

IEC Compliant Active-Energy Meter Design Using The MCP3905A/06A

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OVERVIEW

The MCP3905A/06A energy-metering Integrate Circuits (ICs) supply active real power measurements for single-phase residential meter designs. These devices include features specific for the International Electrotechnical Commission (IEC) requirements, such as no-load threshold and start-up current. In addition, the MCP3905A/06A Energy Meter Reference Design demonstrates a system-level design that passes EMC immunity requirements per the IEC standard.

EMC testing for IEC compliance was performed on the Energy Meter Reference Design Demonstration Board used in this application note. These tests were performed by a third party, the results of which are included at the end of this application note.

Accuracy tests were also performed by a third party using the MCP3905A device across current ranges, power factors, V_{DD} and line frequency conditions outlined in the IEC standard. The results of these tests are also included in the "Summary".

This application note can be used for both MCP3905A/06A stand-alone meters and PICmicro[®] microcontroller-based meters using the MCP3905A/06A devices as the Analog Front-End (AFE). In both meter designs, the system accuracy and EMC immunity rely on the AFE design. Examples of both meter types are shown in [Figure 1](#).

GOALS

The goal of this application note is to discuss design decisions relating to meter design when using the MCP3905A or MCP3906A device that will lead to IEC compliance.

Meter ratings and current sensor choices are discussed first. The trade-off between shunts and Current Transformers (CTs) is described, as well as how the more accurate MCP3906A can be used as an advantage in meter design. This application note will focus on the shunt as a current-sensing choice and contains sections for compensation relating to the shunt only.

Proper selection of the PGA gain is covered, describing how shunt sizes, power consumption goals and signals with high crest factors guide this decision. Directions follow that describes how to select the proper output frequency and calibrate the meter using single-point calibration to a typical meter constant of 100 imp/kWh (impulses/kilowatt hour).

System-level design decisions follow, including EMC compliance and LRC filter design. Microchip's free FilterLab[®] design tool is used to show anti-aliasing filter design and phase-shift trade-offs. A spreadsheet with complex frequency analysis is included to assist in the compensation calculations, best showing graphically the effects of component values in the frequency domain.

Power supply design is also discussed, including how to choose capacitor values based on meter-specific current requirements.

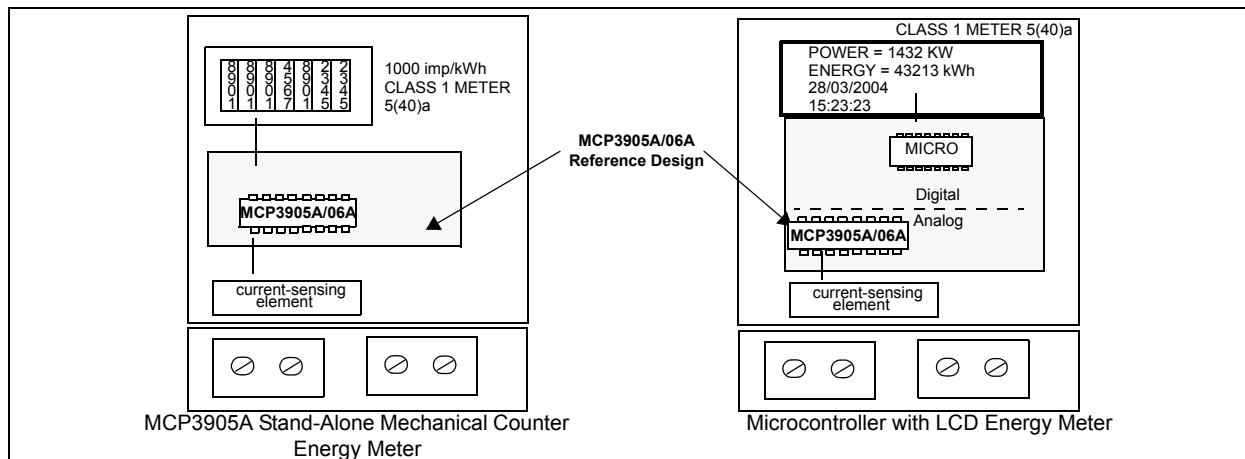


FIGURE 1: MCP3905A/06A Stand-Alone Energy Meter and Microcontroller-Based Energy Meter.

ENERGY METER STANDARDS

Many international standards committees and energy meter specifications (ANSI, IEEE, etc.) exist with local government standards and in areas of large population. Selection of IEC standards to demonstrate MCP3905A/06A performance requirements is purely arbitrary; there is no recommendation from Microchip on which standards should apply.

Purchasing the standards to design for IEC compliance is recommended. This application note does not act as a replacement for the actual standard. This document will only describe how the MCP3905A/06A and the MCP3905A/6 Energy Meter Reference Design relate to the IEC specifications and compliance.

The IEC distinguishes between active-energy meters and reactive-energy meters. This application note and the MCP3905A/06 devices apply to active-energy meters only. There are four classes of meters: Class 2, Class 1, Class 0.5 and Class 0.2 meters. Accuracies described apply to both single-phase and poly-phase meters, with slight differences described within the standards. The standards are all a part of the IEC 62053 specification. This specification replaces the previous specifications IEC61036, IEC1036 and IEC687.

CURRENT SENSING AND DYNAMIC RANGE REQUIREMENTS

Current-Sensing Shunt

The current-sensing shunt is a small piece of metal (typically made of manganese and copper) that is manufactured with a variety of mounting holes and wired connections. It acts as a simple resistor, with the voltage drop across it proportional to the current flowing through it. Parasitic inductances create frequency dependant voltage drops and mismatch in the phase response between channels. In addition, this parasitic inductance has a frequency response that can affect the anti-aliasing networks protection against RF interference. Methods for compensating for this inductance is described later, along with complex frequency analysis using Excel® spreadsheets, in "**Compensating For Parasitic Shunt Inductance**".

Shunt resistances are typically between 100 $\mu\Omega$ and 500 m Ω , with inductances specified between 1 and 5 nH. Though the shunt is very low in cost, it is ultimately limited by it's own self-heating and can typically not be used in meter designs with large maximum current requirements ($I_{MAX} \gg 100A$). The power consumption that the shunt brings to a meter design is also large compared to that of a CT. While lower-value shunts (<250 $\mu\Omega$) offer less power consumption, the resulting V_{RMS} signal going to the Analog-to-Digital Converter (ADC) can be difficult to measure using lower-resolution (<12 bit) ADCs. See "**Required Accuracy – ADC**" for examples using specific meter ratings.

