AN1291

Low-Cost Shunt Power Meter using MCP3909 and PIC18F25K20

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OVERVIEW

This application note shows a single-phase energy meter solution using the MCP3909 Dual Channel ADC and the PIC18F25K20 8-bit microcontroller. This application note uses Microchip's reference design *"Low-Cost Power Monitor using MCP3909 and PIC18F25K20"*, Part Number: MCP3909RD-PM1.

For cost reasons, the current sensor is a 200 µΩ shunt. To improve the low current measurements, the value can be increased to a larger resistance and the meter would require calibration at this lower current. The shunt used in this application note is a model that is connected to the circuit through screws on the meter case. The meter PCB has a footprint for an SMD Shunt and the two large through hole pads (visible on the lower left side in Figure 1) are available for high current wire soldering.

The meter was tested for a range of current from 0.1A to 20A using the Fluke 6100A Electrical Power Standard. Measurement results are visible on the LCD or on the Pulse Output.

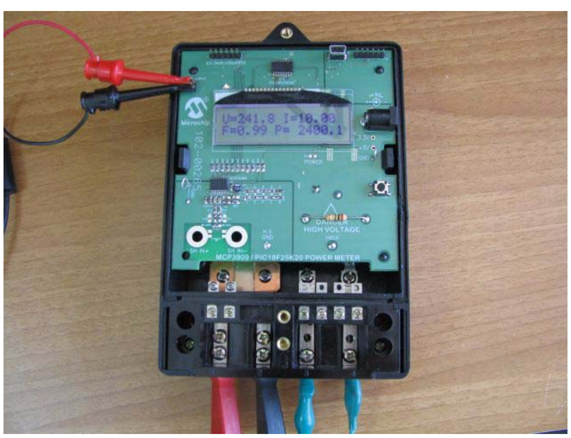

FIGURE 1: *Low-Cost Single Phase Shunt Meter using MCP3909 and PIC18F25K20.*

Microchip's MCP390X metering ADC family has been developed to simplify the design of energy meters.

This document is intended to provide guidance for designers who are interested in using Microchip's MCP3909 Metering ADC with synchronous sampling and PGA on current channel, and the low-cost high performance PIC18F25K20 microcontroller.

The LCD is used to indicate the RMS current and RMS voltage (U_{RMS} and I_{RMS}) on the first row, and the Power Factor and Active Power on the second row.

For systems that require more than local display or monitoring, a second circuit is available for remote monitoring through USB. On the same PCB, there is also a USB circuit, isolated by the meter circuit with optocouplers. This circuit uses the 8-bit PIC18F14K50 microcontroller with USB. The USB circuit is powered from the PC. The firmware given has the USB set up as a virtual COM port with a 19200 baud rate.

The user can easily change the meter firmware to send various information from the meter to PC for different tasks, such as accuracy evaluation, calibration, etc.

HARDWARE DESCRIPTION

The MCP3909 is an energy-metering IC designed to support the IEC 62053 International Metering Standard Specifications. It supplies a frequency output proportional to the average active real power, with simultaneous serial access to ADC channels and multiplier output data. This output waveform data is available at up to 14 kHz with 16-bit ADC output and 20-bit multiplier output words. The 16-bit delta sigma ADCs allow for a wide range of I_B and I_{MAX} currents and/or small shunt (< 200 µΩ) meter designs. The integrated on chip voltage reference has an ultra-low temperature drift of 15 ppm per degree C.

