An Amplitude-to-Time Conversion Technique Suitable for Multichannel Data Acquisition and Bioimpedance Imaging

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Abstract—In this paper we exploit the high timing resolution offered by microprocessors to develop an amplitude measurement approach that is convenient for high channel count portable sinusoidal recording systems such as the bioimpedance measurements used in impedance imaging. This approach reduces the number of components required per channel, reducing cost, size and power consumption compared to the traditional approaches. The setup uses two high performance comparators to convert amplitude difference to a timing difference. This is captured by a high speed microprocessor. A straightforward algorithm removes DC and timing offsets. We suggest three modes of operation: fast: less than one period of the input, normal: exactly one input period and high precision: multiple input periods. The mean signal-to-noise ratio was 40, 81, and 112.4 dB in fast, normal, and high precision mode respectively for a range of resistive loads.

Index Terms—Amplitude-to-time conversion, analog-to-digital converter (ADC), capture unit, bioimpedance imaging, microprocessors, multichannel data acquisition.

I. INTRODUCTION

The electrical bioimpedance spectrum is an inherent characteristic of biological tissue related to the intracellular and extracellular volume and cell membranes. These change due to the physiological and pathological status of tissues so bioimpedance may be used as a powerful diagnostic tool. Bioimpedance is widely used in applications ranging from cell monitoring [1]–[7] and physiology [8]–[10] as it is sensitive to changes as small as the ion channels of the cell membrane to the large, temporary change in tissue structure during respiration or blood circulation.

In bioimpedance, a safe sinusoidal current at controllable frequencies is applied to a pair of electrodes and the resulting voltage changes are measured at additional electrodes [11]. Noise reduction may be greater with bioimpedance than straightforward biopotential measurement as we can demodulate at the known applied frequencies. In this paper we chose a 500 Hz frequency source to test the system as it is above the bandwidth of physiological signals (<300 Hz) and low enough not to be adversely affected by reactance (>1 MHz) [12]–[14].

Imaging systems based on multichannel bioimpedance, such as electrical impedance tomography (EIT) will benefit from high channel counts but this increases the cost, size and power consumption [15]–[20]. The purpose of this paper is to describe an amplitude-to-time conversion approach that can achieve efficient multichannel bioimpedance measurements.

Many EIT research groups have used analog or digital phase-sensitive demodulation techniques to detect the change of impedance distribution inside the body [21]–[23]. While these techniques are accurate, they are not suitable for high channel count systems due to the high cost and calculation overhead of the demodulation algorithm. They require several analog components or multichannel high resolution analog-to-digital converters (ADCs) which are sensitive to temperature variations, mismatch, crosstalk, power supply noise and reference voltage stability.

Recently there has been an interest in developing amplitude-to-time conversion for analog to digital conversion to particularly address the variation effects in the deep sub micron CMOS processes that have been primarily developed for high speed digital logic [24]–[26]. The advantage comes from exploiting the high time resolution of high speed logic or clocking circuits (typically a voltage controlled oscillator) to convert time difference signals that represent the original amplitude changes. A counter is used to convert the time differences to digital values [27]. This is a much more efficient process than flash ADC that requires $2^N - 1$ comparators, successive approximation (SAR) ADC that requires $2^N$ succession steps, and delta-sigma ADC that requires oversampling, filtering and negative feedback [28]–[33]. For bioimpedance and EIT imaging we suggest an off the shelf implementation of amplitude-to-time conversion that is different from the monolithic designs and conventional ADC as it only uses two comparators and the capture unit of the high speed microprocessor.

In this paper, we will describe the multichannel data acquisition method using a time difference for bioimpedance measure-
ments. It is applicable to any measurement using the amplitude response to a sinusoid stimulus such as audio measurements, amplitude modulation or power measurements. There are 3 different operating modes: fast acquisition mode, normal acquisition mode and high precision mode. There is a simple algorithm to convert the time difference to the amplitude of measured signal. When a signal includes DC offset, we can calibrate the measurement to remove the offset using a DC offset nulling method. We apply comparator calibration because the comparators are not ideal. We compared the performance of a range of discrete comparators and operating modes by their mean signal-to-noise ratio (SNR) on a serially connected resistor network.

II. METHODS

A. Amplitude-to-Time Converter (ATC)

Our method of amplitude-to-time conversion is based on the observation that sinusoids of different amplitudes can be observed to intercept a delayed reference signal at different phases or time delays, (1).

\[ A \sin(2\pi ft) = \sin(2\pi ft + \phi) \]  

We are interested in only measuring the amplitude as we plan to measure bioimpedance signals at low frequencies up to several kHz where phase information is minor. We also assume no distortion or non-linearity of the sinusoid which is reasonable for bioimpedance signals below a few volts when using standard Ag/AgCl based electrocardiogram (ECG) electrodes applied to the skin.

