Designing a wireless node-based energy meter to measure both real and reactive power

Timo Engelgeer
(s0150061)

Electrical Engineering, Mathematics and Computer Science (EEMCS)
Computer Architecture for Embedded Systems (CAES)

MENTOR
Dr. Ir. V. Bakker

EXAMINATION COMMITTEE
Prof. Dr. Ir. G.J.M. Smit
Ir. E. Molenkamp
Ir. J. Scholten

16 July 2017

UNIVERSITY OF TWENTE.

1 Introduction

At the energy group of CEAS at University of Twente they research energy management. For efficient use of energy it is not only important to just use energy efficient devices but also how you use those devices. Among other things a major research area of the energy group is time-shifting of the energy usage to reduce the peak power in the grid. For this research area is useful to not only know the static loads but also the real time load. This way the system can make real time decisions and compensate for peak loads.

The currently used meter from Plugwise and other commercial alternatives only measure the real power a device uses but ignore the apparent and reactive power. Because the reactive power is just transported back and forth between the device and the source it causes an extra current which is transported twice. So although the reactive power does not cause power consumption in a device it does needs to be transported over through the grid and cause extra losses in the transport of the power. For power efficiency it is useful to also be able to measure this reactive power. This is explained in more detail in the chapter Theory.

The situation described above illustrates why the energy group is interested in a wireless energy meter that can measure both real and reactive (or apparent) power. The main objective of this bachelor thesis is therefor:

"Design a wireless energy meter which can measure real power, reactive power and voltage."

At the start the requirements were:

- Measures active and reactive power
- Measure voltage
- Wireless communication
- Able to use more than one wireless meter
- Updates every second or at significant changes

These objectives are still pretty broad so they need to be specified into more detail. This is discussed in more detail in the chapter Design under Design requirements.

2   Contents

3 Theory

To be able to design a power meter it is important to understand the theory behind it. This chapter gives a brief overview of the theory of real, reactive an apparent power and the basics for measuring current in the digital domain.

3.1 Real power

Calculating the power in a DC system is pretty straight forward. The power dissipated in the device (or delivered to the system when negative) is defined as the work done in the device. It is the product of the effort and flow in the device thus in the case of electricity is given by:

$$P = U \cdot I \tag{1}$$

With P is power in Watt, U is voltage in Volts and I in Ampere. This also holds for an AC system but voltage and current are time dependent so power becomes time dependent as well. So the instantaneous real power at each moment is given by:

$$p(t) = u(t) \cdot i(t) \tag{2}$$

With $p(t)$ the instantaneous real power at time t, $u(t)$ the voltage at time t and $i(t)$ the current at time t.

But if we want the dissipation in a device we are not interested in the instantaneous power it draws but the average power in time. We can calculate that with:

$$P_{avg} = \lim_{t \to \infty} \frac{1}{t} \int_0^t u(\tau) \cdot i(\tau) \, d\tau \tag{3}$$

With P_{avg} the average real power over time, $u(t)$ the voltage at time t and $i(t)$ the current at time t. P_{avg} is the real power the devices uses. Sometimes also called active power because it is the power that actively does the work (e.g. transformed into heat, light motion etc.).

If we assume sinusoidal voltage with period T feeding an inductive or capacitive load the current will be sinusoidal with period T as well. Or, more general, if the device properties are static in time and we apply a period signal with period T the current will be periodic with period T as well. This means we can determine the average power over one period.

$$P_{avg} = \frac{1}{T} \int_{t_0}^{t_0+T} u(\tau) \cdot i(\tau) \, d\tau \tag{4}$$

With T the period of the signal[1] and t_0 an arbitrary point in time.

The voltage of the power grid is a periodic with a stable frequency of 50Hz[2] which makes it easy to determine the real average power 50 times a second.

Figure 1 shows the grid voltage and the difference between in-phase current (resistive load) and -60° out of phase current (inductive load) for a single period. Figure 2 shows the instantaneous and the

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[1] Although an integer multiple of the period can also be used for static systems.
[2] At least for European countries but 60Hz for America. To keep this thesis simple the European 50Hz is assumed but the same theory can be applied to 60Hz.

average power for the same signals. This shows clearly that although the current is the same (only shifted in time) and thus the cable losses are the same, the average (real) power is half as large[3].
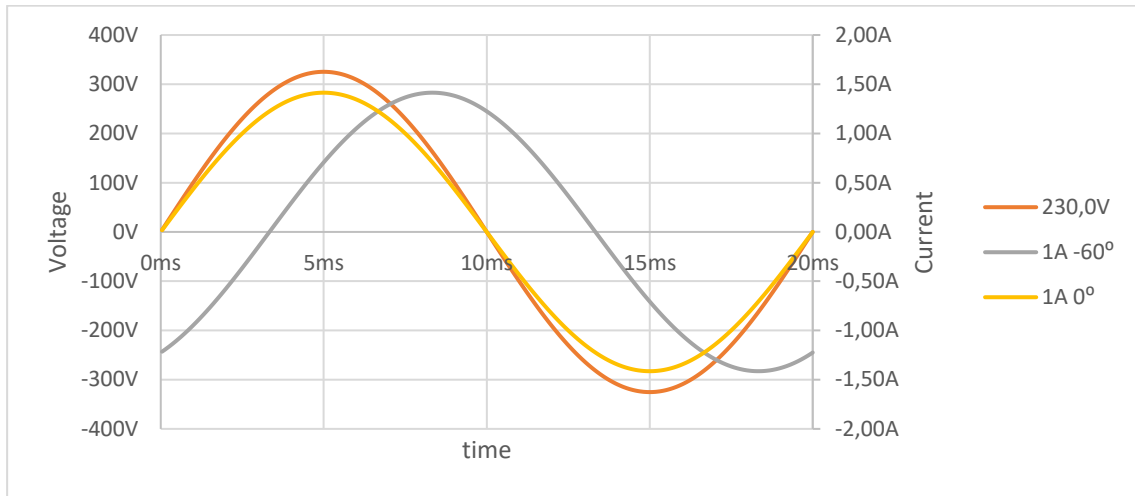
Figure 1: In phase current (resistive load) and -60° out of phase current (inductive load)
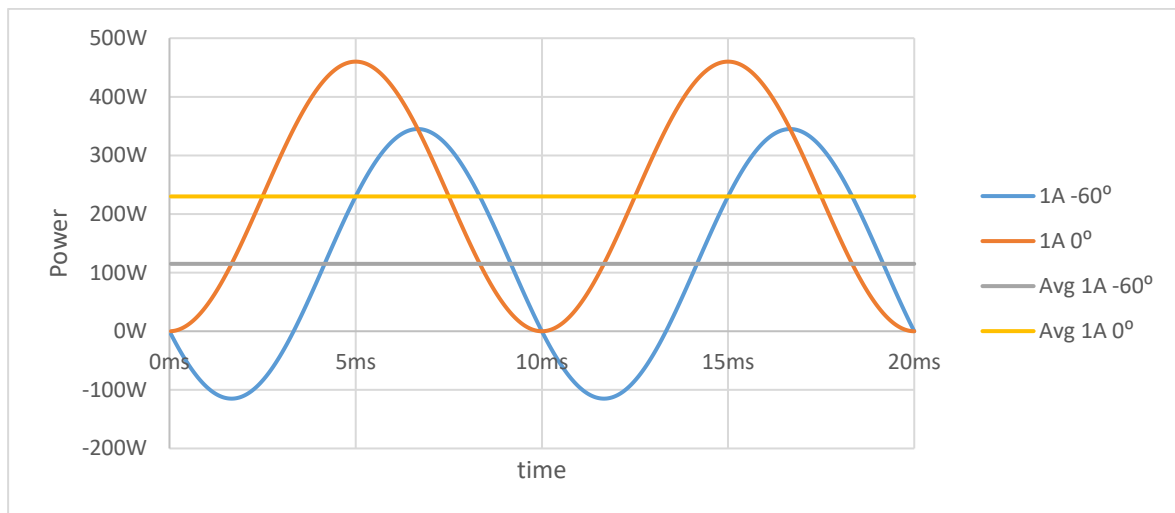
Figure 2: Instantaneous and average power for in phase and -60° out of phase current of 1A

For a resistive load its current is in phase with the voltage. We can describe the net voltage and current to a resistive load in time with

$$u(t) = U_{max} \cdot \sin\left(t \cdot 2 \cdot \pi \cdot \frac{1}{T}\right)$$

(5)

$$i(i) = I_{max} \cdot \sin\left(t \cdot 2 \cdot \pi \cdot \frac{1}{T}\right)$$

(6)

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[3] 230W is delivered in case of the in-phase current and only 115W is delivered in the case of the 60° shifted current.

With RMS values of

$$U_{RMS} = \sqrt{\frac{1}{T} \int_{t_0}^{t_0+T} u(\tau) \, d\tau} = U_{max} \cdot \sqrt{\frac{1}{T} \int_{t_0}^{t_0+T} \sin\left(\tau \cdot 2 \cdot \pi \cdot \frac{1}{T}\right) d\tau} = U_{max} \cdot \frac{1}{\sqrt{2}} \quad (7)$$

$$I_{RMS} = \sqrt{\frac{1}{T} \int_{t_0}^{t_0+T} i(\tau) \, d\tau} = I_{max} \cdot \sqrt{\frac{1}{T} \int_{t_0}^{t_0+T} \sin\left(\tau \cdot 2 \cdot \pi \cdot \frac{1}{T}\right) d\tau} = I_{max} \cdot \frac{1}{\sqrt{2}} \quad (8)$$

If we use the time representation of $u(t)$ and $i(t)$, (5) and (6), to calculate the average power we get

$$P_{avg} = U_{max} \cdot I_{max} \cdot \frac{1}{T} \cdot \int_{t_0}^{t_0+T} \sin\left(\tau \cdot 2 \cdot \pi \cdot \frac{1}{T}\right)^2 d\tau \quad (9)$$

Solving (9) gives

$$P_{avg} = \frac{1}{2} \cdot U_{max} \cdot I_{max} = \frac{1}{\sqrt{2}} \cdot U_{max} \cdot \frac{1}{\sqrt{2}} \cdot I_{max} = U_{RMS} \cdot I_{RMS} \quad (10)$$

So for a pure resistive load where voltage and current are in phase the real power also follows from the same definition as with DC. This doesn't come as a surprise because the RMS value of signal is defined as:

> *"The RMS value of a signal is the amplitude of a DC signal which will transfer the same energy like the applied signal into an ohmic resistor in the same time."*
> *(Jäckle, 2016)*

This makes the RMS values of voltage and current convenient but it only holds for resistive loads. The full time integral of the instantaneous power holds for all (periodic) voltage and current.

3.2 Apparent power

However, if voltage and current are out of phase the current flowing is higher than one might expect when looking at the average power and voltage. That is because for a part of the period voltage and current are out of phase so the instantaneous power is negative. This means the source does not supply power but it absorbs power back. A part of the power that the supply delivers is sourced back to the source so it decreases the average real power. When voltage and current are out of phase the simple product of RMS voltage and RMS current gives a power rating that is too high. But it appears to be the power flowing if looked at (RMS) voltage and current alone.

So for all loads the apparent power is given by

$$S = U_{RMS} \cdot I_{RMS} \quad (11)$$

With S the apparent power. Because it is also a product of voltage and current its unit should also be Watt. But to differentiate apparent power from real power the unit volt-ampere (VA) is used.

This leaves a gap between the real power and apparent power called the reactive power. The reactive power is a measurement of how many power is transported back and forth between source and device without being transformed into work. Because real power and apparent power are vectors the remaining part is calculated with

$$Q = \sqrt{S^2 - P^2} \quad (12)$$

With Q the reactive power. Again, the unit is also Watt but is denoted as volt-ampere reactive (var) to differentiate it from real and apparent power.

