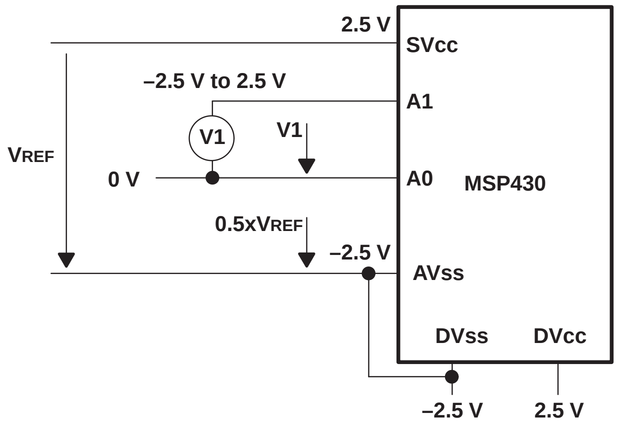

Figure 9. Split Power Supply for Signed Voltage Measurement

The formula for the difference of the ADC results ΔN is:

$$\Delta N = (N_{A1} - N_{A0}) = \frac{V1 + (0.5 \times V_{REF})}{V_{REF}} \times 2^{14} - \frac{(0.5 \times V_{REF})}{V_{REF}} \times 2^{14} = \frac{V1}{V_{REF}} \times 2^{14}$$

This leads to the formula for V1:

$$V1 = V_{REF} \times \frac{\Delta N}{2^{14}}$$

Where:	*V1*	Input voltage to be measured	[V]
	ΔN	Difference of the two ADC results (here N_{A1}–N_{A0})	
	V_{REF}	Voltage between the SVcc and the AVss terminals	[V]

The same software example can be used as shown before with the virtual ground IC.

2.2.3 Use of the Current Source

With the current source method shown in Figure 10, a voltage that is partially or completely below the AVss potential can be shifted into the middle of the used ADC range of the MSP430. This is accomplished by a drop resistor Rh whose voltage drop shifts the input voltage accordingly. This method is especially useful if differential measurements are necessary, because the ADC value of the signal's midpoint (zero point) is not available as easily as with the two methods shown previously. If absolute measurements are necessary, then a calibration or a measurement with a known input voltage equal to the zero point is needed.

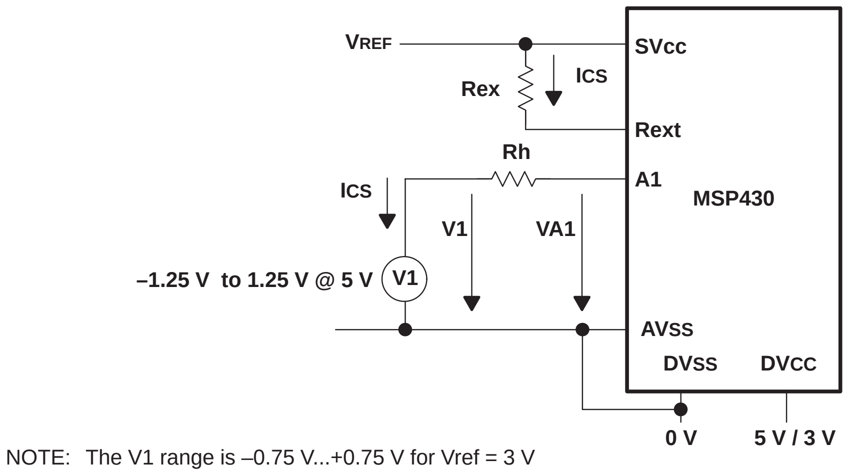

Figure 10. Current Source Used for Level Shifting

The example of Figure 11 shows an input signal V1 ranging from –1.25 V to 1.25 V. To shift the signal's zero voltage (0 V) to the midpoint voltage Vzv of the usable ADC range (this range is approximately 0.5 × Vref, so Vzv is 0.25 × Vref) a current Ics is used. The necessary current Ics to shift the input signal is:

$$I_{CS} = \frac{V_{zv}}{R_h} \rightarrow R_h = \frac{V_{zv}}{I_{CS}} = \frac{V_{zv}}{\frac{0.25 \times V_{REF}}{R_{ex}}}$$

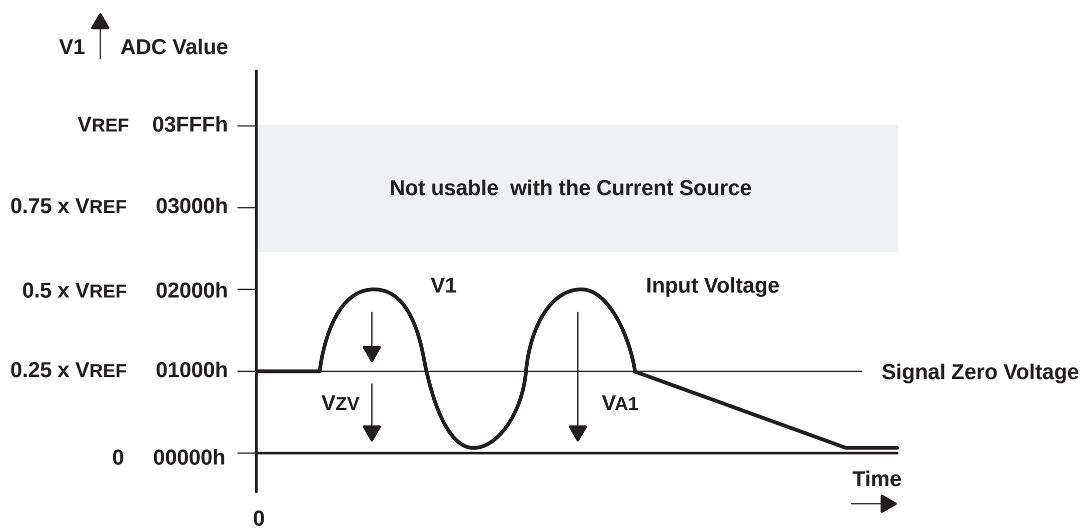

Figure 11. Signed Signals Shifted With the Current Source

Therefore the necessary shift resistor Rh is (Rh includes the internal resistance of the voltage source V1):

$$R_h = \frac{V_{zv} \times R_{ex}}{0.25 \times V_{REF}}$$

with Vzv chosen to: $V_{zv} = 0.25 \times V_{REF} \rightarrow R_h = R_{ex}$

Where:	V_{ZV}	Voltage of the signal midpoint (signal zero voltage)	[V]
	V_{REF}	Voltage at the SVcc terminal (external or AVcc)	[V]
	Rex	Resistor between SVcc and Rext terminal (defines Ics)	[Ω]
	Rh	Shift resistor	[Ω]

The voltage V_{A1} at the analog input A1 is:

$$V_{A1} = V1 + Rh \times I_{CS} = V1 + Rh \times \frac{0.25 \times V_{REF}}{Rex}$$

The offset part (*Rh x Ics*) of the last equation is typically measured during a time when *V1* is known to be zero. This offset is stored in the RAM and subtracted from any measured value for *V1*. This leads to signed, offset corrected values for *V1.*

The unknown voltage *V1* is:

$$V1 = V_{A1} - Rh \times \frac{0.25 \times V_{REF}}{Rex} = V_{REF} \times \left(\frac{N}{2^{14}} - \frac{Rh \times 0.25}{Rex}\right)$$

With Rh=Rex: $V1 = V_{REF} \times \left(\frac{N}{2^{14}} - 0.25\right)$

Figure 12 gives two practical examples for dc and ac measurements using the current source. Both applications measure signed voltages that are partially (the negative parts) out of the ADC range of the MSP430.

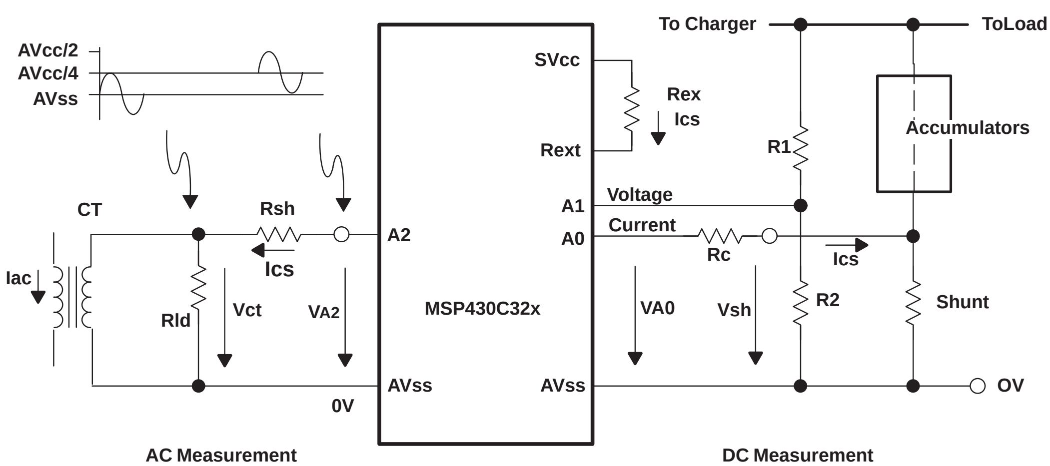

Figure 12. Signed Current Measurement With Level Shifting (Current Source)