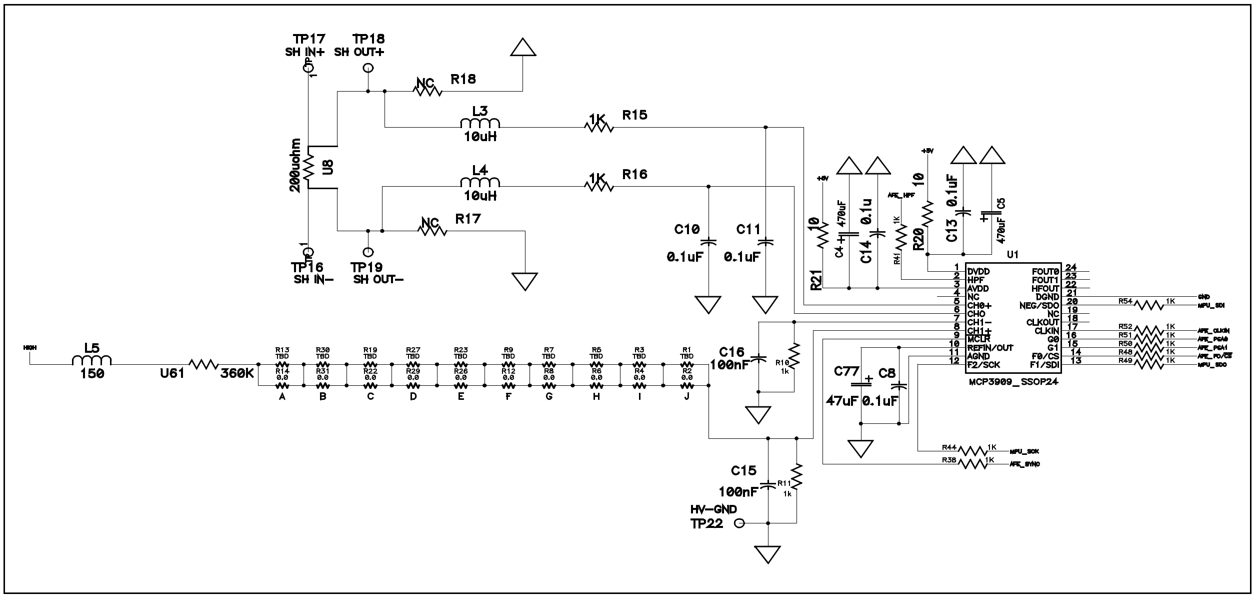

FIGURE 2: *Power Meter Analog Circuit with Shunt.*

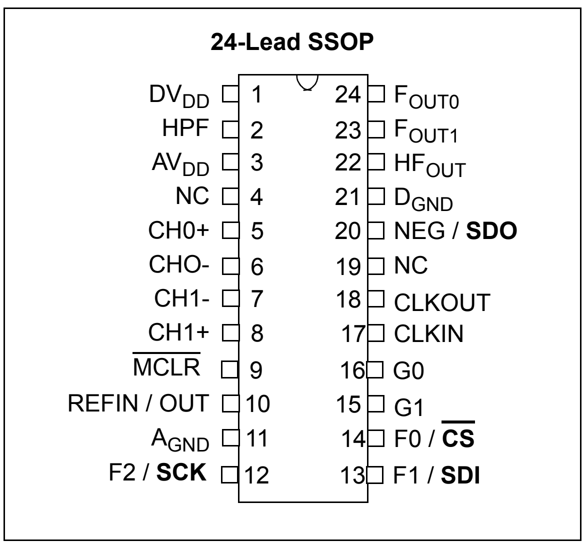

FIGURE 3: *MCP3909 Pinout.*

The CH0 is used for current measurement because the PGA will allow the use of small value shunts. To select a shunt, the desired current range must be established first, the maximum value in particular, so that the voltage drop is less or equal to the maximum CH0 Voltage, as presented in Table 1.

TABLE 1: GAIN SELECTIONS

G1	G0	CH0 Gain	Maximum CH0 Voltage
0	0	1	±470 mV
0	1	2	±235 mV
1	0	8	±60 mV
1	1	16	±30 mV

In the design presented here, the gain used is 16, so the maximum peak voltage on the shunt must not exceed 30 mV, or, for convenience, the maximum RMS value should be under 20 mV. For a 80A meter design, the recommended shunt value is 200 µΩ, as used in this design. Since the MCP3909 is a part with a dynamic range of 1000:1, then the minimum measured current with good accuracy is 0.1A. To be able to obtain this value in an application, the sources of noise around the ADC must be eliminated, otherwise the low current measurements will be less accurate than the 0.1% that can be achieved by the MCP3909.