A graphical representation of how real power, apparent power, reactive power and phase angle relate is shown in Figure 3. This is called the power triangle. The real power is expressed on the horizontal axis. Reactive power is on the vertical axis or the imaginary axis. It is called 'imaginary' because it is the part of the AC power that does not do real work. S shows the apparent power or combined with phase angle ϕ the complex power of the device. Now relation (12) can be easily seen.
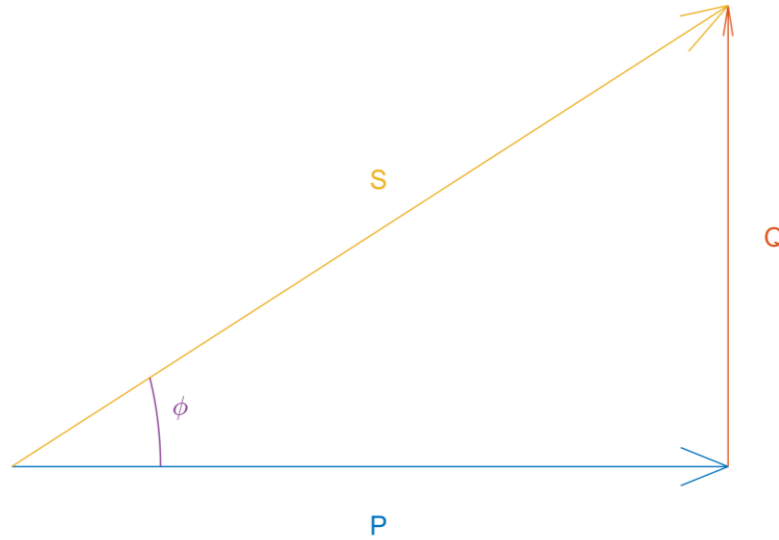

Figure 3: The power triangle for phase angle ϕ, real power on the X-axis and reactive power on the Y-axis

3.3   Power factor

The ratio between real power and apparent power is a measure of the amount of power which is used by the device as work and the amount of power that is only transported back and forth. This is called the power factor and has no unit. It is given by

$$PF = \frac{P}{S} \tag{13}$$

Because the real power is the same as the apparent power (but might be negative) when maximum power transfer is reached the power factor gives a ratio between -1 and 1. When the power factor is negative it means the power transfer is from device to source. This case not covered by this thesis because we are only interested in measuring net powered devices. This gives a power factor between 0 and 1. A power factor of 1 indicates all the transported power is used as work as in the case of a resistive load but also for balanced reactive loads. A power factor of 0 indicates we only transfer power back and forth and do no real work in the device.

Note that the definition of real power, apparent power and power factor holds for all periodic signals, not only sinusoidal signals. But if we assume a pure sinusoidal voltage into a time static linear load it can represent voltage and current with

$$u(t) = U_{max} \cdot \sin\left(t \cdot 2 \cdot \pi \cdot \frac{1}{T}\right) \tag{14}$$

$$i(i) = I_{max} \cdot \sin\left(t \cdot 2 \cdot \pi \cdot \frac{1}{T} + \varphi\right) \tag{15}$$

Where the current is phase shifted by φ compared to the voltage. A positive φ means the current is leading the voltage (capacitive load) and a negative φ means the current is lagging the voltage (inductive load). If the real power (4) is solved for (14) and (15) we get

$$P_{avg} = \frac{1}{2} \cdot U_{max} \cdot I_{max} \cdot \cos(\varphi) = U_{RMS} \cdot I_{RMS} \cdot \cos(\varphi) \tag{16}$$

If we use (16) to calculate the power factor as in (13) we get

$$PF = \frac{P}{S} = \frac{U_{RMS} \cdot I_{RMS} \cdot \cos(\varphi)}{U_{RMS} \cdot I_{RMS}} = \cos(\varphi) \tag{17}$$

So in an ideal world with a clean sinusoidal voltage and a linear device the power factor is given by the phase angle between voltage and current which is fully determined by the (reactance of the) device.

Also note that the power factor gives no indication if the load is leading or lagging in current. The cosine in (17) makes that easy to see. It is also easy to see that a fully capacitive or inductive load, with a phase angle of 90°, corresponds to a real power of 0. Which is what to expect because an ideal capacitor of inductor only stores energy and does not dissipate energy. Hence, no work is done.

Power lines have losses due to their conductivity and the current they need to carry. They form a resistive path (thus induce losses) to the device. Combining Ohm's law[4] and the formula for power to a resistive load we get for the losses in the power lines

$$P = U \cdot I = (I \cdot R) \cdot I = I^2 \cdot R \tag{18}$$

So the losses in the power line are proportional to the current in them. Which is in turn proportional to the apparent power, not the real power, because the real power only contains the current that is used to do real work. This makes the power factor a good measure of how efficient the transported power is used. Another way to look at it is that it is a measure of the current in the power lines to do a certain amount of work. By increasing the power factor the same amount of real power can be transported (so same amount of work can be done) but with lower current in the power lines thus reducing the losses in the power lines.

3.4   Measuring real power and apparent power

Measuring nowadays means going digital. The first aspect that comes to mind is the quantization of the signal levels. The resolution of the ADC[5] determines the resolution we can get for the power and voltage. We cannot measure power directly so we have to stick to voltage and current. The bit depth of the power will be the sum of the voltage and current bit depth because it is the product of those two. Because resolution was not specified at the start of the project a table with arbitrary values is shown in Figure 4. Here we can already see that more than an 8-bit ADC is needed if we want to be accurate within a Volt for the voltage.

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[4]As a reminder, Ohm's law states $I = \frac{U}{R}$

[5] Analog-to-Digital Converter

Case / bit depth	325V[6]	650V[7]	700V[8]	2A	4A	10A	20A
8-bit	1,27V	2,54V	2,73V	7,81mA	15,6mA	39,0mA	78,1mA
10-bit	317mV	635mV	684mV	1,95mA	3,91mA	9,77mA	19,5mA
12-bit	79mV	159mV	171mV	0,49mA	0,98mA	2,44mA	4,88mA
16-bit	5mV	9,9mVmV	10,7mV	0,03mA	0,06mA	0,15mA	0,31mA

Figure 4: Resolution per LSB for different signal ranges

The next issue when going to digital is the sampling. The time continuous math of (4) becomes (19) when the signal is sampled.

$$P_{avg} = \frac{1}{N} \sum_{n=n_0}^{n_0+N} u(n) \cdot i(n) \tag{19}$$

With n_0 an arbitrary start sample, $u(n)$ and $i(n)$ the time discrete values of voltage and current and N the (integer) number of samples per signal period which is given by (20).

$$N = \frac{f_{sample}}{f_{signal}} \tag{20}$$

With f_{sampel} the sample frequency and f_{signal} the frequency of the measured signal, 50Hz in this case.

But this also shows the next issue. The number of samples depends on both the signal frequency and the sample frequency. In an ideal world we could simply pick a sample frequency that is N times higher than the signal frequency and end up with perfect sampling. But although the frequency of the signal might be pretty stable and accurate, because it is the grid frequency, the sample frequency probably is not because it is driven from a local clock source. This gives the need to synchronize the sampling to the signal frequency. An easy and stable reference for the synchronization is the zero crossing of the signal. This time information can be used to adjust the sampling frequency.

The processor power is also something to take in consideration. The computation must be able to keep up with the sampling frequency. The first aspect that effects the speed is the ADC bit depth compared to the microcontroller register width. If the register width is smaller than the ADC bit depth that means the microcontroller has extra overhead doing the calculations. But an even more important consideration is what instructions, or more precise, what mathematical operations, the microcontroller natively can do. A lot of 8-bit microcontrollers do not have a hardware multiplication or division operator. So multiplication and division is done in software at the cost of more clock cycles. For example, a 16-bit times 16-bit signed multiplication on an 8-bit AVR takes up 218 clock cycles (Atmel Corporation, 2009). That does not seem like a lot compared to a clock speed of 16MHz but if we sample with 3,2kHz (50Hz signal, 64 samples per period) that only leaves 5000 clock cycles per measurement to do all the math.

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[6] Peak voltage 230VAC
[7] Peak to peak voltage 230VAC
[8] 230VAC + allowed +5%, rounded up

4 Design

This chapter will explicate the design of the energy meter and will explain the design choices that are made in the process. These choices are all driven by the theory of chapter 3 and further explained here.

4.1 Design requirements

In order to come to a design it is required to extend the initial design requirements because some leave room for interpretation. In this chapter more precise requirements are made and the choices explained.

4.1.1 Measure instantaneous voltage and current at the same time

To measure active and reactive power it is required to measure both instantaneous voltage and current at the same time. From this real power, apparent power, reactive power, RMS voltage and RMS current can be calculated. So this satisfies the initial requirements of "Measures active and reactive power" and "Measure voltage". To do this, two input channels are needed which need to be sampled simultaneously.

4.1.2 Sample frequency

For the sampling the most important part is to keep it in sync with the main frequency. Figure 5 shows the error because of the sample frequency being deviated. So if the sample frequency is 10% slower it will sample part of the new period which creates an error of more than 3%. Or the other way around, if sampled 10% too fast it will not sample the full period which gives an error of around 4%. In both cases it is assumed sampling will start after a zero crossing of the signal and a fixed number of samples, 64 in this case.
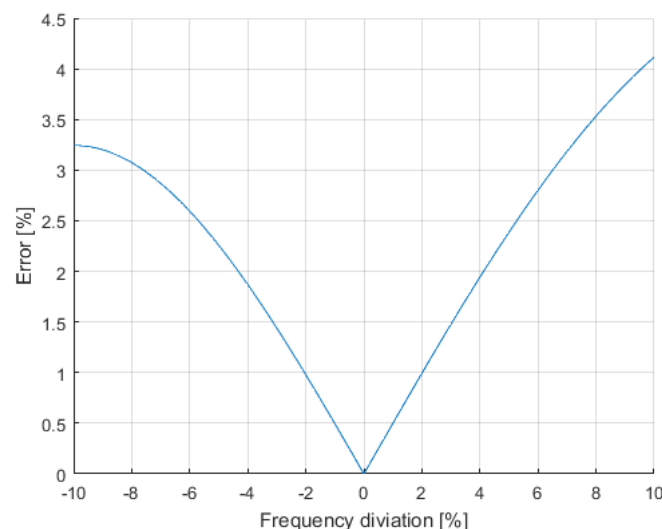
Figure 5: Error from sample frequency deviation of a sinusoidal (50Hz). Nominal frequency of 3200Hz (64 samples/period)

In Figure 5 a sample frequency of 64 samples per period is assumed. But for sampling a sinusoidal signal the number of samples does not really matter as long as it satisfies the Nyquist frequency and, as shown in Figure 5, sampled evenly over the period. But for that, the sinusoidal needs to be sampled in the peaks which makes it harder to synchronize it with the zero crossing. That is easy to fix by taking at least 4 samples per period.

In order to be able to measure non-sinusoidal signals, a higher number of samples is needed as well. It makes sense to make this number a power of two in order to make the division on a

microcontroller easy because it can just use the shift operation. Devices like SMPS[9] will generate non-sinusoidal signals which gives harmonics of a higher order in the signal. With 64 samples per period this will result in a sample frequency of 3200Hz thus according to Nyquist, is possible to sample signals up to 1600Hz. But like stated above, perfect alignment is needed for that. If the minimum amount of samples is increased to 8, a frequency of 400Hz can be sampled correctly. This is the 8th harmonics of the grid frequency. Harmonics play a bigger role in current then in voltage because the voltage in the grid is heavily regulated. The harmonics of the current however depend on the load. However, harmonics in current are regulated and for harmonics greater than the 8th it should be less than 0,5mA/W (Key & Lai, 1997). So 64 samples per period, or 3200Hz, will give an accurate reading for normal household devices including SMPS. But note, the sample frequency of 3200Hz is synchronized to the grid frequency so it will vary with it to keep the same numbers of samples per period.