B. System Design

Fig. 1 shows a block diagram for the proposed system. We first apply a sinusoidal current to the subject and measure the resulting voltage which is modulated by bioimpedance changes due to e.g., respiration of the lungs increasing in size and filling with air. The voltage at each electrode is buffered by a dedicated buffer for each electrode then multiplexed to Comparator B in the 'sensing unit'.

A reference sinusoid is generated by delaying a copy of the applied sinusoidal signal with a fixed phase in the 'phase-delay unit'. Comparator B detects the intercepts or crossing points of the phase delayed reference and sensed voltage resulting in the square wave, Capture B. A second square wave, Capture A is generated by detecting the zero crossings of the phase delayed reference.

The time difference between edges of Capture A and Capture B is related to the amplitude of the sensed voltage as it represents the time taken for the phase delayed reference to pass from zero volts to the amplitude of the sensed voltage. We calculate the estimated amplitude value from time differences in the 'Amplitude Calculation Unit' using (2).

\[ A = \frac{\sin(2\pi ft + \phi)}{\sin(2\pi ft)} = \cos \phi + \sin \phi \cdot \cot(2\pi ft) \]  

C. DC Offset Calibration

DC offsets are a common issue when using skin based electrodes due to the half cell potential that exists between metal and ionic conductors. These might be removed by a filter but that would add an additional component with associated sensitivity to temperature variations and mismatch, particularly relating to the phase for low frequencies close to DC. Digital phase sensitive demodulation techniques reject DC offsets but they require high computation overheads.

However the amplitude-to-time conversion technique presented here can easily remove DC offsets because they are revealed in mismatch of the time difference between rising edges and falling edges in the Capture waveforms. The following (3)–(5) describe the DC subtraction algorithm. Equation (3) describes the time difference between rising edges, \( T_1 \). Here the sensed voltage has an amplitude \( A \) and DC offset \( V_{\text{DC}} \). We have added our deliberate phase offset \( \phi \) to the reference sinusoid which has a different DC offset \( V_{\text{DC}}' \). Equation (4) describes the time difference between falling edges. By taking the difference between (3) and (4) we arrive at (5) and (6) shows that the amplitude \( A \) can be recovered by the ratio.

\[ A \sin(2\pi ft_1) + V_{\text{DC}} = \sin(2\pi ft_1 + \phi) + V_{\text{DC}}' \]  \[ A \sin(2\pi ft_2) + V_{\text{DC}} = \sin(2\pi ft_2 + \phi) + V_{\text{DC}}' \]  

\[ A \sin(2\pi ft_1) - \sin(2\pi ft_2) = \sin(2\pi ft_1 + \phi) - \sin(2\pi ft_2 + \phi) \]  

\[ A = \frac{[\sin(2\pi ft_1) - \sin(2\pi ft_2)]}{[\sin(2\pi ft_1 + \phi) - \sin(2\pi ft_2 + \phi)]} \]

D. Comparator Calibration

Non idealities in the comparators such as DC offsets, threshold voltage variations and effects of implementation such as the PCB setup, reference source and phase delaying will lead to fixed amplitude errors. These were calibrated by measuring the number of clock cycles counted by the capture unit with sources directly connected to the comparator, i.e., with a zero ohm load.
Fig. 2. (a) Waveforms in normal operating mode and high precision mode. In this example the amplitude of the sensed waveform is changed depending on measurement electrodes. Multiplexer was changed its configuration at 1, 5, and 9 ms, respectively. T1 is the time between the positive to negative zero crossing of the phase delayed reference and when it crosses the positive value of the sensed voltage. T2 is the same for the negative case, recorded to remove DC offset. T3 and T4 are the equivalent over 2 periods as an example of the high precision mode by averaging. T5 to T8 are again the equivalent for the sensed waveform coming from the different measurement electrode and are visibly shorter than T1 to T4. (b) Time converted waveforms following the comparators.

E. Operating Modes

The system can be operated in three operating modes that trade-off precision for measurement speed.

1) Normal Operating Mode: In the normal operating mode each channel is converted over a single period. An example with amplitude decreasing is shown in Fig. 2(a). It can clearly be seen that the times between the rising edges of Capture A and B decrease (T5 is shorter than T1), Fig. 2(b).