The PCB layout and grounding scheme used on this demo board was sufficient enough to allow for these low current measurements, as will be described in later sections of this application note.

METER FIRMWARE

In this design, the MCU receives the current and voltage samples from the MCP3909's 16-bit ADCs. Inside the microcontroller, these samples are processed in order to get the U_{RMS}, I_{RMS}, Power Factor, Active Power and Energy Accumulation.

The communication between the MCP3909 and PIC18F25K20 is SPI, and after the last byte is received the PIC starts the calculation phase for all the power quantities. Each data transfer causes an SPI interrupt to begin this calculation on each sample.

Because the calculations are done after each current and voltage sample is aquired, and because it is quite a long process using the PIC18F MCU, the sampling speed of the meter is limited below that which the MCP3909 can sample. The achieved sampling speed on current design is 880 sps.

The sampling speed can be increased by removing some of the computation tasks. For example, if only the Energy Accumulation and Active power is computed, the sampling speed can be increased at 1.6 ksps. The signal processing can be summarized in Figure 4.

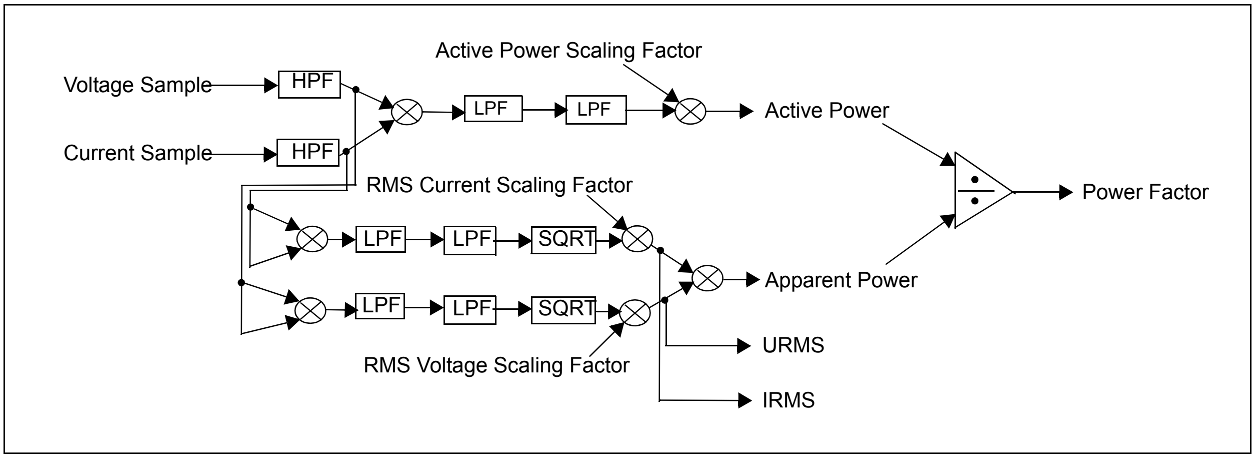

FIGURE 4: *Power Meter Signal Processing.*

As presented in Figure 4, the MCU must compute the result of 8 first order IIR filters, and this takes almost 1 ms. In the existing firmware, to compute the output of IIR filters, numbers in Double format are being used to avoid rounding and saturation. A way to decrease the computation required to get the results is to use 16-bit or 32-bit number formats and write most of the code in assembly. The code written here was done in C and this increases the processing speed of the calculation.

All IIR filters are first-order filters. For the low-pass filters, it is possible to use a second-order IIR filter, but the coefficients will be numbers that will not be represented very accurately in 16 or 32-bit number format; the coefficients will be numbers very close to 1 or to 0. The time needed to compute the two first-order IIR filters is equal to the time required to compute a single second-order IIR filter. The IIR filters structure is Direct Form I, this structure being presented in Figure 5.