CURRENT TRANSFORMER (CT)

The current transformer is another choice for sensing current when designing an energy meter. The device offers isolation through transfer of current from the primary to the secondary winding. The CT can handle higher currents than the shunt, while consuming less power. The trade-off is cost and, in some situations, accuracy. The CT is also susceptible to saturation through either a large DC component or a large overcurrent situation. When the core saturates, the device becomes very non-linear. Mu-metal CTs offer better protection against saturation, though with a higher non-linear phase response. The non-linear phase response of the CTs causes power and energy measurement errors at low currents, with large power factors. Compensating for this non-linearity is not within the scope of this application note; more accurate meter classes than this non-linear phase response must typically be compensated for in the firmware.

Table 1 summarizes some of the trade-offs between the shunt and the CT.

TABLE 1: SHUNT VERSUS CT TRADE-OFFS

Advantage	Shunt	CT
Lower Cost	X	—
High Current	—	X
Lower Power Consumption	—	X
Less Accuracy Issues (Saturation, Phase Response at high-power factors)	X	—

Direct-Connect Meters

Direct-connect meters are those meters connected directly to the power lines using a shunt resistor on the current channel and/or a voltage divider on the voltage channel. Those isolated using a current transformer are not direct-connect meters and some accuracies are defined separately in the IEC standards. The direct-connect or shunt meters have slightly reduced accuracy requirements, as shown in [Table 2](#).

METER RATINGS

The IEC specification requires accuracy down to a percentage of the meters' base or normal current (I_N). For example: a specific energy meter could be rated at 10(40) A, Class 1. This would be an I_N of 10A and an I_{MAX} of 40A.

Power factor also plays a role in the requirements. For low-power factor accuracies, the AFE must have almost zero (<0.1 degree) phase-matching in the channels. The MCP3905A/06A has simultaneous sampling and is phase-matched to <0.01 degree. System-level influences, such as a delay from a CT on one channel, can require additional compensation. System-level phase-matching methods, for use with the MCP3905A/06A and phase-matching calculation examples, are given in a later section. When selecting a shunt size or CT turn ratio, the accuracy of the AFE and power-consumption goals are considered.

SELECTION OF G1, G2 GAIN

The gain selection is dependant on the choice of the current-sensing element, shunt size, meter rating and power-consumption goals. A smaller shunt will consume less power and require larger gain, but at a cost of accuracy in low-current situations. The MCP3905A/06A has an improved ADC on both channels, with almost true 16-bit accuracy, 15.7 Effective Number Of bits (ENOb) across all gain settings. The low-noise ADC will allow for smaller shunts for meter designs, with higher dynamic range requirements. As an example, a 5(80) A Class 1 meter with a meter constant of 100 imp/kWh.

The 80A maximum current could easily dissipate over 2W through a larger shunt, so the goal is to choose as low a value shunt as is possible and still be within the typical accuracy bounds specified in the MCP3905A/06A data sheet. From [Table 2](#), we see that the IEC specification requires our accuracy to be down to that of 5% of I_N for a Class 1 meter. As an example, with an I_N of 5 and an I_{MAX} of 80, this equates to a dynamic range requirement of 320:1.

This current range would be 250 mA to 80A. A 250 $\mu\Omega$ shunt would give V_{RMS} signals in the range of 62 μV to 20 mV. Peak values must be considered. In dealing with loads that cause high crest factor signals, such as those in [Figure 2](#) (and also overcurrent situations), the design goal is to keep the V_{RMS} signal of channel 0 to around half of the full-scale input range at I_{MAX} . Not only must the bandwidth of the ADC be high enough to capture the higher frequency harmonics associated with signals like this, but the input range of the ADC must not be saturated with the peak values. The crest factor for a given signal is the ratio of RMS to peak voltage. In some systems, the crest factor might be limited during design (e.g., a lighting system used in a house or factory might be designed to keep the crest factor below 1.75). Estimating the crest factor depends on your expected energy meter loads.

As mentioned with V_{RMS} signals in the range of 62 μV to 20 mV, the design goal is to keep the full-scale current at around half of the full-scale input range of the ADC. With a gain of 8 for the MCP3905A, the specified range is ± 32 mV, with a typical 0.1% error down to 64 μV , or 500:1 dynamic range. For the 5(80)A meter example with a 250 $\mu\Omega$ shunt, choosing $G = 8$ over a 500:1 dynamic range using the MCP3905A is a satisfactory choice. However, choosing the MCP3906A device would give more spread at both ends of the current range, as this device is specified with an increased dynamic range of 1000:1.

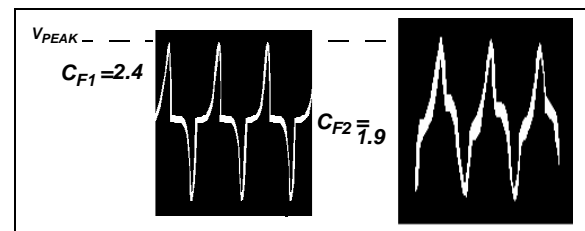


FIGURE 2: Channel 0 Signals With a High Crest Factor, 2.4 and 1.9.

TABLE 2: IEC ACCURACIES AND DYNAMIC RANGE REQUIREMENTS BASED ON METER CURRENT RATINGS AND CLASS FOR DIRECT CONNECT METERS

Current	Power Factor	Class 2	Class 1	Class 0.5	Class 0.2
$0.01I_N < I < 0.05 I_N$	1	—	—	±1.0%	±0.4%
$0.05I_N < I < 0.1 I_N$	1	±2.5%	±1.5%	—	—
$0.1I_N < I < I_{MAX}$	1	—	—	±0.5%	±0.2%
$0.05I_N < I < I_{MAX}$	1	±2.0%	±1.0%	—	—
$0.02I_N < I < 0.1 I_N$	0.5 inductive	—	—	±1.0%	±0.5%
	0.8 capacitive	—	—	±1.0%	±0.5%
$0.1I_N < I < 0.2 I_N$	0.5 inductive	±2.5%	±1.5%	—	—
	0.8 capacitive	—	±1.5%	—	—
$0.1I_N < I < I_{MAX}$	0.5 inductive	—	—	±0.6%	±0.4%
	0.8 capacitive	—	—	±0.6%	±0.4%
$0.2I_N < I < I_{MAX}$	0.5 inductive	±2.0%	±1.0%	—	—
	0.8 capacitive	—	±1.0%	—	—

Required Accuracy – ADC

The required resolution for a meter design depends on the current range and accuracy requirements listed in [Table 2](#). The overall accuracy of the AFE ADCs should be well above the requirements listed in this table. Based on the fact that you are measuring energy (i.e., accumulating power over time), there is time in your budget to use averaging to increase the output resolution.