AC Measurement: A current transformer CT is shown. Its output voltage is shifted into the ADC range by the current Ics of the current source and the resistor Rsh. The tolerable range for Ics is:

$$I_{CSmin} < I_{CS} < I_{dcmax}$$

Icsmin is defined by the ADC specification, and Idcmax is given by the current transformer specification. Current transformers normally are sensitive to dc bias currents. Rcu is the resistance of the transformer’s secondary winding (normally Rld >> Rcu).

$$V_{A2} = Vct + (Rsh + Rcu) \times I_{CS}$$

This leads to:

$$Vct = V_{A2} - Rsh \times I_{CS} = V_{REF} \times \left(\frac{N}{2^{14}} - \frac{0.25 \times (Rsh + Rcu)}{Rex} \right)$$

DC Measurement: The charge and discharge currents of an accumulator cause a voltage drop at the shunt resistor. This signed voltage drop *Vsh* is shifted into the ADC range by the resistor Rc (normally Rshunt << Rc).

$$V_{A0} = Vsh + Rc \times I_{CS}$$

This leads to:

$$Vsh = V_{A0} - Rc \times I_{CS} = V_{REF} \times \left(\frac{N}{2^{14}} - \frac{0.25 \times Rc}{Rex} \right)$$

2.2.4 Resistor Divider

If the input voltages are high – which means normally higher than 10 × V_{REF} – then, as shown in Figure 13, a simple resistor divider may be used for the level shift into the ADC range.

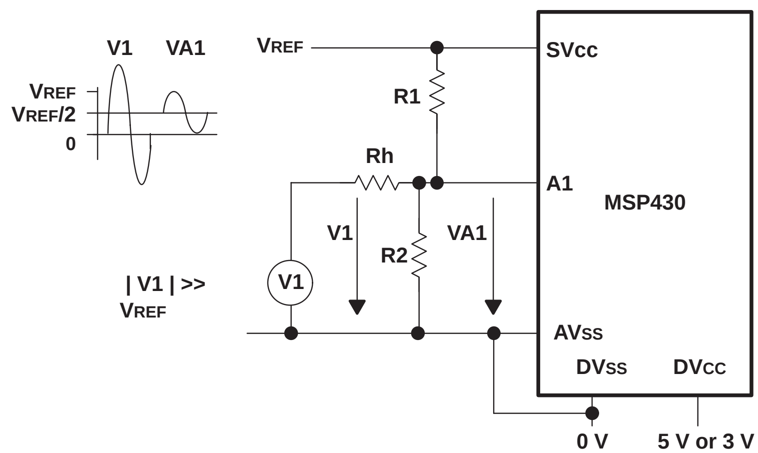

Figure 13. Resistor Divider for High Input Voltages

For input voltages V1 that are much higher than V_{REF}, the following equation is valid (Rh >> R2):

$$V_{A1} = V1 \times \frac{R1||R2}{R1||R2 + Rh} + V_{REF} \times \frac{R2}{R1 + R2} = \frac{N_{A1}}{2^{14}} \times V_{REF}$$

This leads to: $$V1 = V_{REF} \times \left(\frac{N_{A1}}{2^{14}} - \frac{R2}{R1 + R2} \right) \times \left(1 + \frac{Rh}{R1||R2} \right)$$

To get the full accuracy of the ADC, the condition R1||R2 < 27 kΩ must be fulfilled.

For high input voltages V1 the resistors R1 and R2 are normally equal—it is not possible or necessary to correct the small error of the input signal—so the equation simplifies to:

$$V_{A1} = V1 \times \frac{R1}{R1 + 2 \times Rh} + 0.5 \times V_{REF} = \frac{N_{A1}}{2^{14}} \times V_{REF}$$

This leads to: $$V1 = V_{REF} \times \left(\frac{N_{A1}}{2^{14}} - 0.5\right) \times \left(1 + \frac{2 \times Rh}{R1}\right)$$

The dc offset part ($0.5 \times V_{REF}$) of the last equation is typically measured during a time when V1 is known to be zero. This measured offset is stored in the RAM and subtracted from any measured value for V1. This leads to signed, offset corrected values for V1.

For input voltages that have no dc-part (e.g., sinusoidal signals), the zero point can be calculated by an integration of the input signal. After a multiple *m* of the signal period, the integrated sum of ADC results equals *m* times the value of the zero point.

2.3 12-Bit Analog-to-Digital Conversion With Signed Signals

The asymmetrical arrangement of the four ADC ranges reduces the number of solutions that are possible with the 12-bit conversion:

- Normal phase splitter circuits are not able to shift the virtual ground into the middle of range A, B C or D as it is necessary here. See Table 2 column Vvg for the center values of the four ADC ranges.
- The split power supply method would need two voltages to get the zero point into the center of the used range: e.g., 0.625 V and 4.375 V for range A if a 5-V supply is used.

NOTE: The formulas given in this section are valid only if both measurements for differences (ΔN) are measured in the same ADC range. If they are measured in different ADC ranges, then the 12–bit results need a correction (the missing two MSBs of the ADC result must be added). The correction numbers are:
Range A: 0
Range B: 1000h
Range C: 2000h
Range D: 3000h

2.3.1 Virtual Ground Circuitry

The phase splitter TLE2426 delivers only one half of the input voltage at its output terminal; it cannot be used here. With a simple op amp as shown in Figure 14, the necessary output voltages for the four ADC ranges can be obtained:
R = R1 + R2. See Table 1 for the relative resistor values.

Table 1. Resistor Ratios

ADC Range	Voltage VVG	R2	R1
A	0.125 × VREF	0.125 × R	0.875 × R
B	0.375 × VREF	0.375 × R	0.625 × R
C	0.625 × VREF	0.625 × R	0.375 × R
D	0.875 × VREF	0.875 × R	0.125 × R

Resistors R1 and R2 can have relatively high resistances. Only the offset current of the op amp limits these resistor values.

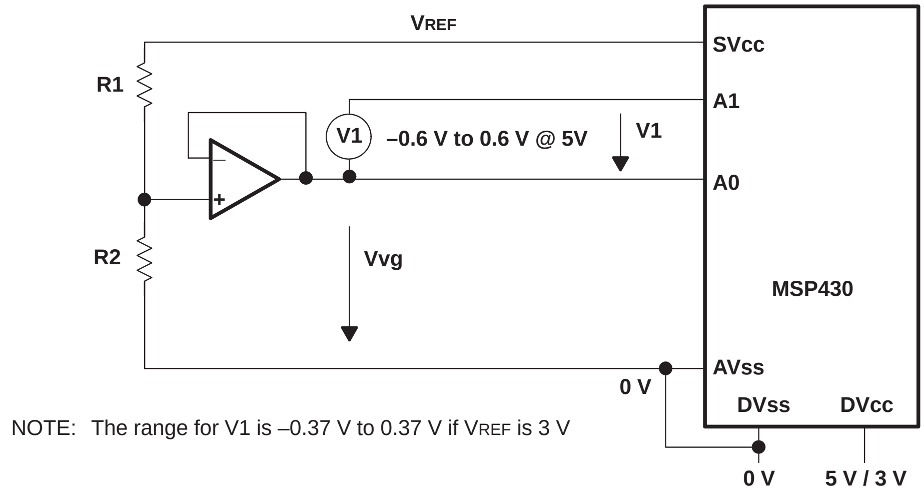

Figure 14. Virtual Ground Circuitry for Level Shifting

The formula for the difference of the ADC results ΔN measured at the analog inputs A1 and A0 is:

$$\Delta N = (N_{A1} - N_{A0}) = \frac{V1 + Vvg}{V_{REF}} \times 2^{14} - \frac{Vvg}{V_{REF}} \times 2^{14} = \frac{V1}{V_{REF}} \times 2^{14}$$

This leads to the formula for *V1*:

$$V1 = V_{REF} \times \frac{\Delta N}{2^{14}}$$

Where:	*V1*	Voltage to be measured inside of one ADC range	[V]
	ΔN	Difference of two ADC results (here N_{A1}–N_{A0})	
	V_{REF}	Voltage at the SVcc terminal measured against AVss terminal	[V]
	Vvg	Voltage at the A0 input (center of the used ADC range)	[V]

EXAMPLE: The center voltage of the C range (at analog input A0) is measured and stored in location VIRTGR (register or RAM). The value of VIRTGR is subtracted from the ADC value measured at analog input A1; this gives the signed, offset corrected value for the input signal at the A1 input. The measurement subroutine MEASR of section 4.1 is used.

```
; Measure center voltage of range C at analog input A0 and
; store value for reference. MCLK = 3.3MHz: divide MCLK by 3
;
        MOV     #ADCLK3+RNGC+CSOFF+A0+VREF,&ACTL
        CALL    #MEASR              ; Measure A0 (center voltage)
```

```
        MOV &ADAT,VIRTGR              ; Store result: 12-bit value
        ...
;
; Measure analog input signal V1 (0 ...0FFFh) and compute
; a signed, offset corrected value for V1 (0F800h...07FFh)
;
        MOV    #ADCLK3+RNGC+CSOFF+A1+VREF,&ACTL
        CALL   #MEASR                 ; Measure A1 (input voltage V1)
        MOV    &ADAT,R5               ; Read ADC value for V1
        SUB    VIRTGR,R5              ; R5 contains signed delta N
        ...                           ; V1 = Vref x deltaN x 2^-14
```

2.3.2 Use of the Current Source

For signed signals it is necessary to shift the input signal V1 to the center of the ranges A or B. See Figure 15.

