

Fluctuation enhanced gas detector for wireless sensor networks

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Abstract—Fluctuation enhanced sensing (FES) is a promising method to improve information extraction from noisy sensor signal especially in the rapidly growing field of semiconductor and nanotechnology gas sensors. Some attempts were made to support the application of the principle in low power, battery powered devices and partial results are available. In this paper we present a complete mixed-signal system that contains a low power analog and digital signal processing part as well. Besides theoretical investigations we have carried out measurements and simulations to optimize the performance and achieve high accuracy. Including all components and processing the system needs only 2mA of active supply current working from a 3V battery and a coin cell battery can provide 250 days of operation at sample rate of 6 samples per hour.

Keywords—fluctuation enhanced sensing; noise; wireless sensors

I. INTRODUCTION

The rapidly developing field of wireless sensor networks need low power, battery operated devices. The efficient information extraction from the raw sensor signal often need extensive analog and digital signal processing, therefore it is quite challenging to keep the power consumption low. This is particularly true for noisy signals when the information source is the noise itself, so estimation of time domain statistical parameters and power spectral density (PSD) is frequently used.

The principle of fluctuation enhanced sensing (FES) had been introduced a few years ago as a candidate to improve the detection of various gases or gas mixtures with even a single sensor [1]. In most cases the resistance of Taguchi and nanotechnology gas sensor is measured to determine the gas concentration; however the selectivity can be rather small, there is no information about possible mixture of several gases. In this latter case typically many sensors are used in parallel that are sensitive to different gases. Fluctuation enhanced sensing is typically utilized to overcome these limitations by measuring the time dependent fluctuations of the resistance that is caused by the random adsorption and desorption process. FES principle is successfully applied to detect different gases with a single sensor, to detect gas mixtures, harmful gases, bacterial odors, scents [2-10]. The analysis of resistance fluctuations needs PSD estimation by

taking several samples and computation-intensive pattern recognition algorithms are often applied to extract the required information.

II. LOW POWER FLUCTUATION ENHANCED SENSING SYSTEM

A typical experimental FES system consists of a low noise power supply, low noise preamplifier, spectrum analyzer or a data acquisition unit connected to a personal computer that runs the FES algorithm. Low power solutions require much smaller systems and simplified algorithms that can be executed even on a small microcontroller. Both hardware realizations and analysis methods are reported, however a complete, stand-alone intelligent sensor nodes including all analog and digital processing are still missing in the field of FES [11-17].

Our complete FES sensor node is based on the principle of approximating the power spectral density using the power measured in a few adjacent frequency bands [11] like in traditional spectrum analyzers. This is typically done by using band-pass filters to divide the signal into different frequency band components [13]. Band pass filters can also be replaced by low-pass filters because they are more simple to realize and the same result can be obtained [13, 14]. The outputs of these filters are used to estimate the power in the different frequency bands by either analogue signal processing components or by analogue-to-digital conversion and digital processing.

We have applied this principle to realize a single supply battery powered complete sensor node that provides the required digital signal processing also. The block diagram of the system is show on Fig. 1.

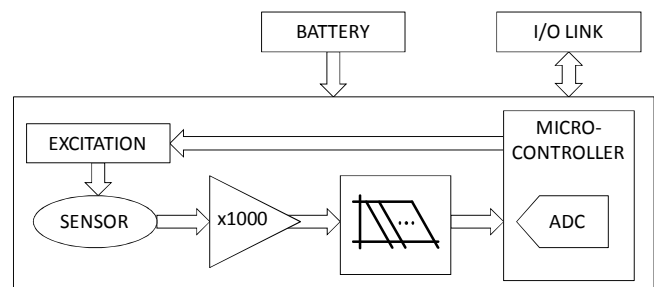


Fig. 1. Block diagram of the low power FES device.

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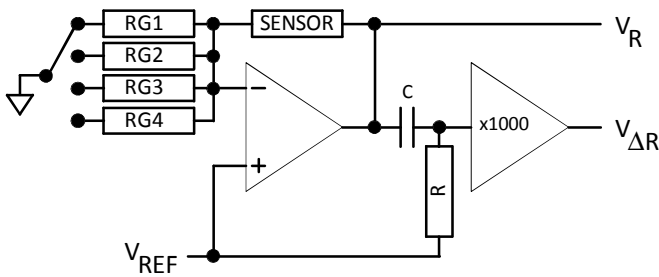


Fig. 2. Sensor excitation and preamplifier circuit.