As an example, let us look at a Class 2 meter with a current rating of 5(80)A. As shown in [Table 2](#), the current range required is from $0.05I_b$ to I_{MAX} for this meter. That equates to a dynamic range of 320:1. Assuming the intended accuracy was 1% (to be below the 2.5% accuracy requirement), the following calculation shows the required accuracy converted to base 2, in bits:

EQUATION 1:

$$\frac{I}{320} \approx 0.01 = 0.00003125$$

$$\frac{\ln(0.00003125)}{\ln(2)} = 14.9 \text{ bits}$$

The MCP3905A and MCP3906A are designed to be 0.1% accurate over a dynamic range of 500:1 and 1000:1, respectively. For the MCP3906A, this equates to an overall required accuracy of 19.9 bits. The MCP3905A/06A devices contain two 16-bit ADCs with an effective resolution of 15.7 bits. The low-pass filter on the output of the multiplier increases this accuracy to fulfill the 500:1 and 1000:1, 0.1% requirement, covering all Class 2, Class 1, Class 0.5 and Class 0.2 meter designs.

METER POWER DISSIPATION

The energy meter cannot consume too much energy, and is typically limited to 2W total power consumption. The power dissipation for static meters of active energy is given in [Table 3](#), Part 21 of IEC62053 standard for Class 1 and Class 2 meters, Section 7.1.

TABLE 3: POWER CONSUMPTION COMPLIANCE

	Class 1	Class 2
Total	2W and 10 VA	2W and 10 VA
Current Only	4 VA	2.5 VA

For 80A maximum current, the 250 $\mu\Omega$ would consume $I^2 \cdot R$ or $80^2 \cdot 250e-6$ or 1.6W. This leaves 400 mW for the remaining meter power. An advanced microcontroller-based meter design would then be limited to less power and a slower clock frequency. Smaller shunts (<200 $\mu\Omega$) can be used with proper attention paid to inductance compensation for noise and phase-shift issues, as detailed in "[Compensating For Parasitic Shunt Inductance](#)".

SELECTION OF F2, F1, F0

After shunt selection, the next step in meter design using the MCP3905A or MCP3906A is selecting the frequency constants, F_C and HF_C . As discussed earlier, based on crest and overcurrent issues, the maximum current for your meter design (I_{MAX}) should give a V_{RMS} signal of no more than half the full-scale input range for a given gain selection. This will be used to determine our frequency settings.

For a meter constant of 100 imp/kWh, Table 4 has been generated for a number of popular I_{MAX} currents.

TABLE 4: OUTPUT FREQUENCY FOR MAXIMUM CURRENT

Line Voltage (V)	IMAX (A)	Power (KW)	Meter Constant K	Output Frequency of F_{OUT0}, F_{OUT1} (Hz)
220	25	5.5	100 imp/kWh	0.153
220	40	8.8	100 imp/kWh	0.244
220	60	13.2	100 imp/kWh	0.367
220	80	17.6	100 imp/kWh	0.489

The output frequency listed in Table 4 is the frequency you should desire when the V_{PP} signal going into channel 0 is at approximately half of the full-scale input range. Table 5 has been generated showing what the output frequency would be with half-scale inputs on both channels for the four F_{OUT0}, F_{OUT1} frequency options on the MCP3905A/06A.

From the output frequency generated in Table 4, select the closest output frequency in Table 5. This determines what the best F1, F0 settings should be for a given meter design.

TABLE 5: OUTPUT FREQUENCY WITH HALF SCALE INPUT

Output Frequency of F_{OUT0}, F_{OUT1} (Hz)	F1	F0
0.085	0	0
0.17	0	1
0.34	1	0
0.68	1	1

The selection of F2 depends on the selected calibration method, the base or normal current used during calibration and the desired calibration frequency. If an optical sensor is used during calibration, the maximum output frequency it can detect is sometimes as low as 20 Hz. The ratio of HF_{OUT} to F_{OUT} frequency is given in Table 6. For each F1, F0 selection, there are two possible multipliers. This allows two options for your calibration constant. Note that all options yield around 1-3 Hz as an output frequency for a given I_B , ideal for optical calibration equipment. I_B , in this case, was

assumed to be 1/4 of I_{MAX} (e.g., 10(40)A meter, also assuming I_{MAX} was set at half of the full-scale input range, as mentioned above).

TABLE 6: HF_{OUT}, F_2 SELECTION

F2	F1	F0	F_{OUT} to HF_{OUT} Multiplier	Output Frequency of F_{OUT0}, F_{OUT1} with Half-Scale Inputs (Hz)	Calibration Output Frequency of HF_{OUT0} with 1/8 Scale ($\sim I_B$) (Hz)
1	0	0	128	0.085	2.72
0	0	0	64	0.085	1.36
1	0	1	64	0.17	2.72
0	0	1	32	0.17	1.36
1	1	0	32	0.34	2.72
0	1	0	16	0.34	1.36
1	1	1	16	0.68	2.72
0	1	1	2048	0.68	348 (Note 1)

Note 1: This logic settings should be used with a MCU connection only, not optical calibration equipment.

Single-Point Meter Calibration

The MCP3905A/06A Energy Meter Reference Design uses a voltage divider network on channel 1 for single-point meter calibration. With the digital high-pass filter turned on in both of the input channels of the MCP3905A/06A devices, active single-point calibration is an option based on the removal of any DC offset by the digital filters. This single-point calibration removes any gain errors in the system due to shunt tolerances, V_{REF} tolerances or any other error.