4.1.3 Scale and resolution

The grid power in Europe is 230V AC but it has a tolerance of +/- 10%. This means the voltage may be as high as 253V AC. But that is the AC voltage, the peak voltage is $\sqrt{2}$ higher at 358V. And we have the same on the negative side so the total range of the mains voltage is 716V. So a total range of 750V is desirable.

The resolution of the voltage ADC is less of an issue because of the full swing through the range. This makes it change steps very fast and because of the large nature of the voltage compared to the current this doesn't affect the power accuracy much. Also for the zero crossing this is not a problem because the rate of change of the voltage around the zero crossing is given by the derivative of the grid voltage, which is given in (14) and results in (21).

$$u'(t) = U_{max} \cdot \sin\left(t \cdot 2 \cdot \pi \cdot \frac{1}{T}\right)\frac{d}{dt} = U_{max} \cdot 2 \cdot \pi \cdot \frac{1}{T} \cdot \cos\left(t \cdot 2 \cdot \pi \cdot \frac{1}{T}\right) \tag{21}$$

Which solved for mains voltage around the zero crossing gives (22).

$$u'(t) = 325 \cdot 2 \cdot \pi \cdot 50,0 \cdot \cos(0 \cdot 2 \cdot \pi \cdot 50,0) = 325 \cdot 2 \cdot \pi \cdot 50,0 = 102kV/s \tag{22}$$

At a rate of 102kV/s around the zero crossing a step of one LSB (750V full range at 8-bit,) just takes 28,7μs which is a lot less than one sample period of 316μs if sampled at 3200Hz (64 samples per period). So even at 8-bit a decisive zero cross detection is possible within one sample. This makes it possible to, even with an 8-bit ADC, synchronize the sampling with the grid frequency which, as shown in section Sample frequency, minimizes the error. So the zero cross detection is not a factor for the resolution of the voltage ADC, even at higher sample frequencies.

The range for the meter was not defined to start with. But it would be reasonable to be able to measure household items. For this, the scale of the current is set at 10A peak. This to accommodate peak currents in the measurements of non-sinusoidal currents. This will allow to measure sinusoidal currents up to 7,1A or roughly 1600W of resistive load. The total scale for the current therefor is 20A.

The power resolution was also not set. But in this world of LED lighting and other low energy devices it would be desirable to have a power resolution of at least 1,0W. That results in a RMS current of 4,3mA which has a peak current of 6,1mA. But to measure it correctly over the span of a period it is necessary to have a LSB of at least an order of magnitude lower than that. Figure 4 shows that a 16-bit resolution is necessary for that.

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[9] Switched-Mode Power Supply

To confirm that, a MatLab model is made in which quantized power calculation is compared to calculated power of in-phase or phase shifted linear signals. From Figure 4 it was already clear that 1W cannot be measured with a 8-bit ADC on a scale of 10A (full scale 20A) because the ADC step size of 78,1mA/LSB is more than an order of magnitude bigger than the 6,1mA peak of a 1W in-phase load. But even if the range is limited 8-bit is not enough for Watt-accurate readings on a useable range.

Figure 6 shows the power of a 1W in-phase load when an 8-bit ADC is used on 1A full scale. Figure 7 shows the quantized voltage and current of the same load. The tracking because of the voltage is sufficient but the current steps are clearly noticeable. The steps in de quantized power match the steps of the quantized current. This gives an error of 3,35% in the quantized power calculation and a 7,68% error in the apparent power calculation. If the signal is given a phase shift of 60° then the error of the real power becomes 3,71% and the error in the apparent power becomes 5,31%. Because the reactive power is calculated from that it's error is even bigger at 5,84%.
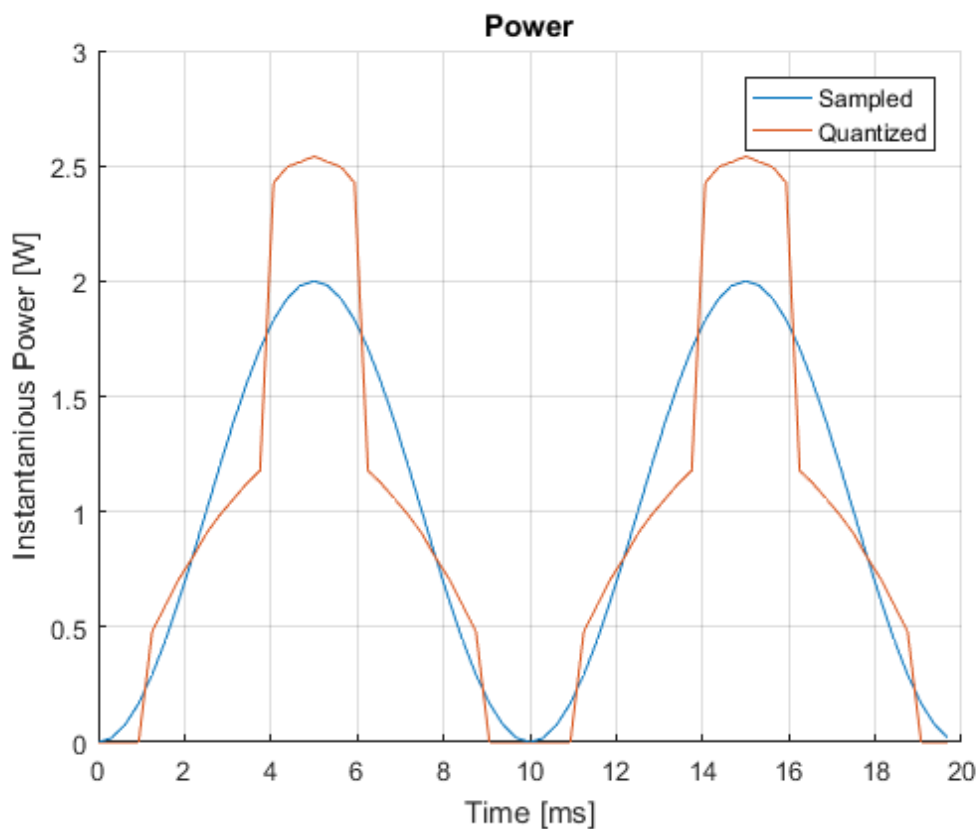
Figure 6: Power sampled and quantized (1W, 0VA at 8-bit for 1A full range, 8-bit for 750V range)
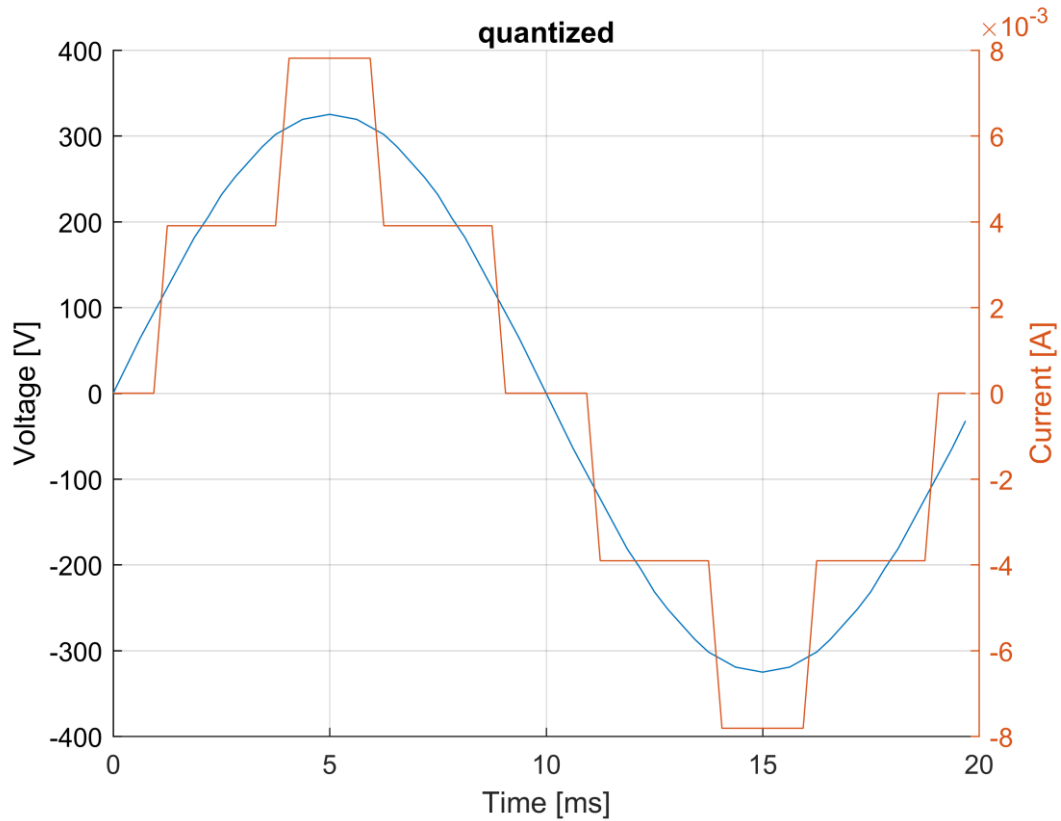
Figure 7: Voltage and current quantized (1W, 0VA at 8-bit for 1A full range, 8-bit for 750V range)

Figure 8 shows the error over the full 500mA range of the same setup (1A full range, 8-bit, 60°).
Figure 9 shows it zoomed in on the lower current part (part with the larger errors) of the same load
as in Figure 8 but with real power, apparent power, reactive power and power factor combined. The
error is as expected alternating with the LSB. This gives a pretty big fluctuation in the error in the
Watt range (lower range). The table in Figure 10 shows the mean error and the RMS error over the
full 500mA range. So even with reduced range an 8-bit ADC will just not give the desire accuracy for
measuring power.