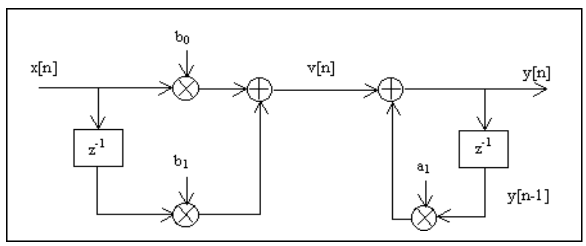

FIGURE 5: *First Order IIR Filter Direct Form I Structure.*

In the case where the ADC noise is much lower than the measured signal, the only adjustment necessary for a correct indication are three scaling factors. However, if the ADC noise is close to the signal level, this could induce an offset like error, so, next to the scaling factor an offset compensation might be required. This happens usually in low-value shunt designs, while measuring low currents.

CALIBRATION

For this design, the calibration is done by modifying the scaling factors in code. As long as the noise of the ADC is much lower than the signal acquired, the offset removal will not be necessary. In this case, a one-point calibration is enough. For situations where the shunt value is low (under 200 µΩ) at low current the noise will have an effect on the calculation, since its RMS value will add to the measurement. For this situation, an offset compensation will be necessary by doing a two-point calibration.

In Figure 6 are presented two situations for the current measurement, using two shunt values: 100 µΩ and 1000 µΩ. This figure shows that for the low value 100 µΩ shunt, the noise affects the lower current ranges. For the higher value 1000 µΩ shunt, the current goes much lower before the noise has an impact on the accuracy.

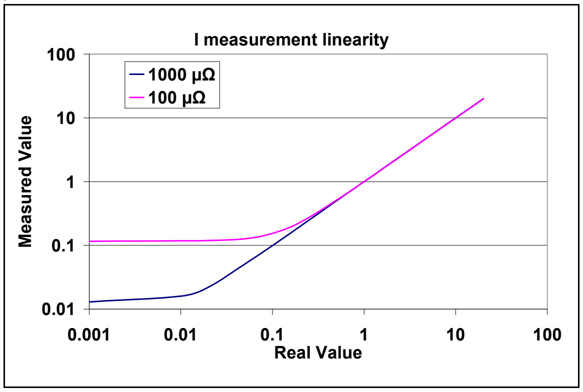

FIGURE 6: *Current Channel Non-Linearity.*

To increase the linear region for the desired range, two solutions are possible: the first one is to use a higher shunt value, while the second solution is to use a non-linear calibration equation for the low current region.

Another problem is the calibration for different power factors (PF). It is possible that for PF = 1, the errors are very small, but for other PF values the errors will increase significantly. This is mostly caused by the difference between the two analog channels (voltage and current). The current channel will have gain introduced by the PGA, and the voltage channel will not. In addition, the RC circuits at the inputs for the two channels can be different because of the part's tolerance.

To ensure low errors at different values for PF, one component from the RC circuits must be modified. To modify just one value with maximum effect the best part is the capacitance in parallel with the low side resistor of the voltage divider from CH1. The value of that capacitance (C15 in this design) will be changed gradually until the measurement error will be minimum on a wide range of PF. Figure 7 shows the evolution of errors for different power factors and different values for C15.