The resistor divider network on the MCP3905A Energy Meter Reference Design has a series of weighted resistors, with shorting jumpers for each. In addition, there are two extra 330 k Ω resistors in front of the calibration network. These two resistors will always be present and are part of a scheme to maintain a 3 dB point for phase-matching.

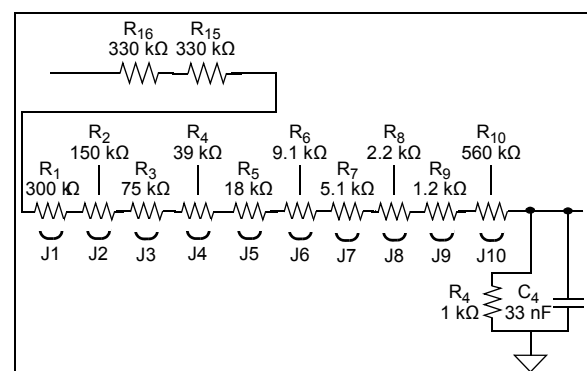


FIGURE 3: Resistor Divider Calibration Network.

To calibrate the meter, start with all jumpers removed. Test each jumper, starting with J1. Solder the jumper on and observe the output frequency. If the output frequency of the meter is too slow with the jumper on, remove the jumper and continue with the next, testing all jumpers once.

With the resistor network populated as shown in Figure 3, the -3 dB point of the anti-aliasing filter is dominated by the two 1 kΩ resistor R₄, regardless of how many jumpers are closed. The two 330 kΩ resistors, R₁₆ and R₁₅, ensure this.

ANTI-ALIASING FILTER DESIGN

The MCP3905A/06A devices contain two 16-bit delta-sigma ADCs that oversample at MCLK/4. Using the NTSC, the standard video subcarrier frequency of 3.579 MHz makes a good low-cost clock source. The MCP3905A/06A devices are designed to operate from a clock source of 1 MHz to 4 MHz, with all data sheet specifications and device characterization performed using a frequency of 3.579 MHz. At this frequency, the sampling rate is approximately 900 kHz. (MCLK/4). Any frequencies above half this should be eliminated by at least -40 dB using an anti-aliasing filter.

The goal of this design choice is to eliminate signals above the Nyquist frequency (half of the sampling frequency). With a 3.58 MHz oscillator, this is ~450 kHz.

The bandwidth of interest depends on the number of harmonics included in the conversion. A channel-to-channel difference in phase of 0.1 degree can cause >0.5% error. Microchip's FilterLab software design tool is used to compare these trade-offs in filter design.

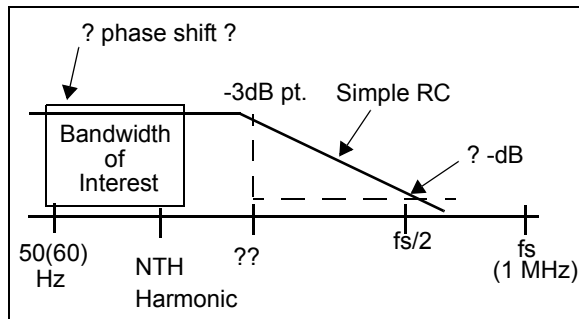


FIGURE 4: Filter Attenuation versus Phase-Shift Trade-off.

The goal is to determine the attenuation at fs/2 and compare that to the phase delay around 50/60Hz, and how component-matching will cause problems in channel-to-channel phase-matching.

Using the FilterLab software, change the filter order to 1 by clicking up on the filter order box. Part 21 of the standard (Class 1 and Class 2 meters) requires accuracy in the presence of up to the 5th harmonic. The MCP3905A/06A/6 devices have an input bandwidth of 14 kHz (or MCLK/256). Start with a cut-off frequency of 5 kHz. To determine the attenuation at Nyquist (MCLK/4/2 = 450 kHz), moving the cursor over the frequency response window. This will show you what the magnitude and phase information is for a given frequency. With this cutoff frequency for our RC filter, we see -39 dB at 450 kHz Nyquist. Figure 5 shows what the FilterLab software filter response screen would look like for a first-order RC with a cutoff frequency of 5 kHz.

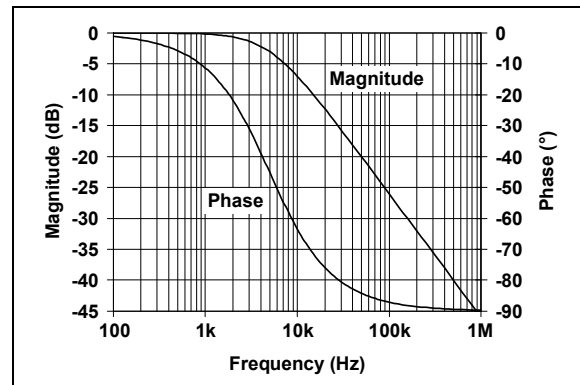


FIGURE 5: FilterLab Software Showing Magnitude and Phase Response.

COMPENSATING FOR PARASITIC SHUNT INDUCTANCE

Note: An Excel worksheet that includes complex frequency has been created to assist in the calculation and to show graphically the effects of component values in the frequency domain. The filename is "s-domain shunt compensation for MCP3905.xls" and can be downloaded at www.microchip.com in the MCP3905A/06A product folder.

The anti-aliasing filter design in the previous section is intended to keep any noise at and above the Nyquist frequency (450 kHz) attenuated below our accuracy threshold (40 dB). The most prevalent noise in this bandwidth comes from AM radio (or other RF signals) coupled through the power transmission lines. When a shunt is directly connected to this input network, the frequency response of the shunt must be included in the analysis of the input network.

With shunt values much less than 1 ohm, and inductances between 1-5 nH, a high-pass filter is created that offsets the low-pass, anti-aliasing filter design that attenuates high-frequency signals. The use of Excel and complex frequency analysis will be required to cancel out this high-pass filter effect from the shunt.

The addition of the shunt creates a high-pass filter in front of our previously designed anti-aliasing filter represented by the circuit in [Figure 6](#).