Fig. 2 shows that the resistive sensor is placed in the feedback loop of an operational amplifier, OPA378. The amplifier operates at supply current as low as $125\mu\text{A}$, the supply voltage range is from 2.2V to 5.5V . The voltage noise is $20\text{nV}/\sqrt{\text{Hz}}$ and what is more important, the current noise is only $200\text{fA}/\sqrt{\text{Hz}}$, since the sensor impedance is typically well above $25\text{k}\Omega$. The current forced through the sensor is equal to V_{REF}/R_{G_i} ($i=1..4$) and can be varied by changing the state of the switch under software control. This allows wide range of sensor resistance values to be used. Indeed, semiconductor and nanotechnology based sensors can have resistance from a few $\text{k}\Omega$ to several $\text{M}\Omega$ s.

In this first stage the sensor resistance is converted to a voltage V_R that contains the small fluctuations as well. The simple 1Hz RC high-pass filter removes the slowly changing components and a gain of 1000 amplifier outputs a signal $V_{\Delta R}$ that is large enough for further analysis of the resistance fluctuations in the range from 1Hz to 10kHz , over four decades of frequency.

At this point the low-pass filter bank is applied to break the signal into different but overlapping frequency bands (see Fig. 3). We have applied eight passive low-pass filters with nominal logarithmic corner frequencies of 10Hz , 27Hz , 72Hz , 193Hz , 518Hz , 1389Hz , 3728Hz , 10000Hz . The eight filtered signals are connected to the inputs of the C8051F410 microcontroller. This mixed-signal microcontroller features an analog multiplexer and a precision 12-bit analogue-to-digital converter therefore no additional external components are needed. The microcontroller also provides the V_{REF} reference voltage, drives the excitation selection quad switch (ADG704). The microcontroller is powered from the same battery that powers the analogue circuitry and runs at 1.5MHz core frequency. The microcontroller digitizes the eight signals

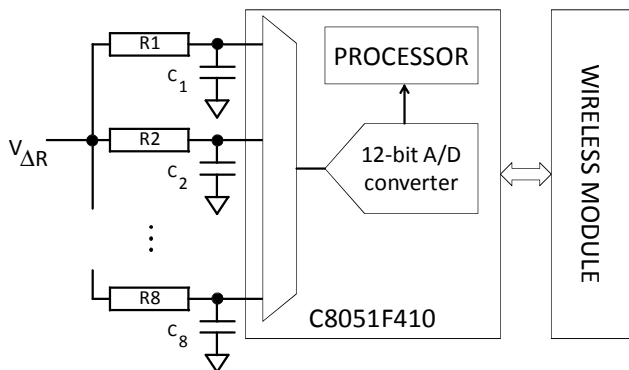


Fig. 3. Low pass filter bank and digital signal processing components.

and calculates the variance over a given amount of time. After this, it can transmit the variance data over the I/O link, typically using a wireless communication method, but wired interfaces are also available including I2C, SPI and UART (RS232, RS485).

III. ACCURACY AND SENSITIVITY OF THE SYSTEM

We have investigated the achievable accuracy and sensitivity of the above described method and realization. Both statistical and deterministic errors were considered.

First of all, the preamplifier stage generates noise approximately equivalent to the thermal noise of a $25\text{k}\Omega$ resistor at room temperature, so this is much less than the typical gas sensor noise. Since it is uncorrelated and can be measured, it can even be subtracted from the variance data. Although the chosen operational amplifier performs well in most sensor applications power consumption or noise level can be optimized by selecting an alternative amplifier. Note that lower power is typically associated with higher noise and lower bandwidth.

The accuracy of the filter bank corner frequencies affects the precision of power estimation also. Resistors can have precision of 1% or below while capacitor values are a bit less accurate therefore they determine the overall accuracy. The error of variance of the signal at the output of the eight filters depends on the corner frequency tolerance. Fig. 4 shows that this sensitivity is quite low and note also that this deterministic error can be rather easily compensated.

The FES principle is based on detecting small changes in the shape of the power spectral density therefore it is worth to determine how small changes can be observed using the variance data obtained at the outputs of the low pass filters. We have carried out numerical calculations and simulations. We have assumed $1/f^\alpha$ ($\alpha=1$) noise typical in gas sensor applications and applied small changes of the exponent α in the range from 0.8 to 1.2 . Fig. 5 shows the eight variance values as a function of α . The variance evaluation time was 10 seconds. One can easily see that 0.05 change in α can be easily resolved and of course this can be further enhanced by taking