Figure 8: Error over the full current input range, 1A full range, 8-bit, 60°
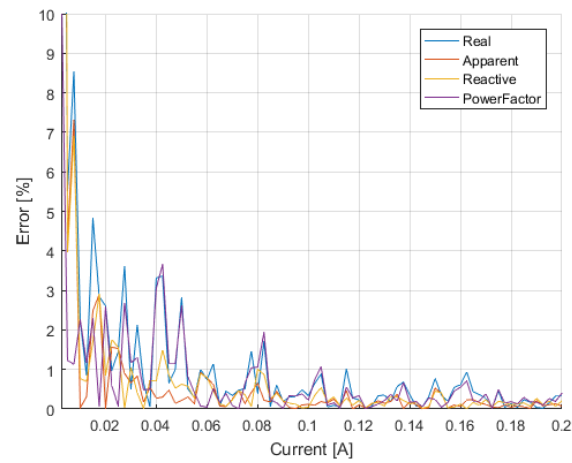
Figure 9: Error over lower rang, 1A full range, 8-bit, 60°

	Mean Error	RMS Error
Real	0,595%	2,36%
Apparent	0,460%	3,34%
Reactive	0,516%	3,68%
Power Factor	0,373%	1,01%

Figure 10: Mean and RMS error over full range, 1A full range, 8-bit, 60°

If the bit depth of the ADC is increased to 16-bit and used over the desired range of 10A (20A full range) the error is reduced significantly. Figure 11 shows the error over the full range (still 60° shift) and Figure 12 shows the error in the lower part. With 16-bit ADC the error is almost reduced to zero, even at the desired range which is 20 times larger than the range for the 8-bit test.
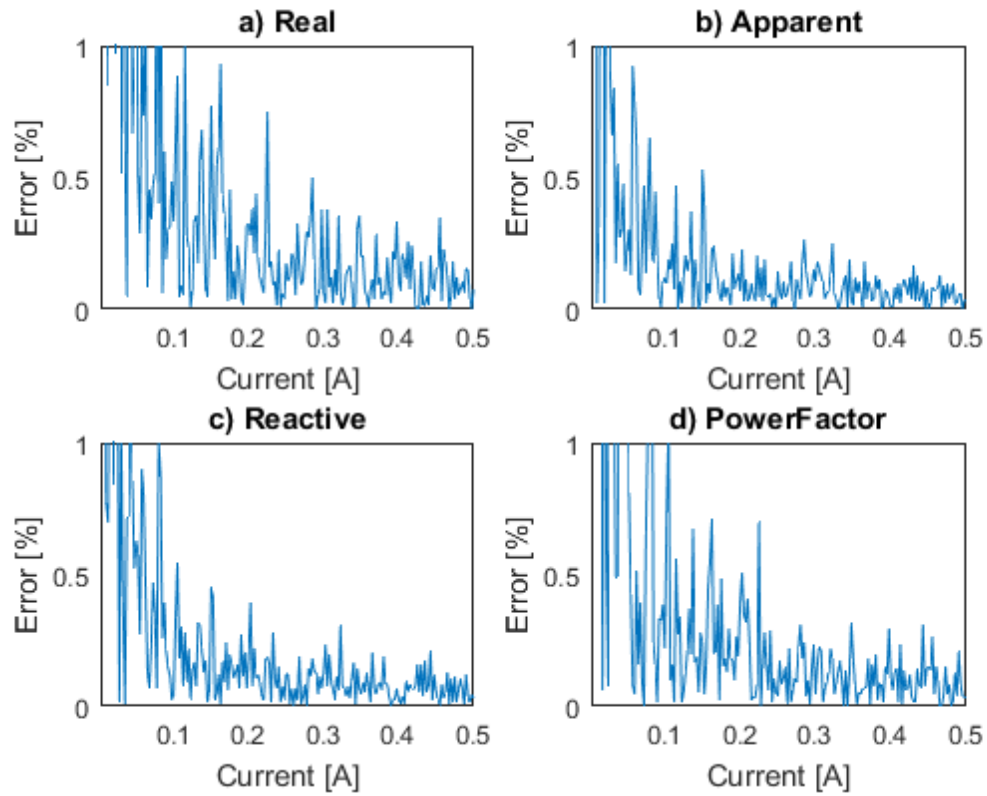

Figure 11: Error over the full current input range, 20A full range, 16-bit, 60°

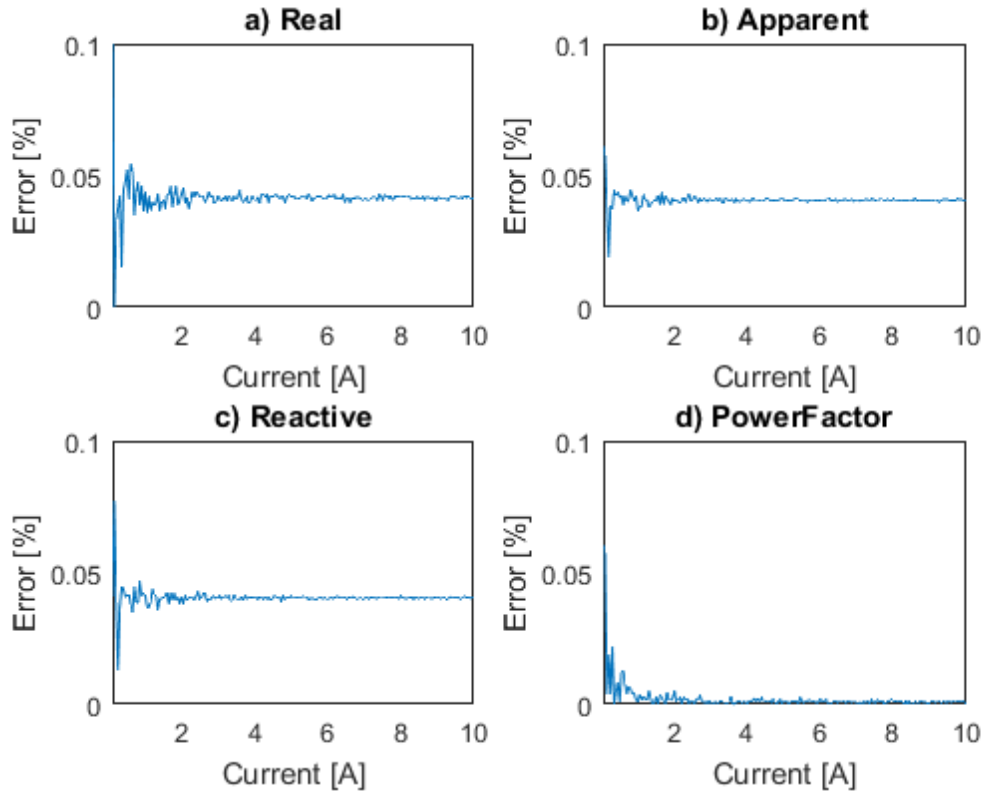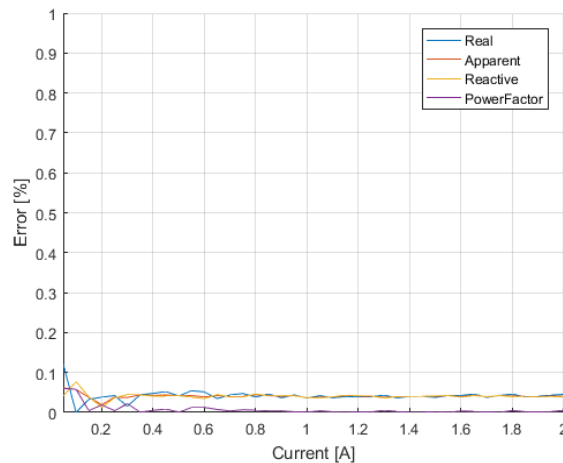Figure 12: Error over lower rang, 20A full range, 16-bit, 60°

	Mean Error	RMS Error
Real	0,042%	0,042%
Apparent	0,041%	0,041%
Reactive	0,040%	0,041%
Power Factor	0,002%	0,007%

Figure 13: Mean and RMS error over full range, 20A full range, 16-bit, 60°

There is no difference in computation time between 16-bit and 10-bit on a microcontroller. In both cases it is 2 words wide on an 8-bit microcontroller and just 1 word on a 32-bit microcontroller. That is why a 16-bit ADC is plotted. But as shown in Figure 14, even a 10-bit ADC will already reduce the error significantly and would suffice. So at least 10-bit for the current ADC is desirable.

	Mean Error	RMS Error
Real	0,146%	0,418%
Apparent	0,069%	0,118%
Reactive	0,089%	0,178%
Power Factor	0,124%	0,402%

Figure 14: Mean and RMS error over full range, 20A full range, 10-bit, 60°

4.1.4   Power supply

The power consumption of the meter should be kept to a minimum and should not consume more than 0,5W on average. To make the design easy a non-isolated power supply is desirable. This makes it possible to directly measure the grid voltage or a shunt resistor because the measuring circuit is already referenced to the grid. No isolation between the measuring circuit and microcontroller is needed because all communication is done wireless and no external devices will be connected. Not being isolated will form a safety hazard.

4.1.5   Wireless communication

The design requirements for the wireless communication were already clear from the start. Because of the power consumption requirement the wireless transceiver needs to be fairly low power. Or at least support a low power mode between transmissions to reduce overall power. This makes the use of Wi-Fi for example undesirable. But the range of the wireless meter needs to cover at least a normal house.

This gives the following design requirements:

- Able to use more than one wireless meter
- Able to update every second or at significant changes (>1%, max one update per second)
- Operates in a free frequency band
- Fairly low power
- Range: within a normal house (at least 10m and a wall)

4.2   Hardware and software design

This paragraph shows the used hardware and software and explains the reason why this is chosen based on the design requirements. The basic design is shown in Figure 15. Each of these blocks are covered in paragraph 4.2 and chapter 5 which specifies the implementation.
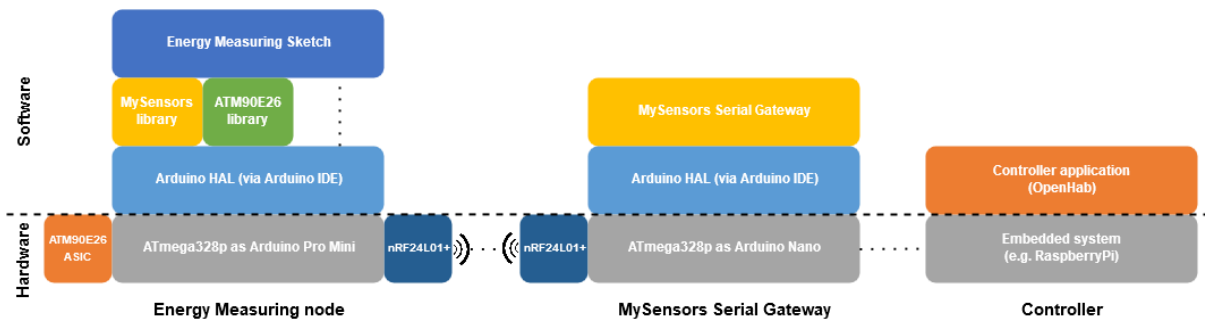
Figure 15: Basic structure of the wireless energy meter including receiver and controller

4.2.1 Measuring

Measuring the current is the biggest challenge in the energy meter. As shown in the design requirements (page 11) an 8-bit ADC would not give the desired resolution even on a scaled down range. This would force the use of an ADC word length of 16-bits. This would increase the load for the calculations on an 8-bit microcontroller drastically. And because timing the sampling is critical it becomes really hard to do all the calculation and keep the sampling accurate.

A solution would have been to use a 32-bit microcontroller. Not only gives this an advantage because of the bigger word length, because the ADC results fit in one word, but also because 32-bit microcontrollers are more advanced. Even simple 32-bit microcontrollers that are competitive in price run at a higher clock speed and feature hardware multiplication and even hardware division. This would really speed up the calculatio, leaving more room to time the sampling and to the wireless transmission.

But because of the lack of experience with 32-bit microcontrollers it was decided to use a dedicated energy meter ASIC[10]. Because the research to use a 32-bit microcontroller and the implementation on it would take up to much time and would fall out of scope of a bachelor thesis. Most big chip manufacturers make energy meter IC's. The ATM90E26 from Microchip (former Atmel chip) was chosen for the design. This is a relative simple IC for single phase metering but it fits the needs. It does all the heavy lifting of the measuring voltage and current, synchronizing the sampling to the grid frequency and calculation of the power, both real and reactive. From a component standpoint it is cheap (less than one dollar), does not need much external components and is easy to interface via SPI, even with an 8-bit microcontroller. The ATM90E26 has a dynamic range of 1:5000 which makes it possible to get the desired power resolution. A downside is that we do not have full control over the sample data which makes it impossible to do more analyzes on it. But because that is outside the requirements of this thesis does not form a problem.

Using a power measuring ASIC instead of a measuring implementation fully on a microcontroller made it possible to realize a working energy meter in a short period without limiting the design requirements. From a cost perspective this is not a big issue because the ATM90E26 is a cheap device. But it does create a manufacturer dependency and might limits future expansion of the measurements, for example measuring the amount of harmonics. But because that is beyond the scope of this thesis that is not an issue.

