Improving the Precision of a Compact Subsampling Impedance Analyzer for Resonating Sensors

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Abstract

In recent years, compact and precise impedance measurement electronics have proved to be reasonable alternatives for dedicated measurement tasks like impedance measurement of resonant sensors. In this contribution, the design of a compact subsampling impedance analyzer is discussed. Furthermore the target application, a quartz crystal resonator (QCR) viscosity sensor for conductive liquids, as intended for the monitoring of a zeolite synthesis process aboard the International Space Station (ISS), is presented. In particular, accuracy aspects of the involved signal processing stage are examined.

impedance analyzer; subsampling; resonating sensors; quartz crystal resonator;

1. Introduction

Resonant sensors are used in a wide range of applications, e.g., as microbalances, chemical sensors in liquid and gaseous environments, and for physical property sensing of liquid and viscoelastic media. For measurements in liquids the Q-factor of such devices is much lower than that of resonators operating in vacuum or gaseous environments. Consequently, oscillator circuits are not first class evaluation circuits for these applications. Besides the common method of using precise laboratory instruments for the measurement of the impedance spectra of these sensors, various approaches were reported. For the interpretation of the measurement results not only the amplitude but also the phase angle of the sensor impedance is of interest, allowing a more accurate detection of the resonance behavior and the elimination of spurious impedance effects. The separation of sensor impedance into real and imaginary part is performed by (de)modulation. Thus analog implementations require nonlinear elements. To minimize negative effects due to these components, it is important, e.g., to keep offset currents, as induced by the parallel electrode-capacitance of the quartz crystal resonator (QCR) sensor, as small as possible or take steps for compensation. Alternative approaches perform (de)modulation in the digital domain increasing the requirements to

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resolution and sampling accuracy. In particular, aperture jitter noise is a key factor in systems with high sampling rates because its impact increases with the signal frequency.

2. Measurement setup

The interface electronics\(^4\) is particularly dedicated to monitor the viscosity in a zeolite synthesis process involving a harsh chemical environment of conductive liquid at a pH-value up to 13. Under these circumstances a fully immersed QCR — where both electrodes are in contact with the liquid — would suffer from spurious shunt admittance in parallel to the resonator\(^5\). While the conductivity of the liquid can be compensated up to a certain amount, it turned out that the electrochemically active solution in the zeolite synthesis distorts impedance measurements severely. To avoid this effect only one side of the sensor is contacted by the liquid in the corresponding realized sensor setup (Fig. 1). The implemented design\(^4\) (Fig. 2) is based upon a direct digital synthesizer (DDS) signal source and a direct sampling technique. The major design criteria were to minimize aperture jitter noise and to omit the utilization of a second analog to digital converter (ADC) for the phase measurement.

3. Signal Processing

3.1. Demodulation

As depicted in Fig. 2, a DDS-synthesizer generates two symmetric sinusoidal signals, an excitation signal

\[ s_e = a \cdot \sin(2\pi f_e t) \]  

(1)

with tunable amplitude \(a\) which, by means of a driver stage, is applied as driving voltage to the sensor, and a trigger signal.
\[ s_i = \sin(2\pi f_i t) \tag{2} \]

for clocking the ADC. As these signals are digitally generated from the same system clock, the frequencies and phase angles are in a well-defined relation to each other. In particular, in the proposed configuration, the current through the sensor, which is obtained in terms of a shunt voltage is (de)modulated by the ADC clock, yielding a signal

\[ s_s[k] = a \cdot \sin(2\pi f_s k T) \quad \text{with} \quad f_s = f_e \pm f_t , \quad T = \frac{1}{f_t} \tag{3} \]

which is a subsampled replication of the shunt voltage signal. A very fast and simple method of obtaining the real and imaginary part of the sensor impedance is to set \( f_t = 4 f_e \) because computational effort reduces to a offset correction. With respect to different noise sources, this approach only reduces uncorrelated noise but does not reduce correlated noise like harmonic distortions or interferences. Therefore it is advantageous to sample the signal at many different phase angles and to perform a second demodulation step in the digital domain. To obtain a scan of \( m \) samples per period of \( s_i \) (in the subsampled domain), the trigger frequency \( f_t \) can then be derived from a known excitation frequency \( f_e \) by

\[ f_t = f_e \cdot \frac{m}{m-1} . \tag{4} \]

Taking samples at many different phase angles of the shunt signal enables the treatment of deterministic noise by appropriate signal processing algorithms. Hence increasing the phase resolution by increasing \( m \) is advantageous over a simple averaging of results acquired at only four phase angles. The second demodulation step, which is required for the determination of the complex sensor impedance, is performed by the DSP using Goertzel’s highly efficient algorithm\(^6\) for computing a single discrete spectral component.

### 3.2. Impacts of frequency resolution

Due to the finite frequency resolution of the synthesizer \( \Delta f_{\text{max}} \), in general it is not possible to precisely set the trigger frequency to the value obtained from Eqn. 4. By introducing a known frequency deviation \( \Delta f \leq \Delta f_{\text{max}} \) the quantized trigger frequency can be found as

\[ f_t' = f_t - \Delta f = f_e \cdot \frac{m'}{m'-1} , \quad \text{with} \quad \Delta f = f_t \% \Delta f_{\text{max}} , \tag{5} \]

where \( m' \) is now a rational number with \( |m' - m| < 1 \) which can be obtained by substituting Eqn. 4 into Eqn. 5 yielding

\[ m' = \frac{f_e \cdot m - \Delta f (m-1)}{f_e - \Delta f (m-1)} . \tag{6} \]

As \( m' \) is not an integer (see Fig. 3), the result of the Goertzel-filter suffers from a leakage effect. In the spectral domain, this error can be described with a sinc-function\(^6\) showing that for \( |m' - m| \ll 1 \) the magnitude error can be neglected. The effect on the phase in contrast is significantly higher, as can be seen in Fig. 3. The deviation can be estimated by

\[ \Delta \varphi = \pi \frac{m' - m}{m'} . \tag{7} \]
For the proposed design using a frequency resolution of 32 bit at a system clock of 250 MHz yielding $\Delta f \leq 116$ mHz the resulting phase error is shown in Fig. 4. By substituting the equations 5 and 6 into 7 it can be seen that the error $\Delta \phi$ increases with $m$, hence it has to be taken into account for precise measurements. With respect to computational effort, compensating this error is clearly preferable to adapting the signal processing algorithms (e.g. by implementing a windowing-function).

Conclusion

The proposed method of evaluating the impedance of a resonator is advantageous with respect to deterministic noise on the measured signal. For very precise determination of the sensor impedance at selected frequencies one can increase the number of samples captured at many different phase angles of the signal. On one hand this decreases the uncertainty of the measurement result, but on the other hand it increases a deterministic error which has to be taken into account to obtain low standard deviation.

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