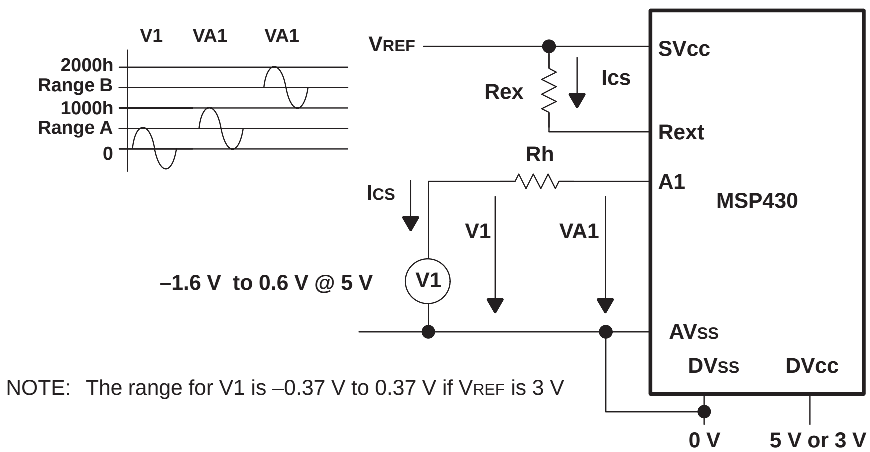

Figure 15. Current Source Used for Level Shifting

To get into the center of range *n* the necessary shift resistor *Rh* is:

$$Rh = 0.25 \times V_{REF} \times \frac{2n+1}{2} \times \frac{Rex}{0.25 \times V_{REF}} \rightarrow Rh = (n + 0.5) \times Rex$$

The unknown voltage *V1* measured to its zero point in the center of range *n* is:

$$V1 = V_{Ax} - Rh \times I_{CS}$$

With the above equation for *Rh* this leads to:

$$V1 = 0.25 \times V_{REF} \times \left(\frac{N}{2^{12}} + n - \frac{Rh}{Rex}\right)$$

2.3.3 Resistor Divider

The same circuitry is used as shown for the 14-bit conversion. See Figure 13. With the 12-bit conversion, it only makes sense to use the A range. This means for resistors R1 and R2, if R = R1 + R2:

$$R1 = 0.875 \times R \text{ and } R2 = 0.125 \times R.$$

For input voltages V1 that are much higher than V_{REF}, the following equation is valid (Rh >> R2):

$$V_{A1} = V1 \times \frac{R1||R2}{R1||R2 + Rh} + V_{REF} \times \frac{R2}{R1 + R2} = \frac{N_{A1}}{2^{14}} \times V_{REF}$$

With the above values for R1 and R2 this leads to:

$$V1 = V_{REF} \times 0.125 \times \left(\frac{N_{A1}}{2^{11}} - 1\right) \times \left(1 + \frac{Rh}{0.125 \times 0.875 \times R}\right)$$

To get the full accuracy of the ADC, the condition R1||R2 < 27 kΩ must be fulfilled. This means R < 247 kΩ.

2.4 Reference Resistor Method

A system that uses sensors normally needs to be calibrated, due to the tolerances of the sensors themselves and of the ADC. A way to omit this costly calibration procedure is the use of reference resistors. Two methods can be used, depending on the type of sensor:

1. Platinum sensors (e.g., PT500, PT100): These are sensors with a precisely known temperature/resistance characteristic. Two precision resistors are used with the sensor resistances of the temperatures at the two limits of the temperature range.
2. Other sensors: Nearly all other sensors have insufficiently tight tolerances. This makes it necessary to group sensors with similar characteristics, and to select the two reference resistors according to the sensor resistances at the upper and the lower measurement range limits of these groups.

If the two reference resistors have—within the needed accuracy—the values of the sensors at the measurement range limits (or at other well-defined points) then all tolerances are eliminated during the calculation. Therefore, no calibration is necessary.

> **NOTE:** For voltage measurements, the reference method described above can be used with two reference voltages instead of two resistors. In this case, substitute voltages for the resistances used with the next equations.

2.4.1 Reference Resistor Method Without Amplification

This method can be used for the input range given by the current source—the A and B ranges and part of range C. For details, see *Architecture and Function of the MSP430 14-Bit ADC Application Report.*[1]

The nominal formulas given in the previous section need to be modified if the tolerances of the ADC, the current source, the external components, and the sensor are considered. The ADC value *Nx* for a given resistor Rx is now:

$$Nx = \frac{Rx}{Rex} \times 2^{12} \times Slope + Offset$$

The slope and the offset are used for the correction of the measured result *Nx*. For the calculation of the slope and offset measurements with different resistors, Rx are necessary. With the hardware shown in figure 16 this calibration process can be omitted.

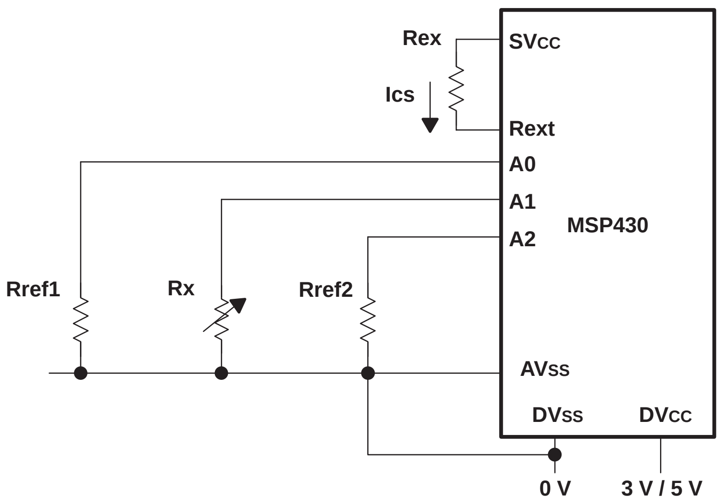

Figure 16. Referencing With Precision Resistors – No Amplification

With two known resistors Rref1 and Rref2 as shown in Figure 16, it is not necessary to know the slope and the offset to measure the value of the unknown resistor Rx exactly. Measurements are made for Rx, Rref1, and Rref2. The ADC results for these three measurements are:

$$Nx = \frac{Rx}{Rex} \times 2^{12} \qquad Nref1 = \frac{Rref1}{Rex} \times 2^{12} \qquad Nref2 = \frac{Rref2}{Rex} \times 2^{12}$$

The result of the solved equations shown above leads to:

$$Rx = \frac{Nx - Nref2}{Nref2 - Nref1} \times (Rref2 – Rref1) + Rref2$$

Where:	*Nx*	ADC conversion result for sensor Rx	
	Nref1	ADC conversion result for reference resistor Rref1	
	Nref2	ADC conversion result for reference resistor Rref2	
	Rref1	Resistance of Rref1 (equals Rxmin)	[Ω]
	Rref2	Resistance of Rref2 (equals Rxmax)	[Ω]

As shown, only known or measurable values are needed for the computation of Rx from Nx. Slope and offset influences of the ADC disappear completely:

- The offset disappears due to the two subtractions, one in the numerator and one in the denominator of the fraction above.

- The slope disappears due to the division

EXAMPLE: The values of these two reference resistors are chosen here for a PT1000 temperature sensor:

Rref1: 1000 Ω: The value of Rxmin. The resistance of a PT1000 sensor at 0°C (Tmin)

Rref2: 1380 Ω: The value of Rxmax. The resistance of a PT1000 sensor at 100°C (Tmax)

2.4.2 Reference Resistor Method With Amplification

If amplification is necessary to get a better resolution, then the solution shown below may be used. The full ADC range (0 to 3FFFh) can be used at analog input A1 despite the use of the current source at analog input A0. As with the section above, the offset and slope disappear; this is also true for the voltage drop at the outputs TP.x due to RDSon. The TP port of the measured resistor is switched to AVss potential; the other ones are set to Hi-Z.

The only error source of this arrangement is the difference of the internal resistances of the TP outputs (ΔRDSon). To minimize the influence of different internal resistances RDSon, only sensors with a minimum resistance should be used, e.g., PT1000 not PT100.

For the full 14-bit resolution at the analog input A1 the following design equations are valid (Rref2 > Rref1). They simplify this way if Rex is chosen to:

$$Rex = \frac{Rref2}{2}$$

This results in a maximum voltage of V_{REF}/2—the safe maximum output voltage the current source can deliver—at the analog input A0 for the maximum resistor value Rref2.

$$Vm = \frac{Rref1}{Rref1 + Rref2} \times V_{REF} \qquad v = \frac{V_{REF}}{V_{REF} - 2 \times Vm} = \frac{R1}{R2 || R3}$$

The calculated amplification *v* of the op amp needs to be reduced by 10 to 15% to be sure that V_{A1} does not saturate under worst case conditions.