An additional advantage of using an ASIC for the energy metering is that is automatically compliant with exciting and accepted measuring requirements. The ATM90E26 is compliant with the requirements of IEC62052-11, IEC62053-21 and IEC62053-23 for single phase Watt-hour and VAR-hour meters. Being compliant is not a design requirement for this energy meter but it does show the

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[10] Application-Specific Integrated Circuit

algorithms used by the ASIC are proven. Being compliant can also be an advantage if the design is use commercially in the future.

4.2.2   Wireless framework

For the wireless and data transmission the MySensors framework is chosen. This framework takes care of all the data transmission and can be linked to most home automation controllers like OpenHAB. Using a framework like MySensors was preferred over designing it from scratch because of the limited time for the assignment. The MySensors frameworks covers all requirements for the wireless transmission. It handles multiple nodes, even combined with other sensors. It is designed for home (automation) use. This covers not only the range requirement but also the frequency of transmission, it transmits in free radio bands and uses low power transmitters.

In addition to covering the initial requirements, the MySensors project is chosen because of the experience with the Arduino IDE and C++ in general. This was a great advantage because that made prototyping quick.

The MySensors framework also makes it possible to expand the implementation in the future. Not only does it allow for more energy metering modes, it also allows for different type of nodes. The nodes can be measured as well as controlled. It is also possible to expand the wireless communication with encryption. This is not used in this thesis because of time constrains and it is not a design requirement.

4.2.3   Wireless transmitters

The use of the MySensors framework does limit the possible transmitters that can be used. It has support for the RFM69 from HopeRF Electronic and the nRF24L01+ from Nordic Semiconductor. The RFM69 works transmits on 315MHz, 433MHz, 868Mhz and 915Mhz whereas the nRF24L01+ transmits on the 2,4GHZ band. So both transmitters are able to transmit on free bands (Agentschap Telecom, 2014; Minister van Economische Zaken, 2015). The nRF24L01+ is the most common, readily available and available in a ready builds module for prototyping. Because of this and the fact it meets the requirements for the wireless the nRF24L01+ is chosen.

Another obvious choice for the wireless transmission would have been Wi-Fi. This would also easily integrate with existing setups because most already use Ethernet. But there is deliberately chosen not to use Wi-Fi for two important reasons. First, there is significant overhead when transmitting the relatively small data from the energy meter. And second, Wi-Fi is not particular energy efficient. And because the energy meter is designed to run continuously, is powered transformer-less and is meant as a tool to save energy the energy consumption is important. The nRF24L01+ only uses 11mA when transmitting and only transmits in small bursts. (Nordic Semiconductor, 2008). For example, the ESP8266EX from Espressif (a common wireless SOC) uses 170mA while transmitting (Espressif, 2017). This makes the nRF24L01+ more suitable for the energy meter compared to Wi-Fi although operation in the same frequency band.

4.2.4   Microcontroller

The MySensors project is based around an Arduino IDE library and is tested to work with the Atmel ATmega328p (the Arduino default) and the Espressif ESP8266. Because it is largely just C++ it can work with every other Arduino compatible microcontroller (third party or native). In the design for the energy meter the ATmega328p is picked. There are a couple of advantages for this:

1. It is the most common microcontroller for MySensors so the framework is well tested for it.
2. It is the Arduino default so any other library is well tested as well.
3. A lot of experience with writing for and working with the microcontroller.

4. It is a high performance (for 8-bit), low power microcontroller.
5. Available as cheap and easy to use development board (Arduino Nano or Arduino Pro Mini).

Point 3 combined with point 5 makes it possible to make a prototype in the limited amount of time of the assignment and not have to spend a lot of time on research about the microcontroller. That said, it is all programmed in C++ with the Arduino IDE as Hardware Abstraction Layer (HAL). It should not be too complicated to port the code for this energy meter to any other capable microcontroller with a C++ capable compiler. Although using an Arduino IDE compatible microcontroller will make it considerably easier to do.

And because one of the design requirements is to make the energy meter fairly low power it makes sense not to go with an overpowered microcontroller. As point 4 makes clear the ATmega328p is a simple low power microcontroller but it has all the necessary features to interface with the energy metering ASIC and the MySensors wireless.

4.2.5   Wireless gateway

Having an energy meter that sends its data wireless is one thing, receiving it and use it in an application is another. To receive the data, the MySensors Serial Gateway is used which passes the data to a controller application. This is basically the same as a MySensors node (an ATmega328p and an nRF24L01+ transceiver) but combined with a serial to USB converter. It sends all the received data to the connected host computer over USB.

The MySensors Serial Gateway comes with plugins for a lot of home automation platforms including Domoticz, OpenHAB, Vera and Home Assistant. OpenHAB is used extensively in the energy group. Although it was not a requirement to make it compatible with OpenHAB it is an advantage to be able to use the sensor data easily within the energy group. That way the energy meter can be a substitute for the Plugwise with the addition of being able to measure reactive power in real time.

4.2.6   Schematics

Figure 16 shows the schematic for the Energy Metering node. The energy metering parts are based on the reference design of the ATM90E26 (Atmel, 2014) with exception of the neutral current sensing and the power supply. The neutral current sensing via a Current Transformer is not implemented because the redundancy and anti-tampering facilities are not required in this energy meter.
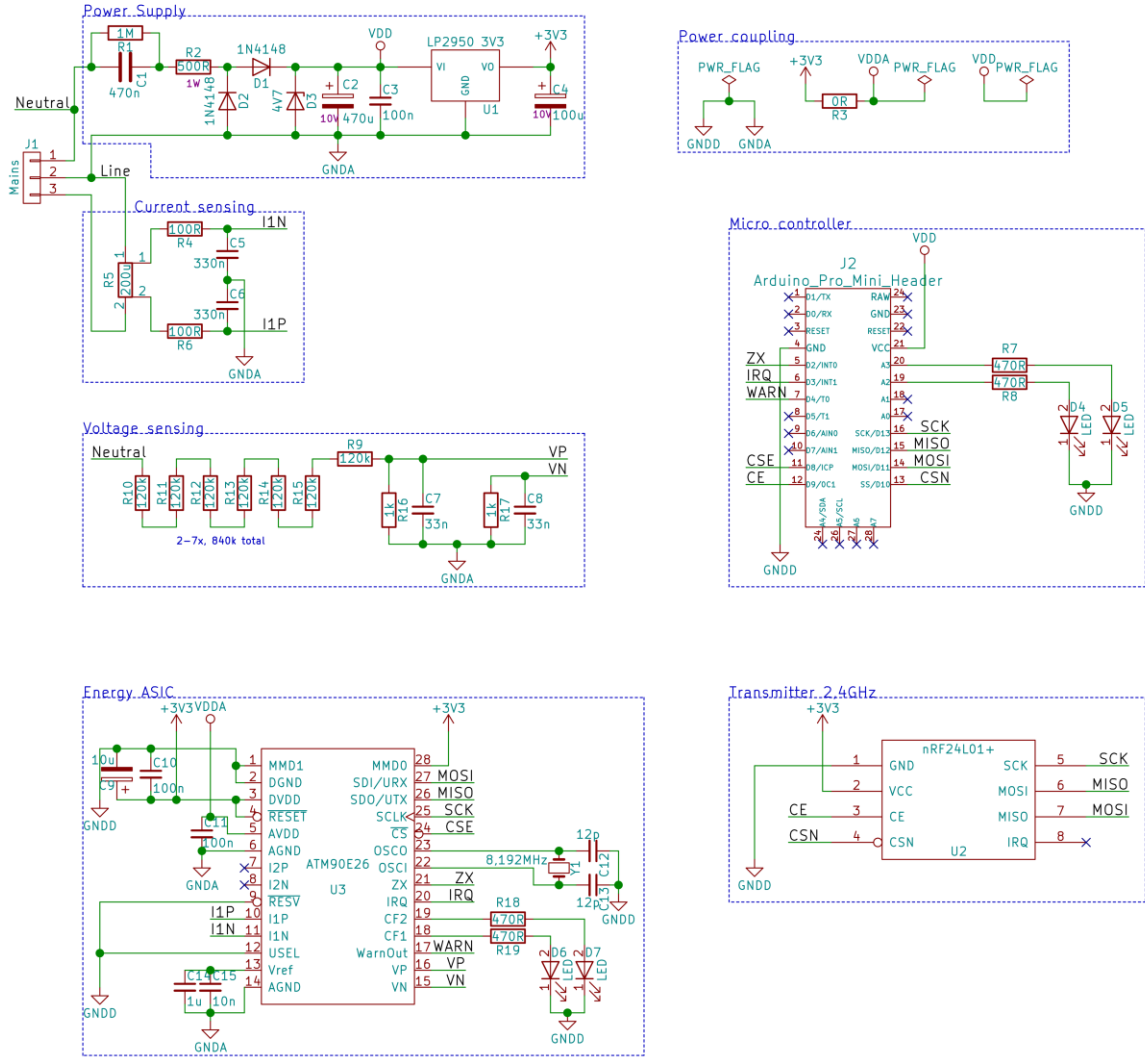
Figure 16: Schematic of the Energy Metering node

4.2.7 Power supply

Instead of a transformer a transformer-less capacitive dropper design is used. This is shown in Figure 16. This drops the mains voltage down to around maximum 4,7V with the use of the reactance of a capacitor and a Zener diode to limit the voltage (Williams, 2017). This is used to power the microcontroller but is fairly un-regulated in voltage. To power the Energy metering ASIC and the transceiver, which need 3,3V, a LP2950 Low Dropout, low cost voltage regulator is used. This is not a critical part and can be substituted with any other voltage regulator if the regulators guidelines about capacitance is followed. An advantage of this circuit is the low cost and small footprint. The disadvantage is that it has a low power factor. The current in de low voltage is the same as in the high voltage so it has a large apparent power. Only a small portion is used as real power.

C1 is the main part of the power supply and together with the Zener diode D3 forms a voltage divider. The reactance of C3 will mainly set the maximum current of the circuit. The reactance of a capacitor is given by

$$Z_C = \frac{1}{2 \cdot \pi \cdot f \cdot C} = \frac{1}{2 \cdot \pi \cdot 50Hz \cdot 470nF} = 6770\Omega \qquad (23)$$

R2 is there to limit the current in case it is powered on when the grid voltage does not pass true 0V. It would then try to instantaneous charge the capacitor resulting in a current spike. R2 limits this to

$$I_{inst} = \frac{U_{max} \cdot \sin(\omega)}{R_2} = \frac{230V \cdot \sqrt{2} \cdot \sin\left(\frac{1}{2}\pi\right)}{500\Omega} = 650mA \tag{24}$$

With $\omega = \frac{1}{2}\pi$ because that gives the biggest instantaneous voltage change which is the worst case. This is de worst case value and is only seen very brief but gives more than sufficient limiting not to trip fuses or breakers.

The current into the supply will be

$$I_{RMS} \approx \frac{U_{RMS}}{Z_C + R_2} = \frac{230V}{6770\Omega + 500\Omega} = 31,6mA \tag{25}$$

Here the voltage over the Zener is neglected because it's small compared to the grid voltage.

But only the positive half of the wave is used to charge the capacitor. The current is also sinusoidal and therefor (26) applies for the peak current and (27) for the average current if only the positive half is used.

$$I_{max} = I_{RMS} \cdot \sqrt{2} \tag{26}$$

$$I_{avg} = I_{max} \cdot \frac{1}{2\pi} \cdot \int_0^\pi \sin(\omega)\, d\omega = I_{max} \cdot \frac{1}{\pi} \tag{27}$$

Filled in (26) and (27) give

$$I_{avg} = I_{RMS} \cdot \frac{\sqrt{2}}{\pi} = 31,6mA \cdot \frac{\sqrt{2}}{\pi} = 14,2mA \tag{28}$$

That this supply can deliver on average. The transmitter makes up for the major part of the current in the circuit and is 13,5mA maximum.