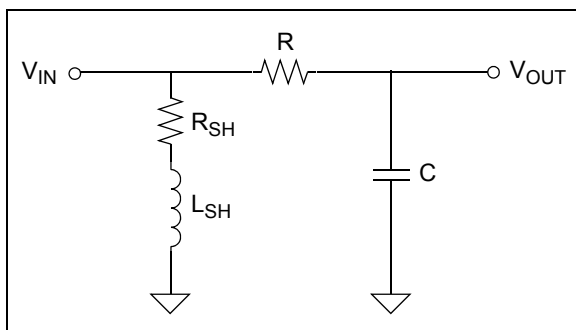


FIGURE 6: Shunt Inductance Causes Zero, Possibly Cancelling RC, Depending On R_{SH} and L_{SH} Values.

The transfer function for this circuit contains a zero from the shunt and a pole from our anti-aliasing filter, represented by:

$$H(s) = \frac{\left(s + \frac{R_{SH}}{L_{SH}}\right)}{\left(s + \frac{1}{RC}\right)}$$

Excel can be used to perform complex number operations using a number of functions in the analysis toolkit. Some of these functions are IMDIV (division), IMPRODUCT (multiplication), IMSUM (addition) and IMPOWER (exponent). The transfer function above is represented by the following complex formula in the Excel spreadsheet:

```
=IMDIV(IMSUM("s",("RSH"/"LSH")),IMSUM("s",("1/("R"*"C"))))
```

Where the s, R, L and C values are to be changed to include the proper R, L, C and s-domain numbers for your application.

The spreadsheet included with this application note demonstrates this and has columns for both the magnitude and phase response. There are also cells to enter the specific Lshunt, Rshunt and RC values. The following figure is the plot from this spreadsheet showing what the magnitude and phase response might look like for a given RSH,LSH, R and C (220 $\mu\Omega$, 3 nH, 1.2 k Ω and 33 nF, respectively).

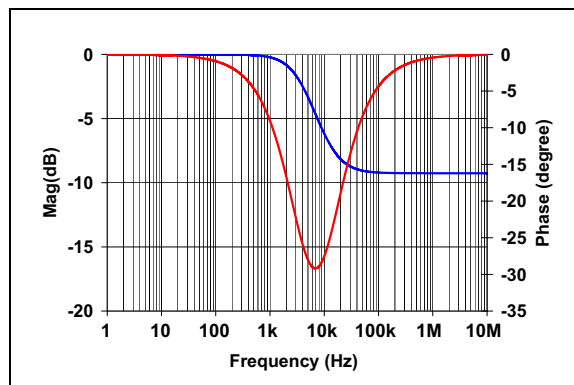


FIGURE 7: Frequency Response With R_{SH} and L_{SH} Zero Close to RC Pole.

As you can see from the plot generated with the Excel spreadsheet, an addition of a zero from the shunt anywhere around the pole of the anti-aliasing RC causes problems at higher frequencies (1 MHz). There is now only -10 dB of attenuation above the zero frequency caused by the shunt. Note that the parasitic capacitances associated with the system would cause another zero higher out, with the frequency response coming back down past 100 MHz. However, this is not included with the transfer function supplied in the Excel spreadsheet.

The addition of another pole to compensate for this zero is required. The least expensive solution would be to add another RC in series, as in [Figure 8](#).

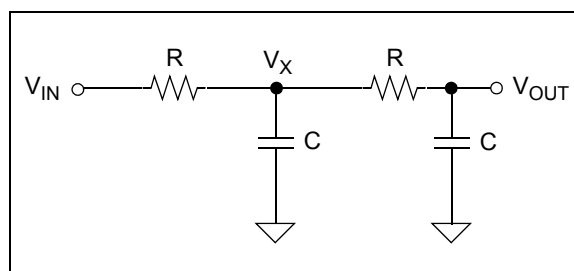


FIGURE 8: Second-Order Filter To Compensate For Shunt Inductance.

By solving for the transfer function of this circuit, and equating one of the $1/RC$ pole frequencies to the R_{SH}/L_{SH} zero frequency, we will have accomplished the compensation.

Use the s-domain impedances and apply Kirchoff's Law to nodes V_X and V_{OUT} to find the transfer function for the circuit:

EQUATION 2:

$$\frac{(V_x - V_{IN})}{R} + sC \cdot V_x + \frac{(V_x - V_{OUT})}{R} = 0$$

$$\frac{V_{OUT} - V_x}{R} + sCV_{OUT} = 0$$

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Solving for V_{OUT}/V_{IN} :

EQUATION 3:

$$2sCRV_{OUT} - V_{IN} + sCRV_{OUT} + s^2C^2R^2V_{OUT} + V_{OUT} = 0$$

$$V_{OUT}(s^2C^2R^2 + 3sCR + 1) = V_{IN}$$

$$\frac{V_{OUT}}{V_{IN}} = H(s) = \frac{1}{s^2C^2R^2 + 3sCR + 1}$$

Solving for the roots of this second equation require the use of the quadratic formula:

EQUATION 4:

$$s = \frac{-b \pm \sqrt{b^2 - 4ac}}{2a}$$

One of the two roots for the equation must be used to cancel the high-pass filter effect caused by the shunt inductance and resistance. By using the shunt zero and picking an arbitrary capacitor value (C), solve for R:

EQUATION 5:

$$s = \frac{-3RC \pm \sqrt{(3RC)^2 - 4(RC)^2}}{2(RC)^2}$$

$$s_1 = \frac{-3RC}{2R^2C^2} + \frac{\sqrt{5}RC}{2R^2C^2} = \frac{-3}{2RC} + \frac{\sqrt{5}}{2RC}$$

The location of the other pole is given by the other root of the quadratic equation solution.