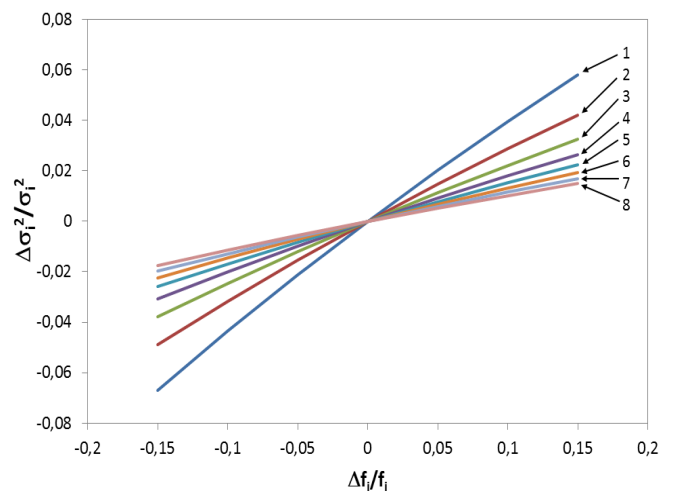


Fig. 4. Sensitivity of the noise variance at the filter output as a function of the filter corner frequency for the eight filters.

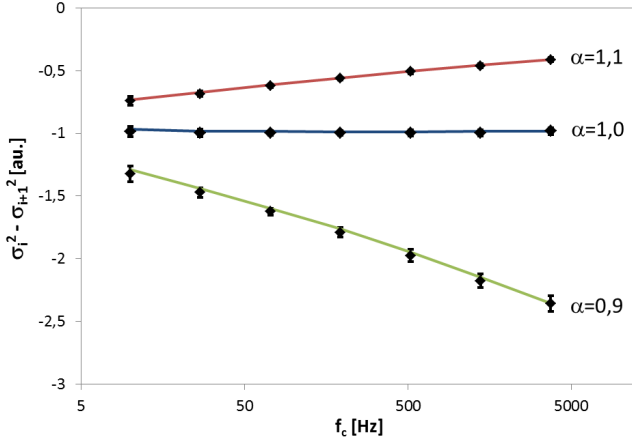


Fig. 5. Difference of the noise variance of neighboring filters as a function of the filter corner frequency for three different values of alpha.

longer measurements.

The last part of the signal processing chain is the analogue-to-digital conversion and calculation of the variance in the digital domain. In order to avoid overdriving the analogue-to-digital converter input the noise signal must fit in the range of the voltage reference. Assuming Gaussian distribution the signal only uses a fraction of this range in most of the time, therefore the effective resolution can be significantly reduced. The quantization noise and the resistance fluctuations are uncorrelated therefore the variance of the quantization noise is an additional term in the total variance of the fluctuations. The theoretical standard deviation of the quantization noise σ_q is equal to $V_{LSB}/\sqrt{12}$, where V_{LSB} is the voltage resolution. V_{LSB} is equal to $V_{REF}/2^b$, where V_{REF} is the measurement range and b is the number of bits of the analogue-to-digital converter. This means that even if the effective resolution is small, the noise contribution can be small as well. The measured variance of the total noise can be expressed as (1).

$$\sigma^2 = \sigma_n^2 + \sigma_q^2 = (c \cdot V_{REF})^2 + \left(\frac{V_{REF}}{2^b \sqrt{12}} \right)^2 \quad (1)$$

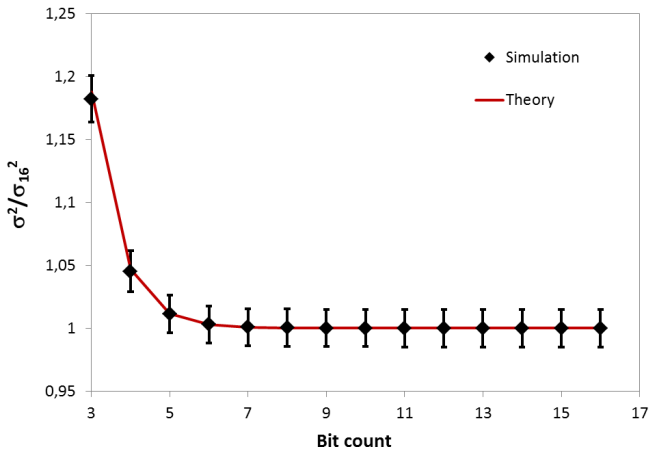


Fig. 6. Normalized variance as a function of the resolution of the analog-to-digital converter.

where σ_n represents the standard deviation of the noise of the sensor. The value of σ_n can be written as $c \cdot V_{REF}$, so c can be used to relate the noise amplitude to the full scale. The value of c must be small enough to avoid overdriving the input of the analogue-to-digital converter. The peak-to-peak noise amplitude is often approximated as $6 \cdot \sigma_n$, therefore c must be less than $1/6$. The ratio of the measured total noise and the noise coming from the sensor is given by (2).

$$\frac{\sigma^2}{\sigma_n^2} = 1 + \frac{V_{REF}^2}{c^2 V_{REF}^2} = 1 + \frac{1}{c^2 \cdot 4^b \cdot 12} \quad (2)$$

We have also carried out numerical simulations to estimate the error introduced by the finite resolution of the analogue-to-digital converter. In our simulations we have varied the resolution (number of bits) of the analogue-to-digital and assumed noise peak-to-peak value of half the input range, $6 \cdot \sigma_n = V_{REF}/2$, therefore the value of c is $1/12$. The simulation results show good agreement with the prediction, Fig. 6 shows that even 5 bits of resolution introduces an error less than the error caused by averaging over ten seconds assuming no quantization. Therefore the 12-bit analogue-to-digital converter of the microcontroller fits well to all of the practical FES applications even if the noise only uses a fraction of the input range of the analogue-to-digital converter.