4.2.8   Shunt and voltage divider

The ATM90E26 is a versatile chip which can calculate current, voltage and power up to $65,535A$, $65,535V$ and $\pm32,768kW$ with a dynamic range of 1:5000. The actual range of the sensor is determined by the shunt resistor and the voltage divider. But the closer the desired maximum signal is to the maximum input range of the chip, the better the resolution will be. But an overshoot of the input range of the chip will result in invalid readings and may even damage the chip.

The input range on the ADC pins of the chip is $600mV_{RMS}$ when the internal gain is set to 1. The voltage ADC has no programmable gain so is always 1. A voltage divider with a 1:431 ration is made which can consist of standard E24-series resistors of $1k\Omega$ and $430k\Omega$. This will result in $534mV_{RMS}$ when $230V_{RMS}$ is applied. With that ratio the maximum input voltage (which gives the full$600mV_{RMS}$ on the ADC input) is $259V_{RMS}$ which is more than the maximum allow grid voltage in Europe.

The input range of the current ADC is also $600mV_{RMS}$ when gain is 1. But the L-line current ADC has a programmable gain of 1, 4, 8, 16 or 24 which will reduce the input range by the same amount. Another aspect to consider is the heat generated in the shunt resistor. Not only to not stress the shunt but also to not waste a lot of energy in the metering process. Shunt resistors of 1W are common and 1W is a reasonable heat loss for the maximum current. But that rating is its maximum

rating under optimal circumstances so it is better to keep a safety margin. On the other hand, a larger shunt value will give less variance due to contact, PCB and solder resistance because they can easily be of the same order. Figure 17 shows the power dissipation in the shunt resistor at 5A and 10A (the desired maximum measuring current) together with the voltage over the shunt at 10A and the maximum allowable gain to stay in the input range. It also shows the maximum current for that particular shunt to stay in the input range at that gain (Max current) and the power dissipation in the shunt at that current level (Power @ max current).

	100mΩ	50mΩ	5,0mΩ	3,0mΩ	2,0mΩ	1,0mΩ
Power @ 5A	2,50W	1,25W	125mW	75,0mW	50,0mW	25,0mW
Power @ 10A	10,0W	5,00W	500mW	300mW	200mW	100mW
Shunt voltage @ 10A	1000mV	500mV	50,0mV	30,0mV	20,0mV	10mV
Highest gain @ 10A	1	1	8	16	24	24
Max current	6,00A	12,0A	15,0A	12,5A	12,5A	25,0A
Power @ max current	3,60W	7,20W	1,13W	469mW	313mW	625mW

Figure 17: Shunt values and there properties

Figure 17 shows that a shunt value of 3,0mΩ is a nice tradeoff between range, power dissipation and value for the prototype. But a shunt resistor of 2,0mΩ would be preferred in a final design because it has the same current range but a lower power dissipation. And on a specific designed PCB the parasitic resistance can be better controlled. And even 1,0mΩ should not be a problem if a bigger range is desired. But increasing the measuring range will reduce the resolution because the overall dynamic range will stay 1:5000.

5 Implementation

This chapter describes the build of the hardware and the implementation of the software.

5.1 Hardware

The hardware used consists of two parts, the actual energy meter or called a Sensor Node, and the Serial Gateway. The Serial Gateways links the wireless communication to a controller application.

5.1.1 MySensors Serial Gateway

The hardware is implemented on a prototype board. A simple MySensors Serial Gateway is build according to the instructions from the MySensors website and shown in the schematic of Figure 18. It consists of an Arduino Nano breakout board, an nRF24L01+ breakout board and LEDs to indicate the status. It can just be seen as a bridge between the wireless and the controller. The main advantage of an Arduino Nano over an Arduino Pro Mini (as used in the Sensor Node) is the presence of a serial to USB chip which gives the Arduino Nano an USB interface which can be directly used by the controller application. The build prototype can be seen in Figure 20.
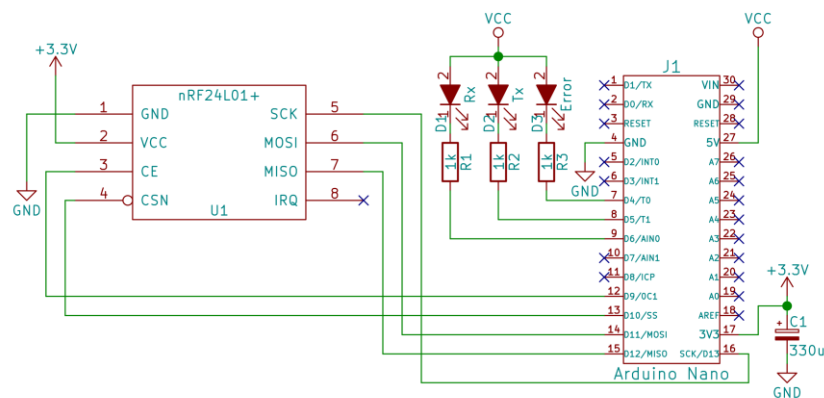

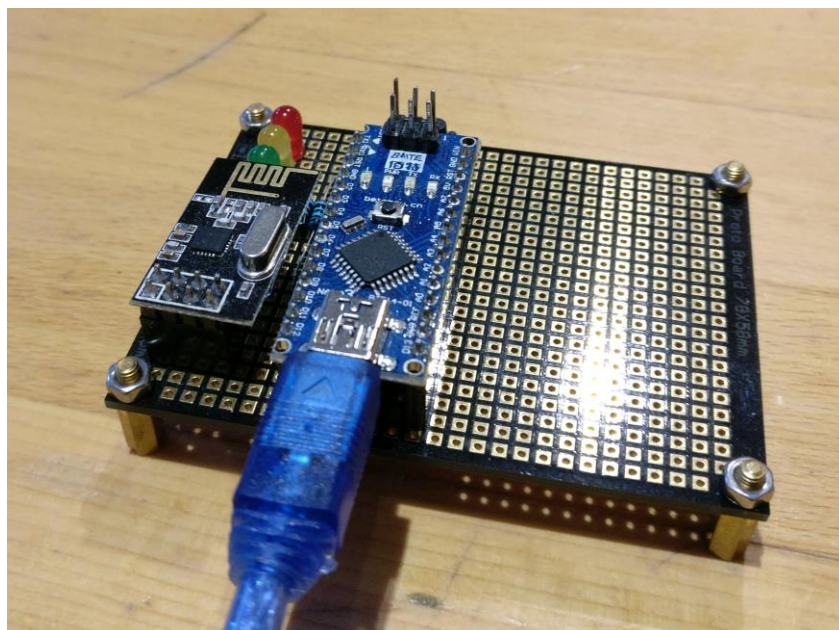*Figure 18: Schematic MySensors Serial Gateway*

Figure 19: MySensors Serial Gateway prototype

5.1.2 Energy measuring node

The energy measuring node is also implemented on a prototype board. This is done according to Figure 16 and consists of an Arduino Pro Mini breakout, an nRF24L01+ breakout, the ATM90E26 ASIC on a breakout board, the capacitive dropper power supply and the external components for the ATM90E26. This board is more complex than the Serial Gateway and took some time to assemble. The complete prototype can be seen in Figure 20. To make it easy to work with, all breakout boards are fitted in headers so they can easily be removed or swapped. All status LEDs from the schematic are left off. They are not needed for the device to work, consume energy and those statuses can also be read over the SPI connection to the ATM90E26.
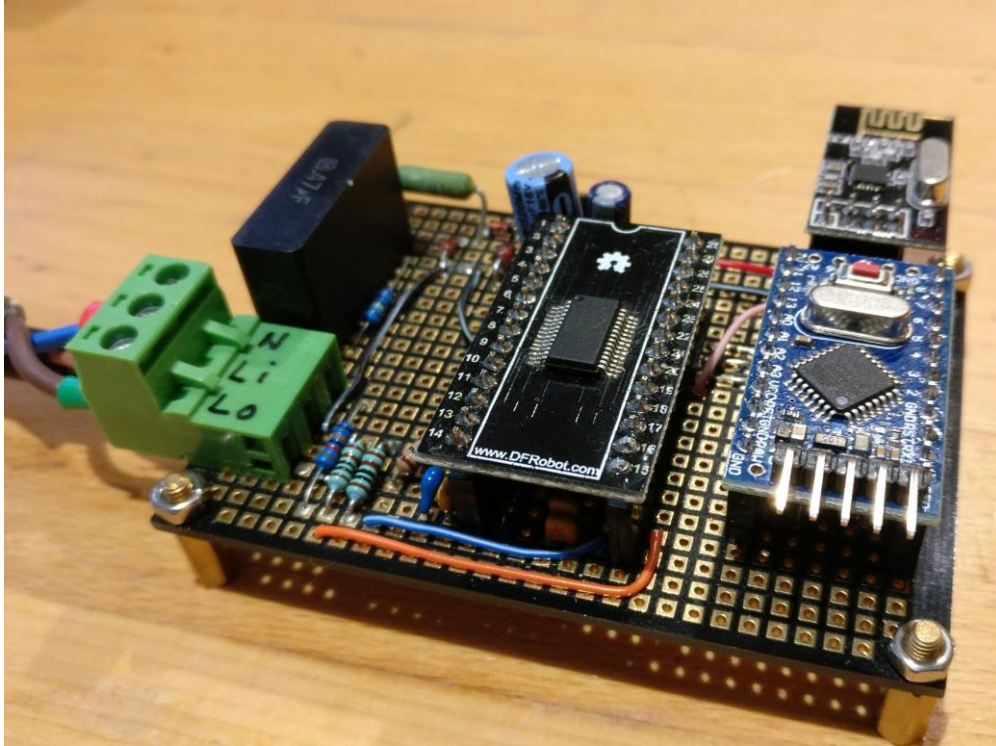
Figure 20: Energy measuring node prototype

Because the whole circuit is referenced to the grid voltage we cannot directly connect a computer (via a USB to serial converter). This is not a problem when we communicate wireless with it because that will already give the isolation. But in order to make an isolated wired connection an isolator is needed for the TTL serial signal. This was designed based on optocouplers which is shown in Figure 21. This was also build on prototype board.
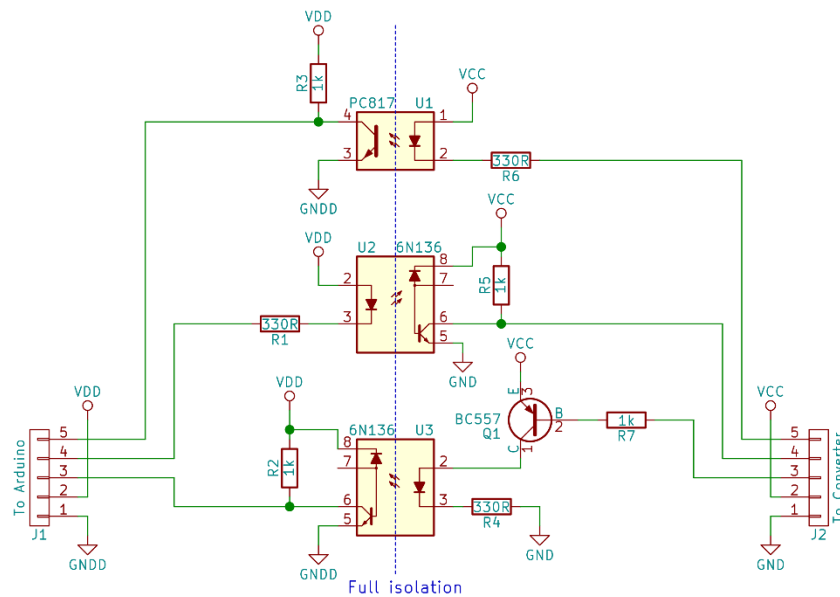
Figure 21: Schematic of TTL serial isolator

Unfortunately, because an optocouplers is based around a LED, the isolator needed more current then the power supply of the node could deliver. So in order to isolate the node from the grid an isolation transformer was used in front of the node. This gives the node the same mains voltage but fully isolated from the grid. This makes it safe to touch and connect to a computer. Downside to this is the limited current capacity of the isolation transformer because this will have to need to power any load connected to the measuring node. But once programmed and calibrated the connection to the PC is no longer needed thus the isolation transformer can be removed.