EQUATION 6:

$$s_2 = \frac{-3RC}{2R^2C^2} + \frac{\sqrt{5}RC}{2R^2C^2} = \frac{-3}{2RC} - \frac{\sqrt{5}}{2RC}$$

The values for the second-order RC can now be calculated and put into the Excel spreadsheet to evaluate the performance of the compensation circuit. A second-order filter is also included in the transfer function in the Excel spreadsheet that accompanies this application note. The transfer function and formula using the complex number Excel notation is shown here:

EQUATION 7:

$$H(s) = \frac{\left(s + \frac{R_{SH}}{L_{SH}}\right)}{\left(s + \frac{1}{RC}\right)\left(s + \frac{1}{RC}\right)}$$

In Excel:

```
=IMDIV(IMSUM("s",("RSH"/"LSH")),
IMPRODUCT(IMSUM("s",
(1/("R"*"C"))),IMSUM("s", (1/("R"*"C")))))
```

Figure 9 shows the final result (i.e., the magnitude and phase response using this cancellation technique):

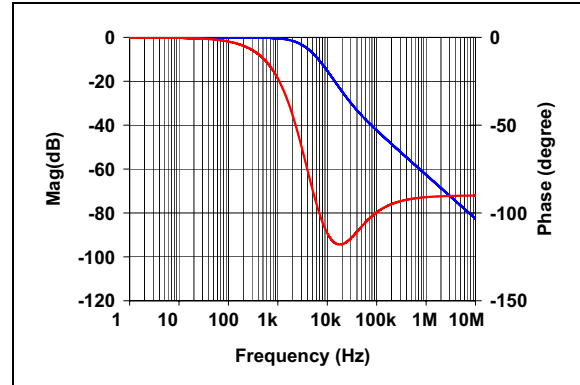


FIGURE 9: Frequency Response With R_{SH} and L_{SH} Zero Being Compensated by Second RC Pole.

As can be seen from this frequency response, the second pole was able to bring the magnitude response near 1 MHz back down to below -40 dB, proving the calculation. The spreadsheet should be used with your own specific shunt values.

POWER SUPPLY DESIGN ANALYSIS

The power on the MCP3905A Energy Meter Reference Design comes from a LM7805 +5V regulator driven by a half-wave rectified zener diode limited AC signal. This circuit is represented in the following figure:

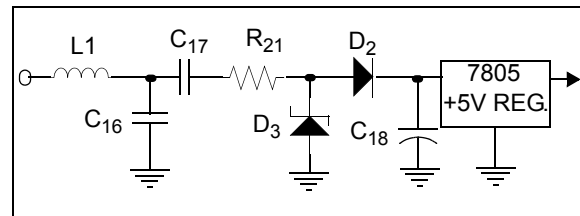


FIGURE 10: MCP3905A/06A Energy Meter Reference Design Power Supply Circuit.

The high-voltage capacitors C₁₆ and C₁₇ divide the line voltage down with a peak current through C₁₇ given by the following equations; for a line voltage of 220V, a line frequency of 50 Hz and a C₁₇ capacitor value of 0.47 μF:

EQUATION 8:

$$\begin{aligned}
 I &= C \frac{dv}{dt} = C \frac{d[V_{sin}(\omega t)]}{dt} \\
 &= C \cdot V \omega \cdot \cos(\omega t) \\
 &= (0.47 \mu\text{F})(220)(\sqrt{2})(2\pi)(50) \cos(1) \\
 &= 45.9 \text{mA (peak)}
 \end{aligned}$$

The presence of diode D₂ leads to the half-wave, rectified waveform shown in Figure 11. This current will charge capacitor C₁₈, required to hold an average DC voltage of 14V to completely drive the input to the 5V regulator LM7805. It is important to analyze the average charging current of this capacitor (C₁₈) when designing system power limitations, especially at system power-up. This analysis can also be used to reduce capacitor values and board cost, depending on the current requirements for your system.

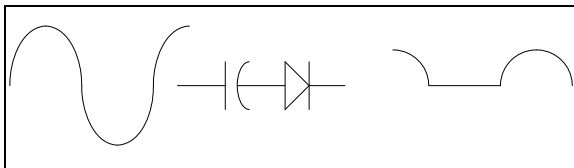


FIGURE 11: Half-wave rectified waveform charging C₁₈ supplies input voltage for regulator and power for meter.

The average current of this half-wave, rectified sine wave is a ratio of 1/π to that of a full sine wave. The average current is calculated from the peak current in that manner.

Table 7 gives I_{AVERAGE} values for different line voltages and for 60 Hz line frequencies. For a 50 Hz line cycle, the average values in the table drop by 50/60 (or 83.3%). An example using 220V, C₁₇ = 0.47 μF and a 50 Hz line cycle frequency would give an I_{AVERAGE} of 14.16 mA.

TABLE 7: CHARGING VALUES, 60 Hz LINE

V _{RMS}	C ₁₇	I _{PEAK}	I _{AVERAGE}
250	0.47 μF	0.63	0.20
220	0.47 μF	0.55	0.17
180	0.47 μF	0.45	0.14
117	1.0 μF	0.62	0.20
110	1.0 μF	0.59	0.19
100	1.0 μF	0.53	0.17

The average current presented here charges C₁₈, which supplies power to the regulator and the rest of the meter. During system power-up conditions, if too much current is required from your system during this time, capacitor C₁₈ will not build enough charge and the regulator will not be fully turned on. The total discharge current for your system must not be greater than the average current charging C₁₈ during this time.

Zener diode D1 limits the voltage at C₁₈ to 14V, at which time D3 is completely turned on. It is necessary to know the number of line cycles required to charge C₁₈ to 14V, as this is used in determining system power-up schemes for your application, as well as MCP3905A Power-On Reset (POR) conditions. Again, here we are using the 220V, 50 Hz numbers, with C₁₈ = 470 μF.

EQUATION 9:

$$\begin{aligned}
 I &= C \frac{dV}{dt} & V(t) &= \frac{1}{C} \int_0^T I_C dt \\
 V_{CHARGE} &= \frac{I_{AVERAGE} \cdot T}{C} & N_{CYCLES} &= \frac{14V}{V_{CHARGE}} \\
 V_{CHARGE} &= \frac{(14.16 \text{mA}) \cdot (20 \text{ms})}{(470 \mu\text{F})} = 0.61 \text{V} \\
 N_{CYCLES} &= \frac{14}{0.61} = 23 \text{ line cycles}
 \end{aligned}$$

Table 8 shows how many line cycle events are required to reach 14V at different line voltages, line frequencies and C₁₇ values. The MCP3905A Energy Meter Reference Design comes populated with a C₁₈ value of 470 μF for a 220V line voltage. If desired, the above formulas can be used to reduce the size of C₁₈ and subsequent board cost.