IV. POWER EFFICIENCY

The power requirements are determined by the power consumption of the operational amplifiers, the microcontroller (the core, analogue-to-digital converter, port loading and internal oscillator) and dissipation on the resistive components of the analogue signal conditioning part including the filter bank. We have carried out simulations and measurements with large noise values, rather low (10kOhm) sensor resistance where the required power is the highest. The applied supply voltage was 3V.

The analogue part of the electronics needs about 600 μ A of total operating current, including the filter bank that requires only about 30 μ A. The filter bank's resistors are in the range of 4,7kOhms to 47kOhms, capacitors are chosen from 10nF to 330nF in order to reduce power consumption and to keep the impedance low enough to reduce external noise contamination in the same time (see Table I.).

TABLE I. PARAMETERS OF THE FILTER BANK

filter	1	2	3	4	5	6	7	8
R[k Ω]	47	18	22	18	30	11	13	4,7
C[nF]	330	330	100	47	10	10	3,3	3,3
f_c [Hz]	10,3	26,8	72,4	188	531	1448	3712	10267

The microcontroller runs at 1,5MHz that is enough to convert the eight input signals at 8kHz ($f_s=1$ kHz/channel) rate and perform the variance calculations. In order to evaluate the noise variance for each channel, the following algorithm was used:

1. Start an analogue to digital conversion on the signal of the k -th filter;
2. Read the data x_k after the end of conversion;
3. Set the multiplexer to the next filter's output;
4. Calculate $v_k = v_k + (x_k - x_{ok})^2$;
5. $k = (k+1) \bmod 8$;
6. if $k=0$ then $i=i+1$;
7. if $i = f_s \cdot T$ then
 - send v_k/i over the wireless port;
 - reset v_k to 0 for each channel ($k=0..7$);
 - reset i to 0.

Here k is a zero-based index of the selected filter, i counts the samples taken per channel during the averaging time T (10 seconds, for example), v_k is the variance of the k -th filter's output signal and f_s is the effective sample rate per channel (1kHz). The x_k sampled raw data are represented by 16-bit signed integer numbers, while the v_k variance needs to be a 32-bit integer number. This allows fast enough execution therefore the processor clock can be kept as low as 1,5MHz at this sample rate.

We have measured the power consumption of the microcontroller performing the above described algorithm (see Table II.). The supply current if the microcontroller is less than 1200 μ A that includes the analogue part of the chip as well (sample-and-hold, analogue-to-digital converter, voltage reference) therefore the total supply current of the system is below 2mA. At 3V supply voltage this means less than 6mW of power consumption.

TABLE II. POWER DISSIPATION OF THE DEVICE

Circuit	Power dissipation [mW]
Sensor excitation circuit (10kOhm sensor resistance)	0,66
Low noise preamplifier	1,11
Passive low-pass filters	0,09
Microcontroller (oscillator, analog-to-digital converter, voltage reference)	3,6
Total power dissipation	5,46

During inactive phase the whole system can be powered down, the microcontroller in the sleeping mode needs less than 3 μ A of operation current. If one assumes 10 seconds of active phase in every 10 minutes – in other words, gas evaluation is taken once in every ten minutes – the average supply current for 10 minutes is 33 μ A. This means that the system can run off a 200mAh CR2032 coin cell battery for 250 days.

Since the wireless or wired communication is needed for transmitting only a few bytes per tens of seconds, the power consumption of this communication part can be neglected.

CONCLUSION

A low power complete fluctuation enhanced gas sensor node has been shown that incorporates all analog and digital signal processing components. The system uses a filter bank introduced in [11] to estimate the power spectral density of the gas sensor resistance fluctuations and applies a low power microcontroller to perform the required digital signal processing and system control functions. The accuracy and selectivity of the system is determined and the power consumption is found to be less than 6mW allowing long operation time of hundreds of days running off a coin cell battery.

The possible applications include gas mixture detection, harmful gas detection, bacterial odor sensing in industrial and commercial applications. The system can serve as a basis of wireless gas sensor node in various wireless sensor networks. The flexibility provided by applying a precision microcontroller allows more advanced signal processing and integration into existing systems as well.

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