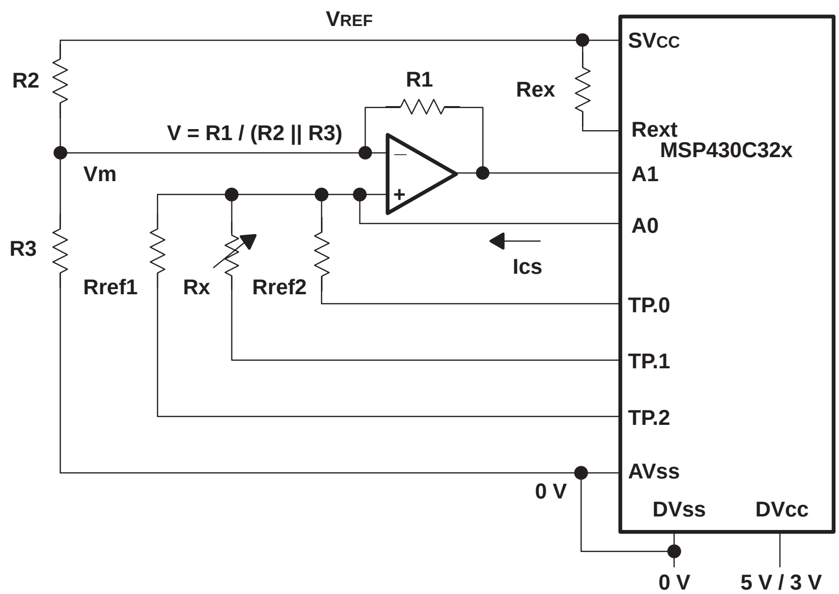

Figure 17. Referencing with Precision Resistors – With Amplification

As Figure 17 shows, with two known resistors Rref1 and Rref2 it is possible to get the values of unknown resistors exactly. The result of the solved equations gives:

$$Rx = \frac{\Delta Nx - \Delta Nref2}{\Delta Nref2 - \Delta Nref1} \times (Rref2 - Rref1) + Rref2$$

Where:	ΔNx	Difference of the two ADC results for Rx	$(N_{A1}–N_{A0})$
	$\Delta Nref1$	Difference of the two ADC results for Rref1	$(N_{A1}–N_{A0})$
	$\Delta Nref2$	Difference of the two ADC results for Rref2	$(N_{A1}–N_{A0})$
	Vm	Voltage generated by the resistor divider R2 and R3	

The differences named above are the differences between the ADC conversion results measured at the analog inputs A1 and A0 for each resistor: $\Delta N = N_{A1} – N_{A0}$.

3 Hum and Noise Considerations

3.1 Connection of Long Sensor Lines

If the distance from the MSP430 to the sensor is long (>30 cm) then it is recommended to use a shielded cable between the microcomputer and the sensor. This avoids spikes at the ADC input that cause measurement errors, and also gives protection to the ADC input. Figure 18 shows this schematic on the left side. In the same way, four-wire circuitry may be connected to the MSP430.

If a shielded cable cannot be used, the circuitry shown on the right side of Figure 18 should be used; the AVss line in parallel to the signal line gives a relatively good screening. Twisting the two lines increases the protection.

To protect the measurement against spikes, hum, and other unwanted noise see the *Signal Averaging and Noise Cancellation* section of Chapter 5 in the *MSP430 Application Report*.[2] This section shows additional possibilities for the minimization of these influences by software.

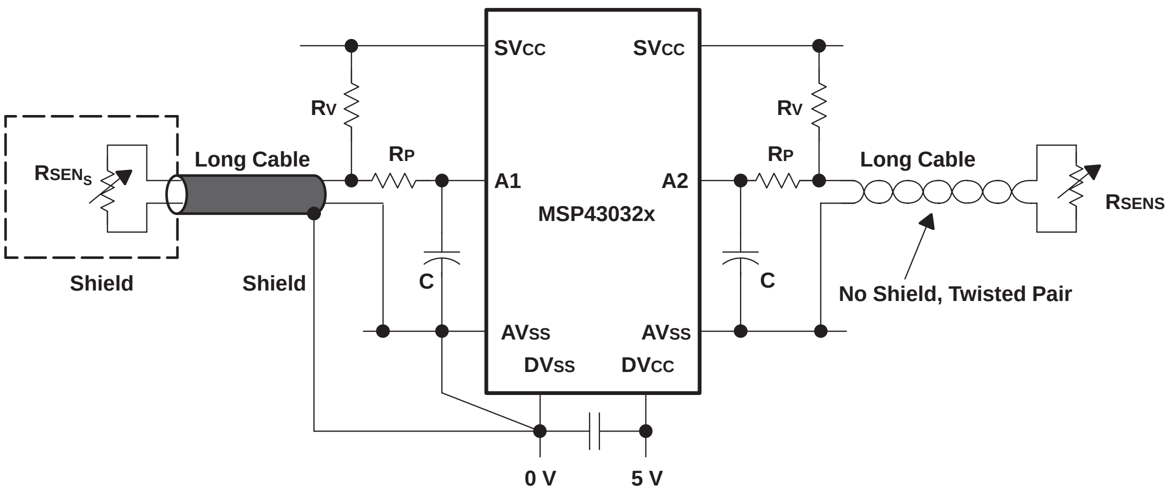

Figure 18. Sensor Connection via Long Cables With Voltage Supply

With the circuitry of figure 18, the minimum time tdelay between the switch-on of the voltage SVcc and the actual measurement—to get the full 14-bit accuracy—is:

$$t_{delay} > ln2^{14} \times \tau_{max} = 9.704 \times \tau_{max} \approx 10 \times \tau_{max}$$

The value of τ_{max} is:

$$\tau_{max} = (Rp + Rsens_{max} || Rv) \times C$$

If the current source is used, then:

$$Rv = \infty : \quad \tau_{max} = (Rp + Rsens_{max}) \times C$$

3.2 Grounding

Correct grounding is very important for ADCs with high resolution. There are some basic rules that need to be observed[1]. See Figure 19 also.

1. Use a separate analog and digital ground plane wherever possible: thin traces from the battery to terminals DVss and AVss should be avoided.
2. The AVss terminal should serve as a star point for all analog ground connections e.g. sensors, analog input signals. The DVss terminal should serve as a star point for all digital ground connections e.g. switches, keys, power transistors, output lines, digital input signals.
3. The battery and storage capacitor Cb should be connected close together (the capacitor Cb is needed for batteries with a relatively high internal resistance). From this capacitor two different paths go to the analog and the digital supply terminals. Two small capacitors are connected across the digital (Cd) and the analog (Ca) supply terminals. See Figure 19.
4. Rules 1 to 3 above are also true for the Vcc paths (DVCC and AVCC).
5. The AVss and DVss terminals must be connected together externally; they are not connected internally. The same is true for the AVcc and DVcc terminals. These connections should be made with the configuration shown in Figure 19.
6. The coil L should be used in very difficult cases.
7. The connections of the capacitor Cb are the star point of the complete system. This is due to the low impedance of this capacitor.

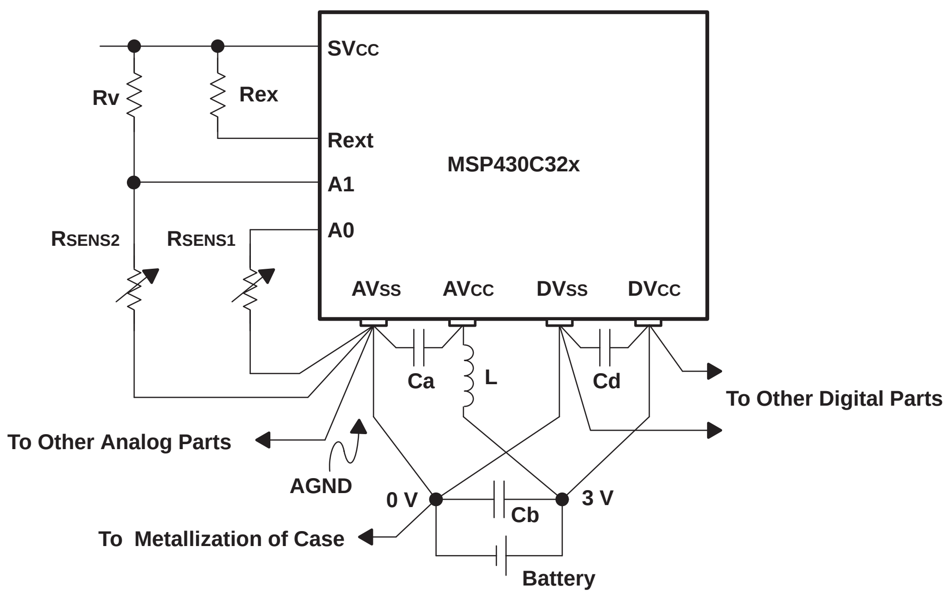

Figure 19. Analog-to-Digital Converter Grounding

If a metalized case is used around the printed circuit board containing the MSP430 then it is very important to connect the metallization to the ground potential (0 V) of the board. Otherwise the behavior is worse than without the metalization.

[1] These grounding rules were developed by E. Haseloff of TID.

3.3 Routing

Correct routing for a PC board is very important for minimum noise. Figure 20 shows a simplified routing that is not optimal; the gray areas receive EMI from external sources. For a minimum influence coming from external sources these areas must be as small as possible.