5.2 Software

This paragraph describes the software used and the software developed to make the energy meter and to use it.

5.2.1 MySensors Serial Gateway

The software for the Serial Gateway is just as provided by MySensors only configured for the nRF24L01+ transceiver. It will work directly with controllers like OpenHAB. The sketch is called GatewaySerial and is part of the MySensors-library.

5.2.2 Energy measuring node

To make the communication with the ATM90E26 easy and to keep it universal an object oriented library is written called ATM90E26-library. It uses the Arduino SPI library for the communication and provides an easy way to communicated with the ASIC. Because it is made object oriented it is also easy to read multiple ASIC's connected to a single microcontroller (with shared SPI-bus) by simply creating multiple objects.

The library is fully documented and HTML reference pages are generated with Doxygen.

In order to get measurements it is necessary to set up the ATM90E26. Mainly this is setting the correct current gain for the current measurement, doing a self-check and starting the measurement.

But in order to get sensible values it is also necessary to calibrate the ATM90E26. Not only to accommodate for differences in component values (of the voltage divider and current shunt) but mainly to set it up for those component values in general. This because the ATM90E26 can be used over a wide range of voltages and currents and has no setting for that. It is all calculated from the

calibration values. If in future meters, no explicit calibration is desired, the values used in this prototype can be used. This meter will then lack precise calibration but the values will act as setting for the used component values.

The basic calibration is for the voltage and the current. This is done by comparing the readout value of the ATM90E26 with the value of that parameter (voltage or current) externally measured. From that the desired (software) gain is calculated (for both voltage and current individual) which forms the calibration. The formula for that is

$$\alpha_{new} = int(\frac{\alpha_{old} \cdot x_{measured}}{x_{read}}) \tag{29}$$

With α_{new} the calculated gain to use as calibration value, α_{old} the power on value of the gain register, $x_{measured}$ the external measured value of the parameter and x_{read} the current readout value of that parameter. $int()$ means transforming the value into an integer value.

For the current calibration it is necessary to have a fully resistive load (0° phase shift) connected to the Energy Metering node. This is done with a +-42W halogen bulb. The current and voltage of are measured with a Hameg HM8115-2 Power Meter and checked with a Uni-T UT136B multimeter.

After the calibration values were determined the ATM90E26-library and the MySensors-library are linked in a program. The basic structure of this program is very simple and is shown in Figure 22. The MySensors-library takes care of all the communication with the controller application and the ATM90E26-library takes care of all the communication with the ASIC.

```c
//Include the libraries
#include <ATM90E26.h>
#include <MySensors.h>

void setup(){
  /*Set up the MySensors-library*/

  /*Set up the ATM90E26*/
}

void loop(){
  if(/*One second passed*/){
    readings = /*Take readings from ATM90E26*/

    /*Send readings to gateway*/
  }
}
```

Figure 22: Pseudo-code of the Energy Metering node

It will simply send all the reading to the controller every second which meets the design requirements. This is not a problem for the MySensors-framework. But it would be pretty easy to extend it to only send on significant changes because the libraries take care of the reading of the data and sending the readings.

5.2.3   Controller application

As a controller application OpenHAB is used because it is already used extensively in the energy group. For testing OpenHAB was used under Windows but it is also easy to install on an embedded platform like a Raspberry Pi. For that, OpenHAB provides installation via openHABian as an image file. Because OpenHAB is already used extensively in the energy group and the controller application is out of the scope of this thesis, only the basics of the Energy Metering node in OpenHAB is covered.

To link the MySensors-framework to OpenHAB the MySensors binding is used. The only configuration that needs to be done is to select the right COM-port of the MySensors Serial Gateway. After that the MySensors-framework is linked to OpenHAB. Every Energy Metering node that is powered up in range of the Serial Gateway will automatically present itself to OpenHAB. In OpenHAB the Energy Metering node will identify itself as two "Things" as shown in Figure 23. The first is power- sensor which can provide the "Items" real power, reactive power, apparent power and the power factor. The second is a multimeter-sensor which can provide the "Items" voltage and the current. From there the "Items" can be used in OpenHAB in the same way as the PlugWise.
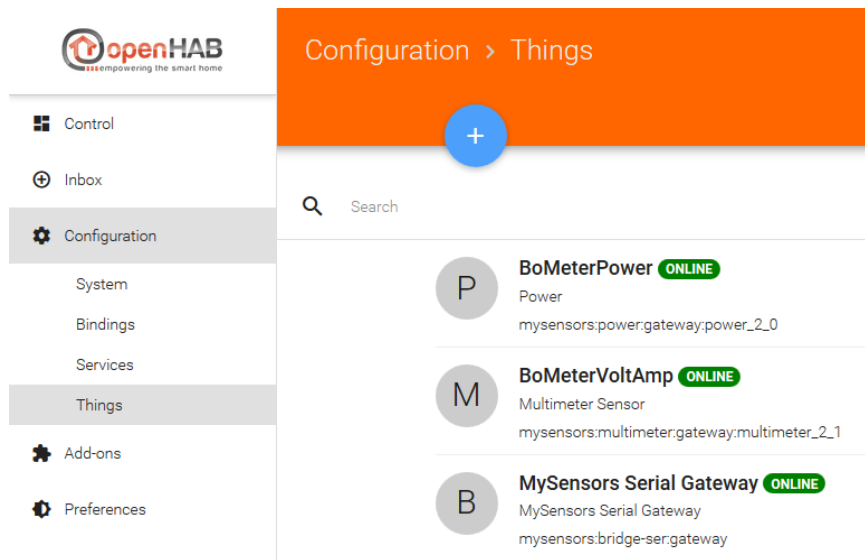

Figure 23: Energy Metering node as "Things" in OpenHAB

5.2.4   Summary

Below a summary of all the software used in this thesis in alphabetical order:

- Arduino IDE 1.8.3
- Doxygen 1.8.11
- KiCad 4.0.6
- MatLab R2016b (9.1.0.441655)
- MySensors library 2.1.1
- OpenHAB 2.0.0 Release build
- OpenHAB MySensors binding 2.1.0

6 Results

This chapter covers what tests were performed during this thesis in order to determine if requirements were met. It covers the Wireless range, the power supply issues and results from the ASIC.

6.1 Power supply

The power supply turned out to be the weakest part of the circuit. Although the current of the transceiver is lower than the average current of the power supply, it was not able to maintain the desired voltage and thus made the circuit reset. The main reason for this is that the supply only uses half of the sinewave. So half of the time all the power needs to come from the buffer capacitors. But the voltage drops too much during the negative halve and is not able to recover in the positive halve. An option would be to increase C1 (see Figure 16) in order to increase the average current. But with 470nF the current is already 31,6mA which results in an apparent power of

$$S = i_{RMS} \cdot U_{RMS} = 31,6mA \cdot 230V = 7,27VA \tag{30}$$

This is around the same apparent power of a Senseo coffee maker. Increasing C1 will result in an even higher apparent power and is considered to be too much for an energy meter which is used to reduce reactive power. For later results an external power supply is used.

6.2 Wireless

The MySensors framework in combination with the nRF24L01+ worked as expected. That framework takes care of being able to have multiple energy meters connected. But it is also possible to mix this with other sensor or actuator nodes.

Only the signal strength was tested. As stated in the chapter Design the design requirement was to be able to transmit 10m and through a wall. The reason for this is so that the Energy Metering node can be used in a normal house without communication problems.

There was no lab setup available to test the requirement. Instead more real life tests were conducted in the office and in a home.

In the office two tests were conducted. A clear line of sight test along the hallway and a between offices test. The clear line of sight test had no wall in between and transmission can through without error over a distance of more than 45m. The between offices test was able to transmit over more than 18m with two walls in between.

The in home test was conducted in a home with all reinforced concrete walls. It was able to transmit without error over a distance of around 10m with two walls in between. After 15m with 3 concrete walls no transmission was received.

	Distance	Number of walls
Office, line of sight	45m	None
Office, between offices	20m	2 office
Home, medium distance	10m	2 reinforced concrete
Home, long distance	15m	3 reinforced concrete

Figure 24: Rang tests Nrf24L01+ with MySensors-framework

The key of the tests was not to look for the longest possible range but arbitrary scenarios were picked. This because it was not an ideal lab situation so not every scenario was possible and all had a lot of variables. It is just to verify the nRF24L01+ was able to transmit in a home.

6.3   Power measurement

To verify the Power Measuring node a Hameg 8115-2 Power Meter is used. As a load, common household items where used and compared against the Power measuring node. This because there was no programmable load available which could set real and reactive power. And this would also just allow to check with sinusoidal loads and not more complex loads like for example SMPS. These can have a lot of harmonics due to their switching frequency and the use of non-linear components like diodes. And after all, the designed operation mode is to measure devices in a (smart) home in order for the controller to react. All measurements are done with the same calibration.

The first test is done with an assortment of different lightbulb. Mainly LED lightbulbs where used because they can should the biggest diversity and are getting coming in (smart) homes. Figure 25 shows the different bulbs. From left to right (**a**) 42W halogen, (**b**) 5W compact fluorescent, (**c**) 5W dimmable LED, (**d**) 6W dimmable LED, (**e**) 9W dimmable LED, (**f**) 5W dimmable LED and (**g**) 2W capacitive dropper LED. These are all the figures as advertised so might not always match the measurement from the power meter or Power Measuring node because it is not uncommon for the power rating to be exaggerated.

Figure 25: Different lightbulbs used to test the Energy Measuring node

In the table of Figure 26 the test results of the lightbulbs are shown. With P for real power, Q for reactive power, S for apparent power, PF for power factor, I for current and U for voltage. These tests are done with the isolation transformer in line which explains the dropping voltage with higher loads. As expected, not all bulbs meet there advertised rating. The Energy Measuring node follows the power meter pretty close but it has as an error. The power meter is not able to display the sign of the reactive power but the Energy Measuring node is. Because the LED and compact fluorescent bulb are both capacitive loads which lead to a negative reactive power that explains the sign. Voltage and current are just shown because they are used to initial calibrate the ATM90E26. Because these parameters are not the main interest of the energy meter these values are nor used to determine errors.

	Power meter						Energy node					
	P	Q	S	PF	I	U	P	Q	S	PF	I	U
a	31,1W	2,6VAR	31,1VA	1,00	172mA	180V	31,1W	-1,3VAR	31,1VA	1,00	172mA	180V
b	5,9W	6,1VAR	8,4VA	0,69	36mA	235V	5,8W	-2,7VAR	8,2VA	0,70	35mA	236V
c	4,4W	3,7VAR	5,7VA	0,76	24mA	237V	4,2W	-1,4VAR	4,2VA	1,00	20mA	238V
d	5,7W	3,2VAR	6,5VA	0,87	28mA	236V	5,7W	-1,7VAR	6,3VA	0,89	27mA	236V
e	4,5W	3,8VAR	5,8VA	0,76	24mA	237V	4,3W	-1,4VAR	4,2VA	1,00	26mA	238V
f	4,6W	2,8VAR	5,4VA	0,85	23mA	237V	4,4W	-1,6VAR	4,4VA	1,00	15mA	237V
g	1,7W	5,1VAR	5,3VA	0,33	22mA	243V	1,6W	-5,2VAR	6,3VA	0,25	16mA	243V

Figure 26: Measurements light bulb

In the table of Figure 27 the relative error is calculated for each bulb and an average for this test. The biggest error is in the measurement of the reactive power. The ATM90E26 seems to make a big error while calculating that. This might be due to no phase angle calibration and no offset compensation.