TABLE 8: C₁₈ CHARGING CYCLES

V _{RMS}	Line	C ₁₇	I _{AVERAGE} on C ₁₈	# Cycles to 14V on C ₁₈
250	50 Hz	0.47 μF	0.017	23
220	50 Hz	0.47 μF	0.015	26
180	50 Hz	0.47 μF	0.012	33
130	60 Hz	1.0 μF	0.013	30
110	60 Hz	1.0 μF	0.011	35
90	60 Hz	1.0 μF	0.009	44

To further simplify your power supply circuit, you can also use the formulas to predict a rise from a different voltage level or zener diode level. Most devices in your system will have a POR circuit that keeps the device in reset until it hits a certain voltage. For example, to complete a rise from 7V, where V_{DD} is regulated to 5V, to a voltage of 14V, you can use:

$$(14-7)/0.61 = 12 \text{ cycles} \Rightarrow 12 * 20 \text{ms} = 0.24 \text{ seconds}$$

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The IEC standard states that the meter should be powered and functional 5 seconds after the reference voltage is applied to the terminals (IEC62053, Part 21, Section 8.3.1). The MCP3905A has a POR delay of 1 second; neither the HF_{OUT} for LED drive or the F₀/F₁ for the mechanical counter drive will occur during this time.

Current Budget

When changing the power supply design (in this case, capacitor values totaling the system current requirements are necessary). The MCP3905A has typical I_{DDD} and I_{DDA} specifications of 1.2 and 3.0 mA, respectively. Assuming there is an output LED on the MCP3905A, as well as a mechanical counter with a 400 ohm coil for the two-step motor, the current budget would be:

EQUATION 10:

$$\begin{aligned}
 I_{TOTAL} &= I_{DDA} + I_{DDD} + I_{LED} + I_{STEPPER\ MOTOR} \\
 &= 3mA + 1.2mA + 4mA + 12.5mA \\
 &= 20.7mA
 \end{aligned}$$

The current required during the system start-up charge time of C₁₈ represents a discharge current for this capacitor. Thus, the total system current for the circuit cannot be larger than the value of I_{AVERAGE} given in Table 8. When this situation arises, the charge on C₁₈ will completely turn on the regulator, possibly causing system brown-out conditions to occur. The following scope plot shows the effect of sagging line voltage, the current demands and the effect on the power supply circuit.

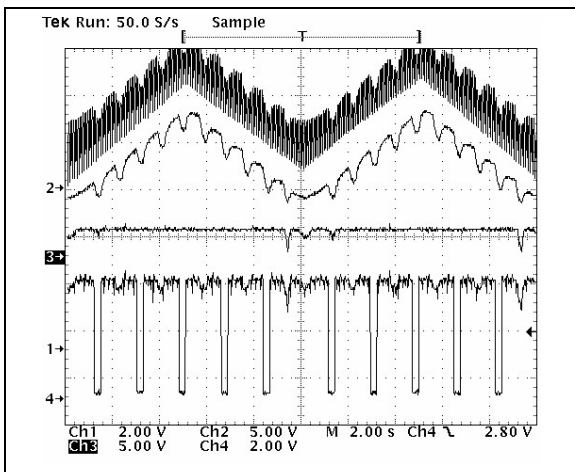


FIGURE 12: Waveforms of power showing increased current demand effects on the power supply.

Channel 2 is the unregulated voltage at the input to the LM7805 (or the voltage on C₁₈). The triangle waveform was generated to simulate a sagging line voltage across a very long period (in this case, a period of

20 seconds). The pulses on channel 4 are the mechanical counter pulses on F₀ from the MCP3905A. The channel 2 ripple shows the effect of the 0,1 pulses to turn the mechanical counter on the unregulated C₁₈ voltage. The sags in this voltage coincide with the increased current demand of the system (in this case, the 0,1 pulses).

MCP3905A POWER-ON RESET (POR)

The MCP3905A has a POR delay of 1 second. During brown-out, in order to prevent false pulses, this 1 second delay is reset and restarted (i.e., any pulses out of the MCP3905A are delayed until the POR timer is complete).

This POR timing is shown in Figure 13.

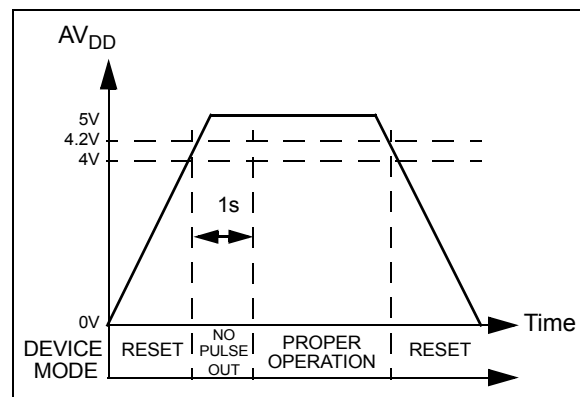


FIGURE 13: Power-on Reset Operation.

NO-LOAD THRESHOLD

Energy meter standards state that when there is no load connected to an energy meter, the device should not slowly register energy. For the MCP3905A, this is achieved by placing a threshold-detection circuit below the minimum load requirement (start-up current). This is defined as 0.0015% of the full-scale output frequency set by pins f₂, f₁ and f₀. When testing the no-load, IEC62053 states that no more than one pulse should occur during the time (Δt) given here:

EQUATION 11:

$$\Delta t \geq \frac{600 \times 10^6}{kmU_n I_{max}} [min]$$

Where

- k = The meter constant in impulses per kilo-watt hour (imp/kWh)
- U_n = The reference voltage in volts
- I_{max} = The maximum current of the meter
- m = The number of measuring elements

For example, using the calibration output pulse with a meter constant of 3200 imp/kWh, a line voltage of 220V and a maximum current of 80A, the period is about 17 minutes.

STARTING CURRENT

The IEC standard also defines the smallest current that the meter shall be able to register energy for. The standard states that the meter “shall start and continue to register at the starting current values” defined. These current values are shown in [Table 9](#).

TABLE 9: STARTING CURRENT

Meter Type	Class 1	Class 2
Direct-connect Meters	0.004 I_B	0.005 I_b
Transformer Meters	0.002 I_n	0.003 I_n

As an example, for a Class 1 direct-connect meter with a base current of 10 amperes, the starting current would be 40 mA.

MCP3905A V_{REF} TEMPERATURE COEFFICIENT CURVE FITTING

The MCP3905A contains a very low-drift voltage reference. The band gap is specially designed to allow the signature bow across temperature to be trimmed-in to give the lowest drift across the industrial temperature range. [Figure 14](#) shows the effects of this curve fitting and how the low tempco specification of the voltage reference is achieved.