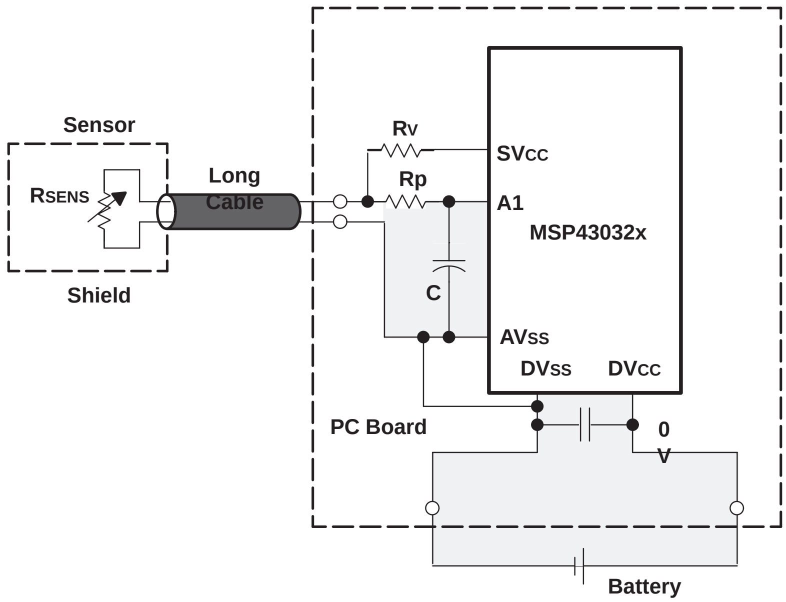

Figure 20. Routing That is Sensitive to External EMI

Figure 21 shows an optimized routing; the areas that may fetch noise have a minimum size.

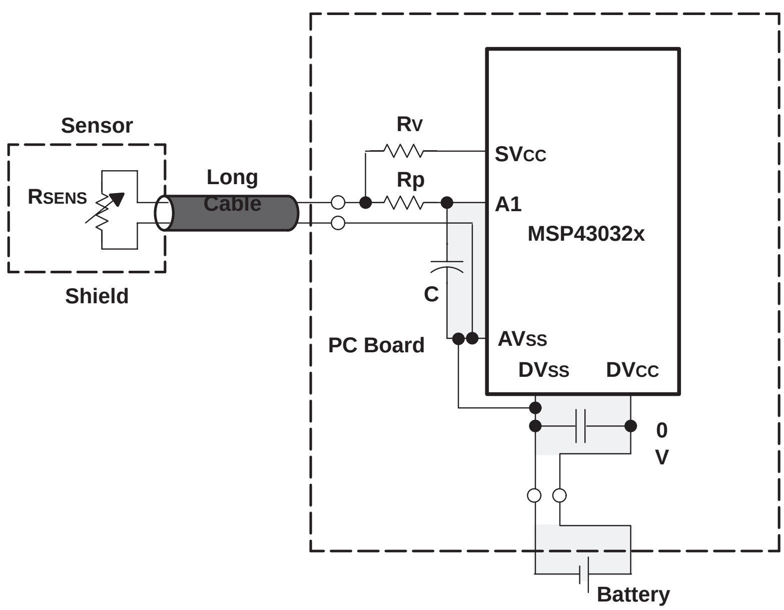

Figure 21. Routing for Minimum EMI Sensitivity

4 Enhancement of the Resolution

Many applications need a higher resolution than the 14-bit ADC can provide. For these applications the following hints may be helpful.

NOTE: These enhancements make it necessary to pay attention to the rules given in Chapter 3. Without observing these rules strictly, no enhancement will be seen.

4.1 16-Bit Mode With the Current Source

With the use of two additional output terminals (I/O-ports or TP-outputs) the 14-bit ADC may be expanded to a resolution of nearly 16 bits. The principle is simple: the resistor Rex of the current source is modified by paralleling two additional resistors (see Figure 23). These resistors have values that represent one half and one quarter of a single ADC-step. Due to the fact that these fractions of a step are accurate only at one point of the ADC-range, this enhancement gives only better resolution, not better accuracy. To get the 16-bit result, four measurements are necessary: one for every combination of the two additional resistors. If the results of these four measurements are added, a 16-bit result is reached. See Figure 22.

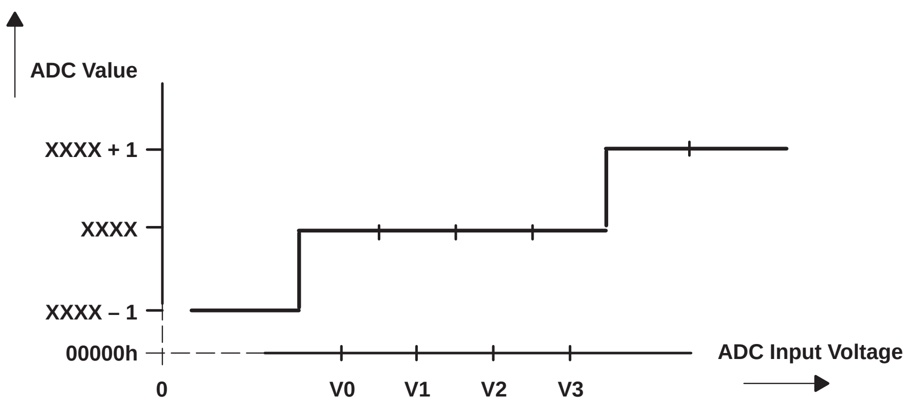

Figure 22. Dividing of an ADC-Step Into Four Steps

Table 2 shows the different results of these four measurements for the four possible input voltages V0 to V3 inside of one ADC-step; the table refers to the hardware shown in Figure 23.

Table 2. Measurement Results of the 16-Bit Method

INPUT VOLTAGE	MEASUREMENT 1 TP.1: Hi-Z TP.0: Hi-Z	MEASUREMENT 2 TP.1: Hi-Z TP.0: Hi OUT	MEASUREMENT 3 TP.1: Hi OUT TP.0: Hi-Z	MEASUREMENT 4 TP.1: Hi OUT TP.0: Hi OUT	MEAN VALUE (BINARY)
V0	XXXX	XXXX	XXXX	XXXX	XXXX.00
V1	XXXX	XXXX	XXXX	XXXX+1	xxxx.01
V2	XXXX	XXXX	XXXX+1	XXXX+1	XXXX.10
V3	XXXX	XXXX+1	XXXX+1	XXXX+1	XXXX.11

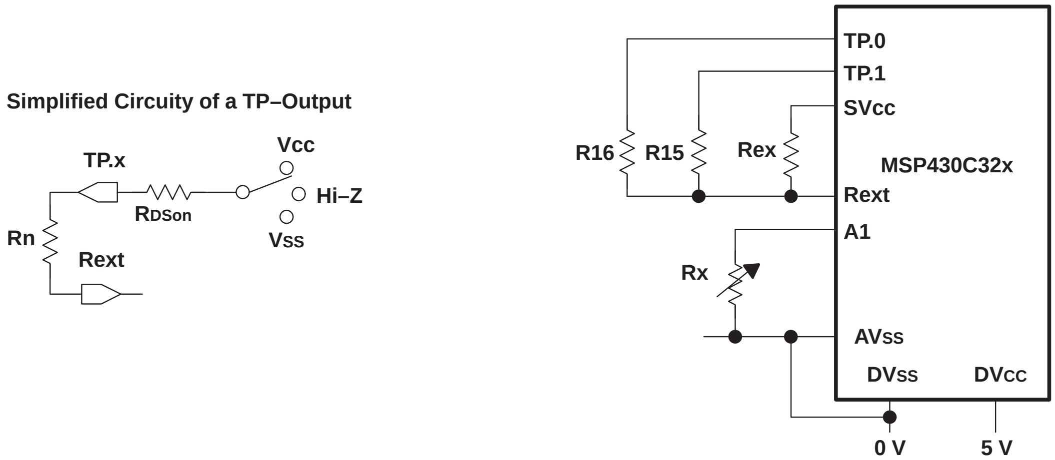

Figure 23. Hardware for a 16-Bit ADC

The values for resistors R16 and R15 are:

$$Rp = \frac{2^{14} \times 0.25 \times Rx0}{m} = \frac{2^{12} \times Rx0}{m}$$

Where:	*Rp*	Parallel resistor to Rex (here R14 and R15)	[Ω]
	Rx0	Sensor resistance at the point of the highest accuracy	[Ω]
	m	Fraction of an ADC step (0.25 or 0.5)	