2) High Precision Mode: When the amplitude is constant over more than one period we can increase our precision by counting the increased time differences, T3 and T4 or T6 and T8 in Fig. 2(a) and (b). This is similar to averaging with the advantage that the large, 32-bit counter in the microprocessor may be used to accumulate the time over multiple samples.

3) Fast Acquisition Mode: In high channel count systems it may be desired to multiplex several electrodes to each measurement channel. Typically in bioimpedance imaging systems there are many voltage recording electrodes situated between current injection electrodes. The electric field decreases rapidly away from the current injection electrodes. This can cause the amplitude of the measured voltages to rapidly decrease. These voltages depend mainly on electrode location, boundary shape and the fixed tissue background so their range can be estimated prior to measurements. The monotonically increasing amplitudes correspond to increasing time difference between the crossing time point and the reference point so these may be allocated time multiplexed positions within a single period. An example with two electrodes is shown in Fig. 3(a). T1 and T2 correspond to the smaller amplitude of electrode 1. The sensed voltage is then switched to electrode 2 where T3 and T4 measure a longer time delay from Capture A, due to the increased amplitude as shown in Fig. 3(b).

F. Implementation

We implemented the system with a high speed microprocessor (Delfino EVM board included TMS320C28346) running at 300 MHz. The resolution of capture units is 1/(300 MHz), i.e., 3.333 nsec. The system relies on the performance of the comparators so we built a custom PCB for testing a number of comparators. In the sensing unit we used buffers on each of 16 electrode inputs for high input impedance and a 16 channel multiplexer (ADG1406, Analog Devices, USA). The operating modes, normal, precise or fast defined the operation of the multiplexer. All 16 channels connected to different electrodes, respectively. In high precision mode the channels sample each electrode for more than one period (in normal mode) and in fast mode the multiplexer switches between electrodes within each source waveform period.

The TMS320C28346 with 6 enhanced capture modules allows the calculation of 6 crossing point times as the reference
time point is common. In our design they can work together to reduce the measurement time (fast mode) or increase the precision (high precision mode). The capture interrupts are sensitive to any noise at the comparator outputs, therefore we operate the capture units in one shot mode to escape the noisy transition periods when the multiplexer switches. We also used the input qualifier provided with the microprocessor as a non-linear digital noise rejection filter to suppress glitches.

In the microprocessor capture unit, there is a 32 bit counter running at 300 MHz in free running mode. When detecting a rising or falling edge, the counter value is stored in a first-in-first-out (FIFO) buffer and can be used to generate an interrupt. Here, we used the edge detection of the signal Capture A (which indicates the reference time point) to generate an interrupt to read the FIFO contents of both capture units.

The microprocessor can communicate with the PC to transmit data and receive commands for control of the ATC circuit. We implemented USB and Bluetooth communication methods in a PC communication module. The speed of USB and Bluetooth were 930 kbps and 230 kbps, respectively.

In initial tests we found that the system is highly sensitive to power supply noise as this produces glitches at the comparator outputs. Therefore in all tests we used the most stable power supply available to us. This was a power supply developed in the lab based on a medically isolated Switched Mode Power Supply (ECM100US07, ECM100US09, XP Power Limited, Singapore) and precision regulators (TL431, Texas Instruments, USA) with the following characteristics: Line regulation and load regulation error 0.05%. Ripple rejection ratio over 105 dB. Noise less than -100 dB up to 100 kHz.

G. Experimental Setup

First, we evaluated the performance of three candidate comparators, HA4905 (Intersil, USA), LM319 (Fairchild Semiconductor, USA) and AD8564 (Analog Devices, USA) in fast acquisition mode on a simple resistor phantom. The resistor phantom consisted of fifteen 18Ω resistors with 1% tolerance soldered together in a serial line. We applied a 2Vpp, 500 Hz sine wave voltage to the source point and measured voltages from each of the 16 measurement points using the proposed amplitude-to-time conversion method. These voltages were measured simultaneously using the 16-to-1 analog multiplexer, ADG1406. The reference waveform lagged 45 degrees behind the source signal. Each time the comparator outputted a rising edge or falling edge, the microprocessor captured the time lag between the cross point and reference point, and calculated the difference as the amplitude. The percentage error from the known impedance was calculated after applying comparator calibration and DC offset calibration. We compared the mean SNR over 512 repetitions. The SNR was measured as the mean recorded amplitude divided by the standard deviation, (7).

$$\text{SNR} = 20 \times \log \frac{\text{average}}{\text{standard deviation}} \text{ [dB]} \quad (7)$$

The best performing comparator was compared with a 24-bit audio sampling card and tested in all three modes: fast, normal and high precision over a range of resistor values.