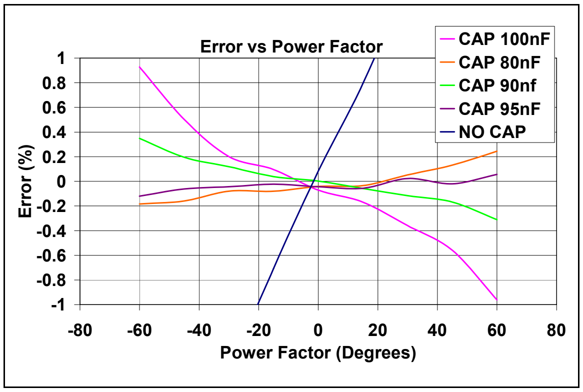

FIGURE 7: *Measurement Error vs. Power Factor for Different Capacitors (C15).*

POWER FACTOR COMPENSATION

One of the main tasks in designing an energy meter is to minimize the effect of power factor variations over the measurement accuracy. In order to have accurate measurements over a wide range of power factors, the input of the MCP3909 ADC has identical current channel and voltage channels. Any difference between channels can cause phase shift between the current and voltage samples and this will cause a decrease in accuracy. The external passive components can induce phase shift because of the parts value tolerances.

A way to minimize the effect of the phase shift on accuracy is to make a multi-point calibration, in order to have information about the effect of different power factors to meter accuracy.

Figure 8 shows that the accuracy varies monotonously with the angle between the voltage and current. On a narrower range (-45, 45 degrees) this variation can be considered linear with a good accuracy. Therefore, it is possible to determine the equation of the best fit line between the calibration points -45, 0, 45 degrees (the yellow points in Figure 8).

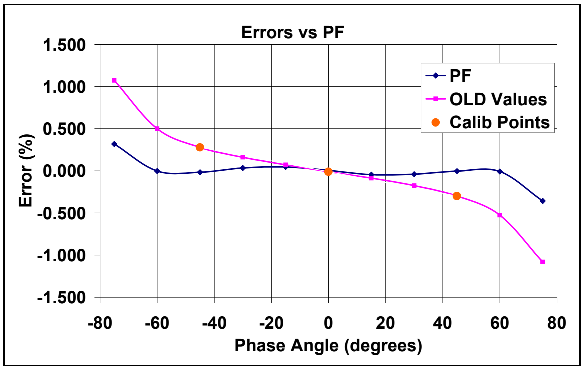

FIGURE 8: *Error vs. Phase Angle.*

The pink line is the variation when calibration was done in one point, without considering the effect of PF over the accuracy of measurements.

The green and blue lines are the errors after the meter was calibrated in three points and the extra information was used to compensate the PF. It is visible how in the -60, 60 degree range, the situation improved visibly from ±0.5% to less than ±0.2%.

The three points calibration is used to determine the best fit line equation that will be used to modify the active power scaling factor relative to the power factor.

The calibration method is quite simple. First, the meter is calibrated at one point: 1A, 0 degrees. The value of the Active Power Calibration Constant is noted. After that, the errors at -45 and 45 degrees are recorded. Using these values, it is possible to compute the ideal value of the Active Power Calibration Constant to have minimum errors in these three points. The three values of the Active Power Calibration Constant are plotted in a graph (Figure 9). On the X scale, not the actual PF is displayed, but a selection of normalized values in order to have the three points in a line.

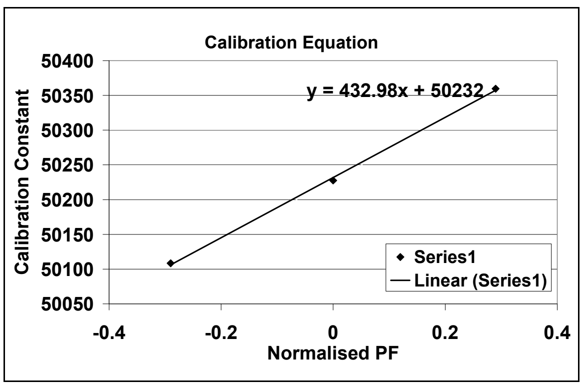

FIGURE 9: *Determination of the Active Power Calibration Constant Equation.*

The equation from the graphs is used to compute the Active Power Calibration Constant at different PF.

The X axis could have been scaled in degrees, but this would have meant to compute the PF value in degrees in the MCU also. The normalized values were used for simplicity and minimum MCU computation power requirement.