EXAMPLE: With the hardware shown in Figure 23, four 16-bit measurements are made. The result is placed into R5. The software may also be written with a loop. The software assumes ascending order for the two TP outputs.

```
        MOV     #RNGAUTO+CSA1+A1+VREF,&ACTL   ; Define ADC
        BIC.B   #TP1+TP0,&TPE                 ; TP.0 and TP.1 to Hi–Z
        BIS.B   #TP1+TP0,&TPD                 ; Set TPD.0 and TPD.1 to Hi
        CALL    #MEASR                        ; Measure with R15 = R16 =
                                              ; Hi–Z
        MOV     &ADAT,R5                      ; 14–bit value to result
        ADD.B   #TP0,&TPE                     ; Set R16 to Hi–Out
        CALL    #MEASR                        ; Measure
        ADD     &ADAT,R5                      ; Add 14–bit value to
                                              ; result
        ADD.B   #TP0,&TPE                     ; Set R15 to Hi–Out,R16 to
                                              ; Hi–Z
        CALL    #MEASR                        ; Measure
        ADD     &ADAT,R5                      ; Add 14–bit value to
                                              ; result
        ADD.B   #TP0,&TPE                     ; Set R15 and R16 to Hi–Out
        CALL    #MEASR                        ; Measure
        ADD     &ADAT,R5                      ; Add 14–bit value to
                                              ; result
        BIC.B   #TP1+TP0,&TPE                 ; TP.n off
        ...                                   ; 16–Bit result 4N in R5
;
; The measurement routine used above:
;
MEASR   BIC.B   #ADIFG,&IFG2              ; Clear EOC flag
        ...                               ; Insert delays here (NOPs)
        BIS     #SOC,&ACTL                ; Start measurement
M0      BIT.B   #ADIFG,&IFG2              ; Conversion completed?
        JZ      M0                        ; No
        RET                               ; Result in ADAT
```

4.2 Enhanced Resolution Without Current Source

The principle is explained in the last section. Figure 24 shows a hardware proposal for the measurement part of a scale using the MSP430C32x. With the resistor Rn, the resolution of the MSP430 ADC is increased to 15 bits:

- TP.0 is off (Hi-Z): normal measurement
- TP.0 is switched to V_{CC}: the current into the right bridge leg increases the voltage at A1 by 0.5 steps of the ADC

Two differential ADC measurements ($N_{A0} – N_{A1}$)—one with TP.0 off and one with TP.0 switched to V_{CC}—are summed-up and provide (nearly) 15-bit resolution. The result of these four measurements is $2 \times \Delta N$.

The formulas derived in the Connection of Bridge Assemblies section are valid here as well.

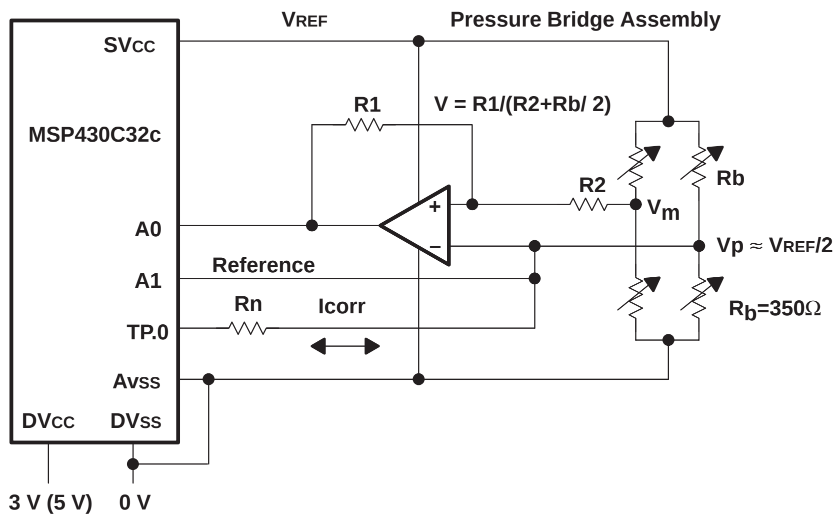

Figure 24. ADC-Resolution Expanded to 15 Bits

The formula for Rn to cause a voltage difference ΔV_{A0} (here 0.5 ADC steps) at the ADC input is:

$$\Delta V_{A0} = \frac{V_{REF}}{2^{15}} = \frac{Rb}{2 \times Rn + Rb} \times \left(V_{CC} - \frac{V_{REF}}{2} \right) \times v$$

This gives an approximate value for Rn ($V_{CC} = V_{REF}$):

$$Rn \approx Rb \times 2^{13} \times \frac{R1}{R2}$$

Where:	ΔV_{A0}	Change of the input voltage at input A0 due to Rn	[V]
	Rb	Resistance of a half bridge leg (here 350 Ω)	[Ω]
	Rn	Resistance of the resistor for 15 bits resolution	[Ω]
	v	Amplification of the operational amplifier: v = R1/R2	
	V_{REF}	Supply voltage at the SVcc terminal (int. or ext.)	[V]
	DV_{CC}	Supply voltage at DVcc terminal (output voltage of TP.0)	[V]
	R1,R2	Resistors defining the amplification of the op amp	

Without any change to the hardware above, the resolution of the ADC can be increased to 15.5 bits (this method is only possible with sensor assemblies like those shown in Figure 24, that deliver output voltages near 0.5 x Vref):

- TP.0 is off (Hi-Z): normal measurement
- TP.0 is switched to Vcc: the current into the right bridge leg increases the voltage at A1 by 0.5 steps of the ADC
- TP.0 is switched to Vss: the current out of the right bridge leg decreases the voltage at A1 by 0.5 steps of the ADC

Three differential ADC measurements ($N_{A0} - N_{A1}$)—one with TP.0 switched to Hi-Z, one with TP.0 switched to Vss, and one switched to V_{CC}—are summed-up and provide (nearly) 15.5-bit resolution. The calculations following these six measurements must be changed for an input value of $3 \times N$.

ΔV_{A0} is chosen to: $\Delta V_{A0} = \dfrac{V_{REF}}{3 \times 2^{14}}$

The circuitry of Figure 24 leads to very high values of Rn with high amplifications v: for the above example Rn = 286 MΩ for v = 100. If these resistor values are too high, then the circuitry shown in Figure 25 should be used. The registor values of R15 and R16 have the same effect as the circuitry in Figure 24, but are much smaller.

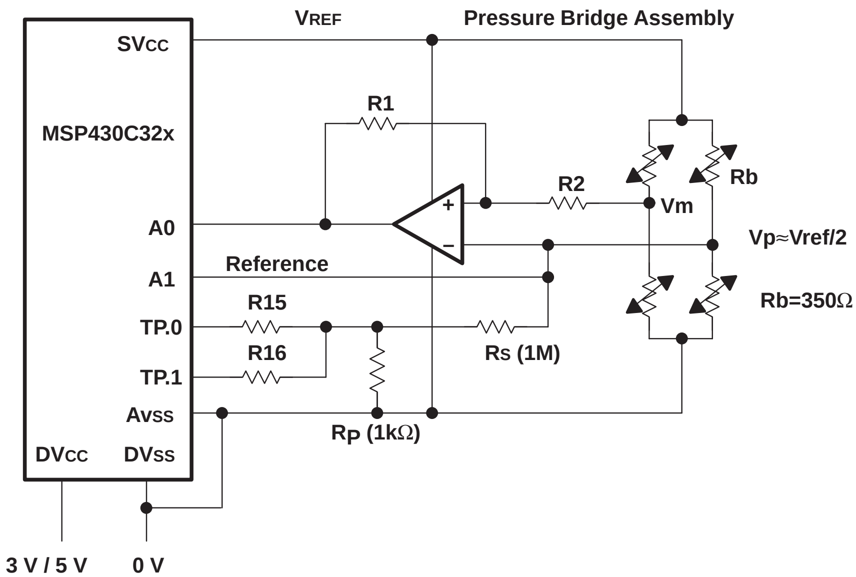

Figure 25. ADC-Resolution Expanded to 16 Bits

The formula for R15 and R16 to cause a voltage difference ΔV_{A0} of 0.5 ADC steps for R15 and 0.25 ADC steps for R16 at an ADC input is now:

$$Rn \approx 2^n \times \frac{R1}{R2} \times \frac{Rp}{Rs} \times \frac{Rb}{2}$$

Where: *n* Bit number of resolution resistor (15 or 16)
Rn Resistance of resolution resistor (bit n) [Ω]
Rb Resistance of a half bridge leg (here 350 Ω) [Ω]
Rp Parallel resistor (chosen to be 1k: small compared to 1 M) [Ω]
Rs Serial resistor (chosen to be 1M: large compared to 350 Ω) [Ω]

With the circuitry of Figure 25 (v = 100) R15 now becomes 573 k and R16 becomes 1.15M.

The necessary four measurements are described in Table 2. Each measurement consists of two ADC conversions that are subtracted afterwards ($\Delta N = N_{A0} - N_{A1}$). The four differences ΔN are summed and deliver a 16-bit result with nearly two bits more resolution than the normal 14-bit result. The result in R5 is $4 \times \Delta N$.