<table>
<thead>
<tr>
<th>Comparator</th>
<th>Count at rising edge (ideally 300,000.0)</th>
<th>Count at falling edge (ideally 600,000.0)</th>
</tr>
</thead>
<tbody>
<tr>
<td>HA4905</td>
<td>304,325.7</td>
<td>599,292.0</td>
</tr>
<tr>
<td>calibration factor</td>
<td>4,325.7</td>
<td>-79.2</td>
</tr>
<tr>
<td>LM319</td>
<td>304,430.8</td>
<td>599,794.3</td>
</tr>
<tr>
<td>calibration factor</td>
<td>4,430.8</td>
<td>-205.7</td>
</tr>
<tr>
<td>AD8564</td>
<td>298,912.0</td>
<td>399,606.5</td>
</tr>
<tr>
<td>calibration factor</td>
<td>-1,088.0</td>
<td>-200,393.5</td>
</tr>
</tbody>
</table>

III. RESULTS

In fast mode, the comparators had the following mean SNR: LM319 37.9 dB, AD8564 34.7 dB and HA4905 40 dB. For the three comparators measured, the SNR varied by less than 0.5 dB over 512 repeated measurements.

The comparator calibration results are shown in Table I for 500 Hz. The HA4905 provided the best performance which may be due to its low offset voltage, current and crosstalk. The LM319 is a widely used comparator without any specific circuitry for high speed or low noise operation. The AD8564 has very low latency of 7 ns which caused it to produce more glitches than other comparators and reducing the measured SNR. This was also evident from our calibration values where the falling edge was often missed, causing the calibration value for the falling edge to be much larger than the other comparators.

We choose the HA4905 as the best comparator and tested its performance against a 24 bit ADC. The best SNR of the ATC with HA4905 was 112.4 dB over 256 periods which compared well with the 24 bit ADC which had an SNR of 113 dB. As expected the SNR of the ATC system with the HA4904 comparator degraded gracefully with less periods: 32 periods 96.1 dB, 8 periods 90.3 dB, 4 periods 87.2 dB, in normal mode (1 period, 81.4 dB) and in fast mode (40 dB).

With 1600 measurement channels at 500 Hz in normal acquisition mode, all measurements were finished in 0.533 seconds. In high precision mode, the measurement time will increase by the number of successive periods measured and the SNR will be improved.

IV. DISCUSSION AND CONCLUSION

We have presented an accurate amplitude measurement system that requires few components and is therefore attractive for developing high channel count systems. The conventional technique of using a high resolution ADC requires enough samples per period to satisfy the Nyquist sampling while the presented ATC technique only requires two samples per period. For computation the ATC technique needs two subtract operations and one division operation. This is efficient compared to the conventional digital phase sensitive demodulation technique which requires many more computations to achieve the same result. In keeping the component count low, the technique may be robust to temperature and electromagnetic noise. In this paper we propose this technique only for the measurement of the magnitude of the impedance signal. It relies on a fixed phase in the ‘phase delay unit’ and has only been tested here for a sinusoidal sensed waveform with no distortion, nonlinearity or significant noise. These conditions may not exist in some.
bioimpedance applications particularly where the sensed signal is very small or when there are artefacts due to electrode movements or changes in the contact impedance. At low frequencies (<1 MHz) the permittivity of tissue is small so we can expect to keep a good relative phase for the phase reference signal. In this paper we chose a phase lag of 45 degrees for the phase of reference signal. This was a compromise between the theoretical best phase lag of 0 degrees which would be affected by noise in the comparators and 90 degrees which will produce a small time difference. In future work the relationship between these variables needs to be investigated.

The high precision stems from using the fine time resolution offered by the microprocessor and in presenting the amplitude information as a time difference between two analog signals. This subtraction process is able to remove DC offset and decrease noise. Comparator calibration can compensate for non-ideal characteristics of the comparator and source. We also expect advantages in developing synchronous systems as we do not rely on multiplexing many channels into a high performance ADC, rather we simplify the amplitude measurement and are able to dedicate one of these systems to a single or small number of channels. The high precision mode resulted in the expected increase in SNR of approximately 6 dB for each period doubling.

We have an interest in developing large channel count systems for a planar EIT system for breast cancer detection from impedance images that would feature 2400 channels and microscopy systems that would require over 8000 channels [34], [35]. Clearly the complexity of the recording channel is very important in this design.

REFERENCES


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