LINE FREQUENCY COMPENSATION

A 50 Hz line frequency was used, and typically this is the frequency most of the time. However, this is not a constant, and it can vary around this value by a few Hertz. This line frequency shift can cause measurement errors because of the Sinc filter characteristics at low sampling speeds. The Sinc filter transfer function is similar to a low-pass filter until a frequency equal with the sampling speed where the signal is completely attenuated. Depending on the sampling speed of the ADC, this low-pass filter can be narrower or wider. Figure 10 shows three situations, for 880 sps (as in this meter), 1200 sps and 3200 sps. In Figure 11, the (48 Hz, 52 Hz) range is magnified and the Y axis is scaled to have a crossing at 50 Hz between all three cases.

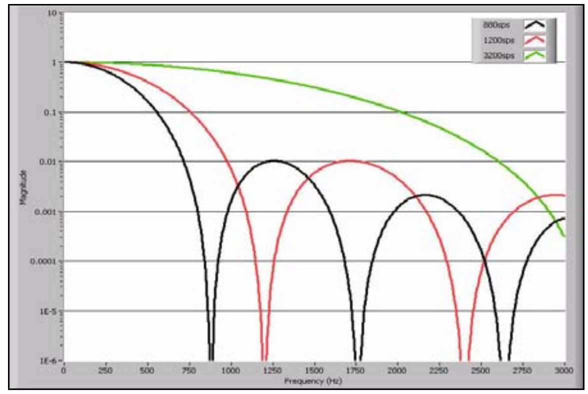

FIGURE 10: *Sinc Filters Transfer Functions.*

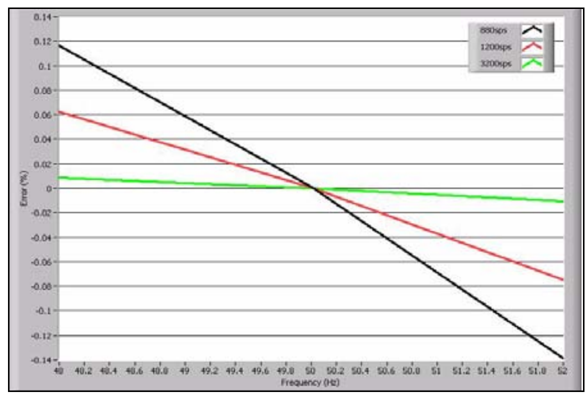

FIGURE 11: *Errors Caused by Line Frequency.*

Here it is visible how the low speed ADC is causing a sensitive attenuation of the signal when the line frequency is higher than 50 Hz relative to the situations when line frequency is lower than 50 Hz. The measurement differences can be higher than 0.2%. In order to have accurate measurements, no matter the line frequency, it is necessary to compensate for these low-pass filter situations.

The compensation is normally done using complex long FIR structures called Sinc Compensation Filters. Such a structure is impossible to be implemented in this application when the MCU is already used close to its maximum computation power. The solution is to adjust the cut-off frequency of the HPF to a value for which the transfer function of the HPF will compensate the Sinc transfer function around the 50 Hz value. The simulation and the measurements indicate that a cut-off frequency of 9 Hz for the HPF will be the best choice in this case (see Figure 12).

The actual error measurements in the 48 Hz to 52 Hz range at the current equal with 1A are presented in Figure 13.

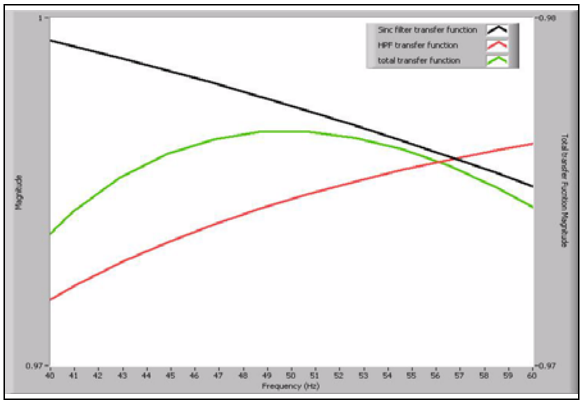

FIGURE 12: Sinc Filter Compensation using HPF.