EXAMPLE: With the hardware shown in Figure 25, four differential measurements for ΔN are made ($\Delta N = N_{A0} - N_{A1}$). The four values for ΔN are summed in R5. The software assumes ascending order for the two TP outputs (TP.x and TP.x+1).

```
       BIC.B     #TP1+TP0,&TPE           ; TP.0 and TP.1 to Hi–Z
       BIS.B     #TP1+TP0,&TPD           ; Set TPD.0 and TPD.1 to Hi
       MOV       #RNGAUTO+A0+VREF,&ACTL  ; Define ADC
       CALL      #MEASR                  ; Measure with R15 = R16 = Hi–Z
       MOV       &ADAT,R5                ; 14–bit value to result
       MOV       #RNGAUTO+A1+VREF,&ACTL  ; Define ADC
       CALL      #MEASR                  ; Measure with R15 = R16 = Hi–Z
       SUB       &ADAT,R5                ; (NA0 – NA1) to result
;
       ADD.B     #TP0,&TPE               ; Set R16 to Hi–Out, R15 = Hi–Z
       MOV       #RNGAUTO+A0+VREF,&ACTL  ; Define ADC
       CALL      #MEASR                  ; Measure
       ADD       &ADAT,R5                ; Add 14–bit value to result
       MOV       #RNGAUTO+A1+VREF,&ACTL  ; Define ADC
       CALL      #MEASR                  ; Measure
       SUB       &ADAT,R5                ; (NA0 – NA1) to result
;
       ADD.B     #TP0,&TPE               ; Set R15 to Hi–Out,R16 to Hi–Z
       MOV       #RNGAUTO+A0+VREF,&ACTL  ; Define ADC
       CALL      #MEASR                  ; Measure
       ADD       &ADAT,R5                ; Add 14–bit value to result
       MOV       #RNGAUTO+A1+VREF,&ACTL  ; Define ADC
       CALL      #MEASR                  ; Measure
       SUB       &ADAT,R5                ; (NA0 – NA1) to result
;
       ADD.B     #TP0,&TPE               ; Set R15 and R16 to Hi–Out
       MOV       #RNGAUTO+A0+VREF,&ACTL  ; Define ADC
       CALL      #MEASR                  ; Measure
       ADD       &ADAT,R5                ; Add 14–bit value to result
       MOV       #RNGAUTO+A1+VREF,&ACTL  ; Define ADC
       CALL      #MEASR                  ; Measure
       SUB       &ADAT,R5                ; (NA0 – NA1) to result
;
       BIC.B     #TP1+TP0,&TPE           ; TP.n off
       ...                               ; 16–Bit result 4xN in R5
;
```

4.3 Calculated Resolution of the 16-Bit Mode

4.3.1 16-Bit Mode With the Current Source

To give an idea of how much better the results of the 16-bit mode can be compared to the 14-bit mode of the ADC, the results of four calculations are shown in Table 3. The table shows the statistical results for the deviations of the corrected result in ADC-steps:

- The first column shows the statistical results for the normal 14-bit ADC
- The second column shows the statistical results for measurements that have the highest accuracy at the lowest sensor value: Rx0 = 1000 Ω
- The third column shows the statistical values if the point of highest accuracy is moved to the midpoint of the sensor resistance: Rx0 = 1190 Ω
- The fourth column shows the same as before if the highest sensor value is used for the highest accuracy: Rx0 = 1380 Ω

Calculation values and explanations:

Rxmax:	1380.0 Ω	Highest sensor resistance (100°C for PT1000)
Rxmin:	1000.0 Ω	Lowest sensor resistance (0°C for PT1000)
Rx0:		Sensor resistance for highest accuracy (3 different values)
ΔRx:	0.01 Ω	Step width for resistance value during calculation
Rex:	690.0 Ω	Calculated external resistor for the Current Source
R15:		Calculated resistor for the 15th bit
R16:		Calculated resistor for the 16th bit

Table 3. Calculation Results for Different 16-Bit Corrections

ITEM	NO CORRECTION 14-BIT	Rx0 = 1000 Ω 16-BIT	Rx0 = 1190 Ω 16-BIT	Rx0 = 1380 Ω 16-BIT
R15	N/A	8.2MΩ	9.7MΩ	11.3MΩ
R16	N/A	16.4MΩ	19.5MΩ	22.6MΩ
Mean value	–0.5001	–0.0538	–0.1250	–0.1767
Standard deviation	0.2887	0.1019	0.0841	0.0898
Variance	0.0833	0.0104	0.0071	0.0081

Table 3 shows the improved resolution especially if the best resolution is programmed for the lowest sensor resistance (Rx0 = 1000 Ω). The result is derived from 38,000 measurements with a step width of 0.01 Ω. The 14-bit results show the (correct) inherent error of minus 0.5 steps that is enhanced with the three 16-bit modes by a factor of 3 to 9.

4.3.2 16-Bit Mode Without the Current Source

Circuitry like shown in Figure 25 is normally used: this means the input voltage of the analog inputs is always near 0.5 × Vref. Therefore the results of Table 3 column Rx0 =1190 Ω (highest accuracy at the center of the resistance range) are valid.

5 Hints and Recommendations

5.1 Replacement of the First Measurement

In certain cases the first measurement is discarded. Instead, a second measurement is started and used. This method is especially useful if the settling time for the ADC is insufficient.

```
MOV      #XX,&ACTL             ; Define ADC
CALL     #MEASR                ; 1st measurement (not used)
CALL     #MEASR                ; 2nd measurement is used
MOV      &ADAT,R5              ; for calculations. Result to R5
```

5.2 Grounding and Routing

With increasing ADC accuracy and CPU frequency, the board layout becomes more important. A few hints may help to increase the performance of the ADC:

- To avoid cross talk from one ADC input line to the other one, grounded lines (AV_{SS} potential) between the analog input lines are recommended.
- Large ground planes (0V potential) should be used wherever possible. Any free space on the board should be used for this purpose.
- Analog input lines should be as short as possible. If this is not possible, input filtering may be necessary. See the *MSP430 Application Report[2]* for details.
- To get reliable ADC results in noisy environments, additional hardware and software filtering should be used. The *MSP430 Application Report[2]* describes several methods to do this in the *Signal Averaging and Noise Cancellation* section of Chapter 5: over sampling, continuous averaging, weighted summation, rejection of extremes, and synchronization to hum. Tested software examples are included.

See also sections 3.2 and 3.3.

5.3 Supply Voltage and Current

Completely different environments exist for battery and mains driven systems. A few hints are given for these two supplies. More information concerning this topic is included in the *MSP430 Application Report[2]* in the *Power Supplies for the MSP430* section of Chapter 3.

5.3.1 Influence of the Supply Voltage

The supply voltage is used for reference purposes if the Vref-bit (ACTL.1) is set. This means a change of the analog supply voltage AVcc during the measurement influences the final ADC result. The same is true for an external reference voltage.

Figure 26 shows a decreasing analog supply voltage with the ADC timing. The error of the ADC result N is mainly introduced during the conversion time for the 12 LSBs of the ADC result. The input sample is taken with an AV_{CC} voltage Vref0, the LSB is generated with an AV_{CC} voltage Vref1. The two results have the ratio:

$$\frac{N_1}{N_0} = \frac{Vref0}{Vref1}$$

The maximum error emax in per cent is therefore:

$$e_{max} = \frac{N_1 - N_0}{N_0} \times 100 = \left(\frac{Vref0}{Vref1} - 1\right) \times 100$$

Where:

	N_0	ADC result measured with a stable AV_{CC} of Vref0	
	N_1	ADC result measured with a stable AV_{CC} of Vref1	
	e_{max}	Maximum error caused by unstable AV_{CC}	[%]
	Vref0	Value of AV_{CC} during the sampling of the conversion sample	[V]
	Vref1	Value of AV_{CC} at the end of conversion	[V]

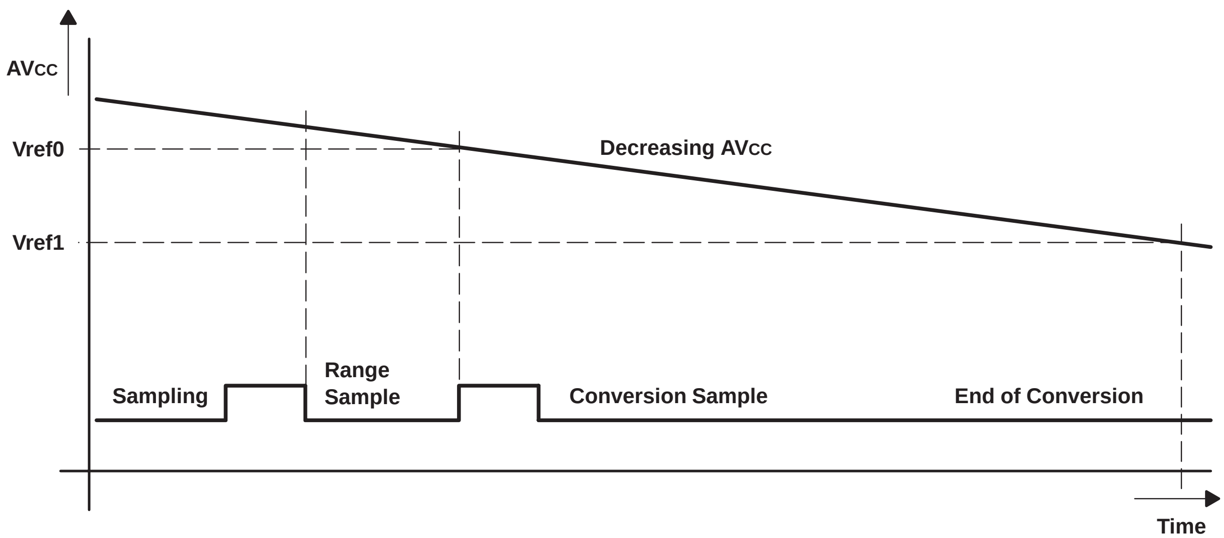

Figure 26. Influence of the Supply Voltage

The result caused by an unstable AV_{CC} can normally be detected by its trailing series of zeroes or ones. If, during the conversion, one of the leading bits is set, or reset, and this bit has the wrong state for the changing reference voltage, then, all remaining bits will have the same value, e.g., 1 for a decreasing AVcc.