	Error P	Error Q	Error S	Error PF
a	0,00%	50,00%	0,00%	0,00%
b	1,69%	55,74%	2,38%	1,45%
c	4,55%	62,16%	26,32%	31,58%
d	0,00%	46,88%	3,08%	2,30%
e	4,44%	63,16%	27,59%	31,58%
f	4,35%	42,86%	18,52%	17,65%
g	5,88%	1,96%	18,87%	24,24%
Average	2,99%	46,11%	13,82%	15,54%

Figure 27: Relative errors of the tested lightbulbs

A second test is conducted with other household devices including higher power devices like an electric kettle and a vacuum cleaner. Because these devices cannot be powered by the isolation transformer these tests are conducted without it. To power the Energy Measuring node because of the problem mentioned in paragraph 6.1 an isolated power supply is used.

	Type of device
a	HTC USB charger
b	OnePlus Dash USB fast charger
c	Cheap brandless USB charger
d	LED lighting with transformer
e	Philips TV
f	Miele 1800W vacuum cleaner
g	2020W kettle

Figure 28: Household test devices

In Figure 29 the results of this test are shown and in Figure 30 shows the relative error of those measurements. Again the sign of the reactive power is ignored because the Power meter is not capable of displaying that. But it is in the line of expectation that d and f show a positive reactive power and the rest negative. The transformer in the LED lighting and the motor in the vacuum cleaner are both inductive loads were the rest are capacitive loads.

	Power meter						Energy node					
	P	Q	S	PF	I	U	P	Q	S	PF	I	U
a	5,9W	11,2VAR	12,6VA	0,46	54mA	231V	5,9W	-1,9VAR	12,5VA	0,46	53mA	231V
b	5,7W	10,9VAR	12,6VA	0,45	55mA	231V	5,8W	-2,0VAR	12,3VA	0,46	54mA	231V
c	20,4W	35,0VAR	40,0VA	0,51	177mA	231V	20,4W	-2,7VAR	40,3VA	0,51	177mA	231V
d	7,2W	10,8VAR	13,0VA	0,55	56mA	233V	7,3W	9,4VAR	13,0VA	0,56	56mA	233V
e	186W	36VAR	194VA	0,98	839mA	230V	187W	-22VAR	195VA	0,84	1mA	231V
f	1,43kW	0,42kVAR	1,48kVA	0,96	6,8A	219V	1,43kW	0,31kVAR	1,51kVA	0,96	6,8A	220V
g	1,75kW	80VAR	1,75kVA	1,00	8,0A	217V	1,76kW	-30VAR	1,76kVA	0,99	8,0A	219V

Figure 29: Measurements of the household devices

	Error P	Error Q	Error S	Error PF
a	0,00%	83,04%	0,79%	0,00%
b	1,75%	81,65%	2,38%	2,22%
c	0,00%	92,29%	0,75%	0,00%
d	1,39%	12,96%	0,00%	1,82%
e	0,54%	38,89%	0,52%	14,29%
f	0,00%	26,19%	2,03%	0,00%
g	0,57%	62,50%	0,57%	1,00%
Average	**0,61%**	**56,79%**	**1,01%**	**2,76%**

Figure 30: Relative error of household devices measurement

Again the real power and apparent power match closely within 2,5%. But the reactive power is off. But this is not completely due to the error in the Energy Measuring node but also due to error in the Power meter. As stated in 3.2 the reactive power can be calculated from the real power and the apparent power with (12). For the kettle (device **g**) that would result in a reactive power of $0,0kVAR$ for both the Power meter and the Energy Measuring node but they both show a small part reactive power resulting in a large relative error.

Also for the TV (device **e**) there is some error in the Power meter result. Because (12) would result in respectively

$$\sqrt{194^2 - 186^2} = 55VAR \tag{31}$$

$$\sqrt{195^2 - 187^2} = 55VAR \tag{32}$$

But this could be due the more fluctuating power consumption of the TV in comparison to the other devices. This makes it harder for the meters to give a time average of the reactive power.

But it is believed that the accuracy of the reactive power can be improved by better calibration of the phase angle and the offsets in the power. Due to time this is left for future work if the energy group is interested in using this energy meter in other tests or real applications.

6.4 Controller application

The MySensors-framework can be linked to many different controller software packages. Because OpenHAB is used a lot that is what was used to showcase the Energy Measuring node. Because the

data from the MySensors Serial Gateway is just a stream of serial data it is even possible to write custom software to read the sensor data.

During testing it was discovered that OpenHAB was not displaying the digits after the decimal point correctly in the interface. But the OpenHAB log-file shows it does parse the values including what is after the decimal point. Because the deployment of the Energy Measuring node in OpenHAB is actually outside of the scope of this thesis this is not further investigated.

7 Conclusion

In this thesis the task was set to design and implement (into a prototype) a wireless energy meter. To reach this goal, research, simulations, calculations and experiments were conducted. In multiple design iterations a final design is made. The main objective, as stated in the Introduction, was:

"Design a wireless energy meter which can measure real power, reactive power and voltage."

During this thesis multiple problems were solved and design choices made to come to a working prototype. So in general it can be said that the main objective is met. But the further validate this thesis three questions have been formed based on the problem introduced in the Introduction and the design choices in chapter 4. These questions are:

1. *"Are all design requirements set during this thesis met in the final prototype?"*

The initial design requirements mentioned in chapter 1 are met. The extended requirements from chapter 4.1 are not all met. As discussed in chapter 6.3 the proposed capacitive dropper power supply turned out to be not suitable. This was a bad design requirement to start with because of the bursts of power needed and the large reactive power inherent for a capacitive dropper.

The rest of the design requirements is met although nothing can be said about the sample frequency. Sampling is all done by the ASIC so this requirement does not apply anymore. The 1W resolution is met because of the more than sufficient dynamic range of the ASIC. But the meter still has a significant measurement error but this can probably be fixed with better calibration.

2. *"Is the designed energy meter a useful replacement for the Plugwise?"*

As a proof-of-concept the Energy Measurement node is a potential replacement of the Plugwise. The wireless aspect works similar to the Plugwise due to the simplicity and the support for multiple meters. Also the link with controller applications is similar.

However there is room for improvement considering the accuracy of the meter. Especially the measurement of the reactive power. But the accuracy of the Plugwise is similar and only specified at the higher range (5% above 1035W (Plugwise, 2011)) and at unknown power factor. Also it is missing the feature to measure reactive power altogether.

This makes the Energy measuring node already a useful replacement for the Plugwise. Especially considering the possible improvement of the accuracy which makes it more accurate and better specified than the Plugwise.

3. *"Was it a wise choice to use an ASIC instead of doing all the measurements and computation with a microcontroller?"*

During the research into measuring power a lot of difficulties where outlined. It would definitely be possible to make an accurate power meter with a microcontroller. This would give the advantage of having full control over the data and the features. For example, to add harmonics analysis. But the use of an energy metering ASIC made the process of designing and debugging the meter faster then what was expected with the use of a microcontroller alone. That made it possible to design, build and analyze a wireless energy meter in the limited time of this thesis. Also the ASIC met all the features set out in the design requirement and was a cheap solution which gave no restrictions to the energy meter.

In hindsight, the choice to use an ASIC instead of doing all the calculations on a microcontroller was certainly a justified choice. Without the ASIC it would not have been possible to come to a tested design and prototype.

7.1   Recommendations

Although this thesis resulted in a working energy meter which met the design requirements there are a couple of improvements which can be made. Mainly because this thesis only made a proof-of-concept prototype were as multiple finished units would be preferred if the energy meter is used in other research or in implemented energy management. But due to time restrictions these improvements are not implemented. These improvements are discussed in this paragraph.

7.1.1   Power supply

As stated in paragraph Power supply in chapter Results the power supply for the Energy metering node was not sufficient. As state there increasing the capacitor value may solve the issue but is undesirable. In hindsight the choice for a transformer-less capacitive dropper power supply did not work out. It would be recommended to use a small SMPS in the design. Although more expensive it is barely bigger then the capacitive dropper but will be able to supply the whole circuit at a much better power factor.

7.1.2   Calibration process

For the prototype the calibration values are determined by hand and hard-coded into the software. For future use it would be convenient to implement a calibration process in software. External hardware will be needed to verify the parameters but it would be possible to let the Energy Metering node calculate the calibration parameters in software and store them in EEPROM (non-volatile memory). That way the calibration process is easy and no change of the code is needed.

7.1.3   Full calibration

At the moment only the voltage and current gain are calibrated. But the ATM90E26 also supports phase angle calibration and offset compensation. For future use it would be recommended to implement those calibrations in the calibration process as well.

7.1.4   Fabricated PCB

The assembly of the prototype took quite some time. So if multiple Energy Metering nodes are desired it would be easier to design a PCB for it which can be fabricated. This can easily be done from the schematics already made. Attention must be paid to the use of proper current capacity traces, clearance between line and neutral, and the voltage rating of the resistors in the voltage divider. Optionally the ATmega328p and the nRF24L01+ can be implemented directly on that PCB instead of using breakout modules. For operation this makes no difference but will create a fully self-contained device.

8 Bibliography

Agentschap Telecom. (2014). *Vergunningsvrije radiotoepassingen.* Groningen: Agentschap Telecom. Opgehaald van https://www.agentschaptelecom.nl/sites/default/files/brochure-vergunningsvrije-radiotoepassingen.pdf

Atmel. (2014). *Application Note: Single-Phase Energy Meterng IC M90E26.* San Jose, CA: Atmel Corporation. Retrieved from http://www.atmel.com/Images/Atmel-46102-SE-M90E26-ApplicationNote.pdf

Atmel Corporation. (2009). *AVR200: Multiply and Divide Routines.* San Jose.

Eijkel, J., Heijden, F. v., & Salm, C. (2008). *Inleiding Elektronica en Elektrotechniek.* University of Twente, faculty EWI.

Espressif. (2017). *ESP8266EX datasheet v5.4.* Espressif. Retrieved from https://espressif.com/sites/default/files/documentation/0a-esp8266ex_datasheet_en.pdf

Jäckle, T. (2016). *Application Note 105: Power measurement and its theoretical background.* ZES Zimmer. Retrieved from http://www.zes.com/en/content/download/655/6442/file/Applikat_105_Power_Rev_2.0_e.pdf

Key, T., & Lai, J.-S. (1997). *IEEE and International Harmonic Standards Impact on Power Electronic Equipment Design.* IEEE.

Kozub, R., & Radhostem, R. p. (2012). *Application Note: Time Domain Based Active Power Calculation.* Freescale Semiconductor.

Minister van Economische Zaken. (2015). *Regeling gebruik van frequentieruimte zonder vergunning en zonder meldingsplicht 2015.* Overheid.nl. Retrieved from http://wetten.overheid.nl/BWBR0036378/2016-12-28

Nordic Semiconductor. (2008). *nRF24L0+ - Single Chips 2.4GHz Transceiver - Preliminary Product Specification v1.0.* Nordic Semiconductor. Retrieved from http://www.nordicsemi.com/eng/nordic/download_resource/8765/2/7021907/2726

Plugwise. (2011). *Getting started with Plugwise; Installation manual.* Plugwise. Retrieved from https://www.plugwise.com/media/wysiwyg/handleidingen_en/home-start-and-basic-manual.pdf

Williams, E. (2017, April 4). *The Shocking Truth About Transformerless Power Supplies*. Retrieved from Hackaday: http://hackaday.com/2017/04/04/the-shocking-truth-about-transformerless-power-supplies