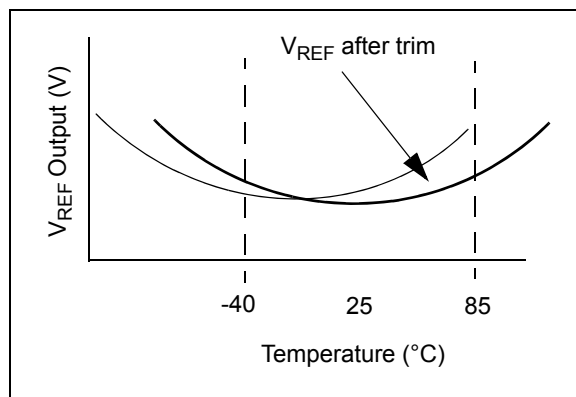


FIGURE 14: Band Gap Curve fitting.

When designing an energy meter, any drift associated with the voltage reference will effect energy meter readings and subsequent billing during winter months. [Figure 15](#) shows the results (i.e., MCP3905A voltage reference data using this curve-fitting technique).

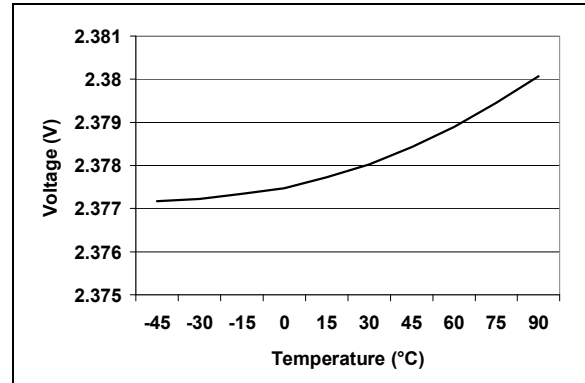


FIGURE 15: MCP3905A/06A Tempco Post Trim.

This device shows a 9.2 ppm/C temperature drift. The typical specification given for the MCP3905A/06A devices is 15 ppm/°C.

IMMUNITY TO EXTERNAL INFLUENCES – EMC

At the time of this writing, there were 16 ‘Influences’ described in the IEC62053 Class 1 and Class 2 Standard, Section 8.2. These influences to the meter are not meant to add additional percentage errors to the accuracy of the meter. Most of these influences pertain to variations in either the frequency content of the signal or variations in supply voltage or frequency that are diffused by the MCP3905A/06A PSRR, input bandwidth or sampling rate. The remaining influences deal with high frequencies (up to 2 GHz).

Electromagnetic RF fields of 30 kHz to 2 GHz, and conducted 150 kHz to 80 MHz, can exist in the application. However, the MCP3905A/06A cannot protect against these signals with internal circuitry. The anti-aliasing filter and series inductor used on the MCP3905A Energy Meter Reference Design PCB eliminates unwanted noise in this range. An EMC certificate follows in the **"Summary"**.

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SUMMARY

The following tables show testing results using the MCP3905A Energy Meter Reference Design. [Table 10](#) shows standard testing conditions. [Table 11](#) and [Table 12](#) show results from variation in line voltage and frequency. The EMC certificate shows compliance to tests outlined in IEC62053, IEC61036, IEC1036 and IEC687 energy meter standards.

TABLE 10: STANDARD POWER METER TESTING, NST-3500 POWER METER TESTING EQUIPMENT

Current	Power Factor (cos ϕ)	Permission Error Change (%)	Actual Error (%)		
			No. 111003502194	No.111003502181	No.111003502956
0.05I _B	1.0	±2.5	0.0	0.0	0.1
0.1I _B	1.0	±2.0	0.1	0.0	0.1
0.5I _B	1.0	±2.0	0.1	0.0	0.0
I _B	1.0	±2.0	0.1	0.0	0.0
0.5I _{max}	1.0	±2.0	0.1	0.1	0.0
I _{max}	1.0	±2.0	0.1	0.0	0.0
0.1I _B	0.5L	±2.5	0.0	0.0	0.1
0.2I _B	0.5L	±2.0	0.1	0.0	0.1
0.5I _B	0.5L	±2.0	0.1	0.0	0.1
I _B	0.5L	±2.0	0.1	0.0	0.1
0.5I _{max}	0.5L	±2.0	0.1	0.0	0.1
I _{max}	0.5L	±2.0	0.1	0.0	0.1

Note: Reference Voltage: 220V Current: 5(20)A Reference Frequency: 50 Hz
Temperature: 23°C Humidity: 45%

TABLE 11: FREQUENCY ERROR TESTING RESULTS, NST-3500 METER TESTING EQUIPMENT

Current	Power Factor (cosφ)	Frequency (Hz)	Permission Error Change (%)	Actual Error Changing (%)
				#1
0.05I _B	1.0	49	±0.5	0.065
I _B				0.046
I _{max}				0.000
0.1I _B	0.5L		±0.7	-0.016
I _B				-0.014
I _{max}				-0.034
0.05I _B	1.0	51	±0.5	0.121
I _B				0.067
I _{max}				0.054
0.1I _b	0.5L		±0.7	0.056
I _B				0.042
I _{max}				0.026

Note: Reference Voltage: 220V Current (I_B): 10A Reference Frequency: 50 Hz
Temperature: 23°C Humidity: 45%

TABLE 12: VOLTAGE ERROR TESTING RESULTS, NST-3500 METER TESTING EQUIPMENT

Current	Power Factor (cosφ)	Voltage (V)	Permission Error Change (%)	Actual Error Changing(%)
				#1
0.05I _B	1.0	198	±0.7	-0.071
I _B				0.063
I _{max}				0.032
0.1I _B	0.5L		±1.0	0.052
I _B				0.030
I _{max}				0.016
0.05I _B	1.0	242	±0.7	-0.017
I _B				0.000
I _{max}				-0.030
0.1I _B	0.5L		±1.0	-0.015
I _B				-0.014
I _{max}				-0.047

Note: Reference Voltage: 220V Current (I_B): 10A Reference Frequency: 50 Hz
Temperature: 23°C Humidity: 45%

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- [4] *"IEC 62053 Standard"*, IEC Commission, ©2003.

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