5.3.2 Battery Driven Systems

If the battery used has a high internal resistance Ri (like some long-life batteries) then the parallel capacitor Cb (see Figure 19) must have a minimum capacity Cbmin: the supply current for the measurement part—which cannot be delivered by the battery—is delivered mainly by Cb; the approximate equation includes the small current coming from the battery:

$$Cb_{min} \geq tmeas \times \frac{I_{AM}}{\Delta Vb} - \frac{1}{Ri}$$

If the battery has a high impedance Ri, then it is recommended to use the kind of measurement shown in *Architecture and Function of the MSP430 14-Bit ADC Application Report*:[1] the CPU is switched off during the ADC measurement which lowers the current out of the battery.

Between two ADC measurements, the capacitor Cb needs a time tch to become charged to Vcc potential for the next measurement. During this charge-up time the MSP430 system runs in low power mode 3 to have the lowest possible power consumption. The charge time tch to charge Cb to 99% of Vcc is:

$$tch_{min} \geq 5 \times Cb_{max} \times Ri_{max}$$

Where:	Cb	Capacitor in parallel to the battery	[F]
	I_{AM}	Medium system current (MSP430 and ADC)	[A]
	$tmeas$	Discharge time of Cb during measurement	[s]
	ΔVb	Tolerable discharge voltage of Cb during time tmeas	[V]
	Ri	Internal resistance (impedance) of the battery	[Ω]
	tch	Charge-up time for the capacitor Cb	[s]

5.3.3 Mains Driven Systems

No hum, noise, or spikes are allowed for the supply voltages AVcc and DVcc. If present, the reliability of the system and the accuracy of the ADC will decrease. This is especially true for applications where the AVcc voltage is used for the ADC reference [ACTL.1 = 1 (Vref bit)]. See section *Averaging and Noise Cancellation* of chapter 5 of the MSP430 Application Report[2] for ways to overcome this problem.

5.3.4 Current Consumption

Often it is important to know the current consumption of the complete MSP430 system—which means including the supply current of the MSP430 and its ADC. The supply current of the CPU increases nearly linearly with the MCLK frequency and the applied supply voltage DVCC, but this is not the case for the ADC: the main component of the ADC supply current is drawn by the resistor divider with its 4×128 resistors. An approximate formula for the nominal current consumption Icc of the MSP430C32x is (internal ADC reference):

$$I_{CC} = I_{CCdigital} + I_{CCanalog} = \left(\frac{V_{DVcc}}{5V} \times \frac{f_{MCLK}}{1MHz} \times 750\ \mu A\right) + \left(\frac{V_{AVcc}}{3V} \times 200\ \mu A\right)$$

Where:	I_{CC}	Complete current consumption of MSP430 (nominal)	[µA]
	$I_{CCdigital}$	Current consumption of the digital parts	[µA]
	$I_{CCanalog}$	Current consumption of the ADC	[µA]
	V_{DVcc}	Voltage at the DVCC terminal	[V]
	V_{AVcc}	Voltage at the AVCC	[V]
	f_{MCLK}	Frequency of the system clock generator (MCLK)	[Hz]

5.4 Use of the Floating Point Package

For the MSP430 a Floating Point Package exists with two selectable bit lengths: 32 bit and 48 bit. For calculations with the ADC, results consisting of several multiplications and divisions, it is recommend to use this package: no decrease of accuracy is caused by the calculation itself. A detailed description of the Floating Point Package and all available mathematical functions is given in the *MSP430 Application Report*. See section *The Floating Point Package* section of Chapter 5.

A small example is given below: the measured ADC result—in ADC buffer ADAT—is corrected with slope and offset. The result (BCD format) is placed into the locations BCDMSD, BCDMID and BCDLSD (RAM or registers).

```
DOUBLE    .EQU      0                   ; Use .FLOAT format (32 bits)
;
          MOV       #xxx,&ACTL          ; Define ADC measurement
          CALL      #MEASR              ; Measure. Result to ADAT
```

```
          CALL      #FLT_SAV            ; Save registers R5 to R12
          SUB       #4,SP               ; Allocate stack for FP result
          MOV       #ADAT,RPARG         ; Load address of ADC buffer
          CALL      #CNV_BIN16U         ; Convert ADC result to FP
;
; Calculate: ADCcorr = (ADC result x Slope) + Offset
;
          MOV       #Slope,RPARG        ; Load address of slope
          CALL      #FLT_MUL            ; ADC result x Slope
          MOV       #Offset,RPARG       ; Load address of offset
          CALL      #FLT_ADD            ; ADC result x Slope + Offset
          ...                           ; Continue with calculations
;
; The final result is converted to BCD format for the display
;
          CALL      #CNV_FP_BCD         ; Convert FP result to BCD
          JN        CNVERR              ; Result too big for BCD buffer
          POP       BCDMSD              ; BCD number: sign and MSDs
          POP       BCDMID              ; BCD digits MSD–4 to LSD+4
          POP       BCDLSD              ; BCD digits LSD+3 to LSD
                                        ; Stack is corrected by POPs
          CALL      #FLT_REC            ; Restore registers R12 to R5
          ...                           ; Continue with program

Slope     .FLOAT    –1.2345             ; Slope (fixed, RAM, EEPROM)
Offset    .FLOAT    14.4567             ; Offset (fixed, RAM, EEPROM)
CNVERR    ...                           ; Start error handler
```

6 Additional Information

This application report is complemented by the *Additive Improvement of the MSP430 14-Bit ADC Characteristic* application report[5] that explains several methods to minimize the error of the 14-Bit ADC. For all methods (linear, quadratic, cubic and others) the actual improvement for a measured ADC characteristic is shown. The enhancement methods discussed are compared completely with statistic results, advantages and disadvantages, necessary CPU cycles, and storage needs.

7 References

1. *Architecture and Function of the MSP430 14-Bit ADC Application Report*, 1999, Literature #SLAA045
2. *MSP430 Application Report*, 1998, Literature #SLAAE10C
3. *MSP430 Family Architecture Guide and Module Library*, 1996, Literature #SLAUE10B
4. *MSP430C325, MSP430P323 Data Sheet*, 1999, Literature #SLAS219
5. *Additive Improvement of the MSP430 14-Bit ADC Characteristic Application Report*, 1999, Literature #SLAA047
6. *Linear Improvement of the MSP430 14-Bit ADC Characteristic Application Report*, 1999, Literature #SLAA048
7. *Nonlinear Improvement of the MSP430 14-Bit ADC Characteristic Application Report*, 1999, Literature #SLAA050

Appendix A Definitions Used With the Application Examples

```
; HARDWARE DEFINITIONS
;
AIN       .equ   0110h     ; Input register (for digital inputs)
AEN       .equ   0112h     ; 0: analog input   1: digital input
;
ACTL      .equ   0114h     ; ADC control register: control bits
SOC       .equ   01h       ; Conversion start
VREF      .equ   02h       ; 0: ext. reference    1: SVcc on
A0        .equ   00h       ; Input A0
A1        .equ   04h       ; Input A1
A2        .equ   08h       ; Input A2
A3        .equ   0Ch       ; Input A3
A4        .equ   10h       ; Input A4
A5        .equ   14h       ; Input A5
CSA0      .equ   00h       ; Current Source to A0
CSA1      .equ   40h       ; Current Source to A1
CSA2      .equ   80h       ; Current Source to A2
CSA3      .equ   0C0h      ; Current Source to A3
CSOFF     .equ   100h      ; Current Source off
CSON      .equ   000h      ; Current Source on
RNGA      .equ   000h      ; Range select A (0 ... 0.25xSVcc)
RNGB      .equ   200h      ; Range select B (0.25..0.50xSVcc)
RNGC      .equ   400h      ; Range select C (0.5...0.75xSVcc)
RNGD      .equ   600h      ; Range select D (0.75..SVcc)
RNGAUTO   .equ   800h      ; 1: range selected automatically
PD        .equ   1000h     ; 1: ADC powered down
ADCLK1    .equ   0000h     ; ADCLK = MCLK
ADCLK2    .equ   2000h     ; ADCLK = MCLK/2
ADCLK3    .equ   4000h     ; ADCLK = MCLK/3
ADCLK4    .equ   6000h     ; ADCLK = MCLK/4
;
ADAT      .equ   0118h     ; ADC data register (12 or 14–bit)
;
IFG2      .equ   03h       ; Interrupt flag register 2
ADIFG     .equ   04h       ; ADC ”EOC” bit (IFG2.2)
;
IE2       .equ   01h       ; Interrupt enable register 2
ADIE      .equ   04h       ; ADC interrupt enable bit (IE2.2)
;
TPD       .equ   04Eh      ; TP–port: address data register
TPE       .equ   04Fh      ; TP–port: address of enable register
TP0       .equ   1         ; Bit address of TP.0
TP1       .equ   2         ; Bit address of TP.1
```