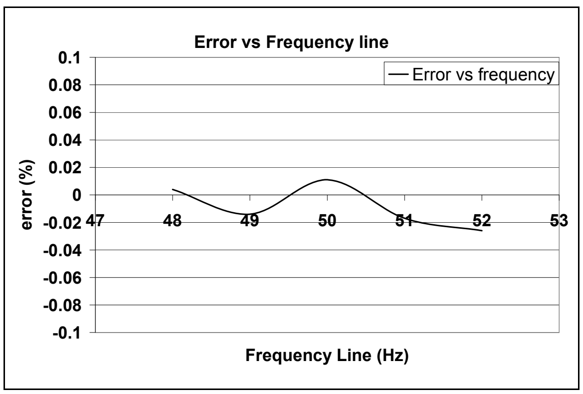

FIGURE 13: Errors vs. Line Frequency.

This is a simplified solution and it will not compensate for frequencies where harmonics might be, yet it improves the overall performance of the meter accuracy significantly. One drawback of this method must be mentioned. As is noticed, the signal will be attenuated a little more that the case when the HPF is having a lower cut-off frequency. This extra attenuation will increase a little the measurement errors at low currents, where the situation is already difficult because of the lower SNR. But in this situation, this accuracy decrease is less than 0.1% and it is considered acceptable.

ENERGY METER ACCURACY

The overall accuracy of the meter was measured using the Fluke 6100A Power standard. The measurement was done on a range of 0.1-20A, PF = 1, 0.5, -0.5, at 50 Hz, 49 Hz and 51 Hz. The results are presented in Table 2 and plotted in Figure 14.

TABLE 2: METER ACCURACY

	PF=1	PF=0.5	PF=-0.5	PF=1	PF=1
I	**50 Hz**	**50 Hz**	**50 Hz**	**51 Hz**	**49 Hz**
0.2	-0.158	-0.947	-0.519	-0.449	-0.648
0.5	-0.116	-0.042	0.027	-0.24	-0.114
1	-0.098	-0.084	-0.178	-0.177	-0.162
2	-0.122	-0.176	-0.118	-0.124	-0.068
5	-0.028	-0.078	-0.066	-0.011	-0.03
10	0.028	-0.064	-0.013	0.028	0.028
20	0.028	0.028	0.028	0.028	0.028
40	0.028	0.028	0.028	0.028	0.028
80	0.028	0.028	0.028	0.028	0.028

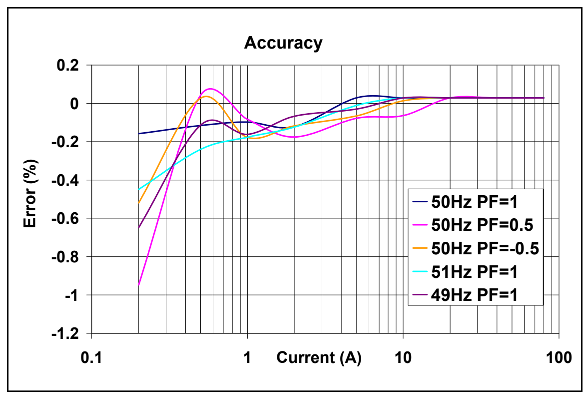

FIGURE 14: Energy Meter Accuracy.

REFERENCES

[1] MCP3909 Data Sheet, *"Energy Metering IC with SPI Interface and Active Power Output"*, Microchip Technology Inc., DS22025, 2008.

[2] AN1151, *"PIC18F2520 MCP3909 3-Phase Energy Meter Reference Design"*, Microchip Technology Inc. Microchip Technology Inc., DS01151, 2008

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