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vice-versa. The benefit of utilizing a constant current excitation is that the large currents, due to low SUT impedances, are prevented, which allows for a stable power consumption and safety. These two elements enable the estimation of the SUT transfer function by using the system identification technique [18]. In this work, we have adopted the linear system identification due to the small signal behavior across the SUT. Nevertheless, the proposed method can be expanded to nonlinear system identification since the post-processing is performed on the sampled time-domain data and implemented fully in the digital domain, regardless of the analog measurement setup.

A. Linear System Identification

The impedance of the SUT is calculated using the adaptive algorithms from the *System Identification Toolbox* in MATLAB (r2012a). The purpose of an adaptive algorithm is to calculate the minimum of a cost function, $I(\underline{\theta})$, which corresponds to finding the coefficients of an equivalent SUT transfer function. The cost function is generally given by

$$I(\underline{\theta}) = \frac{1}{2} \underline{\theta}^T \underline{X}^T \underline{X} \underline{\theta} + (-\underline{X}^T \underline{y})^T \underline{\theta} + \underline{y}^T \underline{y}, \quad (1)$$

where $\underline{\theta}$ represents the coefficients of the SUT transfer function, and $\underline{X}$ is the regression matrix for N measurements. For instance, in the ARX (autoregressive with exogenous input) adaptive algorithm, which is the most widely used in system identification [18], the regression matrix is defined as

$$\begin{bmatrix} u(m) & \dots & u(1) & -y(m) & \dots & -y(1) \\ u(m+1) & \dots & u(2) & -y(m+1) & \dots & -y(2) \\ \vdots & \vdots & \vdots & \vdots & \ddots & \vdots \\ u(N-1) & \dots & u(N-m) & -y(N-1) & \dots & -y(N-m) \end{bmatrix}, \quad (2)$$

where m corresponds to the order of the filter, while u and y correspond to integer samples of the excitation and response waveform, respectively. By finding the derivative of (1) concerning $\underline{\theta}$, the minimum of the cost function can be expressed as

$$\frac{\partial I(\underline{\theta})}{\partial \underline{\theta}} = -\underline{X}^T (\underline{y} - \underline{X} \underline{\theta}) = 0. \quad (3)$$

Since the excitation and response waveform are contaminated with noise, it is only possible to have an estimate of the equivalent SUT transfer function coefficients, $\hat{\underline{\theta}}$, which are found by

$$\hat{\underline{\theta}} = (\underline{X}^T \underline{X})^{-1} \underline{X}^T \underline{y}. \quad (4)$$

Having $\hat{\underline{\theta}}$, an equivalent IIR digital filter, which matches the measured response of the SUT in time, can be constructed.

B. Prototype Implementation

The block diagram of the measurement setup and the prototype are shown in Fig. 1. The setup consists of a linear feedback shift register (LFSR), a 3-level DAC, a 12-bit ADC, a pseudo-differential amplifier, and a logic analyzer.

At the rising edge of f_{CLK} , the LFSR generates the excitation waveform, $D_{in<1:0>}$, which is a 2-bit pseudo-random sequence since it is deterministic and periodic. This means

that given the present state of the register, every next state can be easily calculated and the sequence repeats after all the combinations are cycled through.

The LFSR is implemented using shift registers (SN74LV164A) and a 2-input XOR gate (SN74LVC1G86). The period of the pseudo-random noise waveform is 255 clock cycles. In this way, the LFSR generates an approximate white noise waveform from the outputs of the first two flip-flops ($D_{in<1:0>}$), which control the 3-level DAC [8].

$D_{in<1:0>}$ controls a 4/1 analog multiplexer (ADG804), which functions as a 3-level DAC, and selects either V_{REF-} , V_{CM} , or V_{REF+} , as shown in Fig. 1a. The V_{REF+} and V_{REF-} correspond to the voltage reference levels, which are supplied directly from the ADC and decoupled via a voltage buffer (AD8652). The resulting waveform consists of positive and negative pulses of V_{REF} amplitude centered around the common-mode value $V_{CM} \approx 1.6$ V, which is obtained by averaging $V_{REF+} \approx 2.6$ V and $V_{REF-} \approx 0.6$ V. Hence, the DAC and ADC have the same full-scale, $V_{FS} = V_{REF} = 1$ V. The voltage output of the analog multiplexer, V_Y , is then converted into an excitation current, $I_{exc} \approx 9.88 \mu\text{A}$, by R_{fuse} , resulting in the V_{SUT+} voltage at the output of A_1 , which is

$$V_{SUT+} = V_{CM} - \left[\frac{(R_{stb} + Z_{SUT}) || R_{bias}}{R_{fuse}} \right] (V_Y - V_{CM}), \quad (5)$$

where Z_{SUT} is the SUT impedance highlighted in Fig. 1a. If A_1 malfunctions, the current through SUT is limited by R_{stb} .

The SUT is connected in parallel with $R_{bias} \approx 85.6$ k Ω to ensure that the signal swing across the SUT can not exceed the unloaded (or open) condition and that a proper DC operating point is established for A_1 in the case of purely capacitive SUTs. By placing the SUT in the feedback network of A_1 , a simple circuit solution is achieved for constant current excitation without additional components.

The chosen feedback configuration introduces stability issues since the employed operational amplifier A_1 (AD8652) cannot drive large capacitive loads (≤ 1 nF) without a damping resistor. Hence, the maximum loading capacitance for A_1 , $C_{sut} \approx 930.9$ pF (1 nF $\pm 10\%$), is used in series with R_{stb} resistor for stability. The impact of R_{stb} is analyzed using the small signal schematic in Fig. 2, where the SUT is represented as an RC network. The $C_{p1} \approx 100$ pF is the sum of the input parasitic capacitances of amplifiers A_1 and A_2 . The $C_{p2} \approx 13.4$ pF is the measured parasitic capacitance between the input and output pins of the soldered amplifier and the SUT interconnects. The $C_L \approx 50$ pF models the load, which represents the on-board decoupling and ADC sampling capacitance. A resistor, $R_L \approx 22 \Omega$, is added in series with C_L to decouple the ADC's charge transfer transients from A_1 . The $R_{out} \approx 43.2 \Omega$ is calculated based on the reported phase margin in the data-sheet for a defined capacitive load. The resistive component of the SUT, $R_{sut} \approx 98.8$ k Ω , is chosen empirically to bound the signal swing at the ADC input, which prevents quantization overflow due to switching glitches introduced by C_{sut} . For the chosen RC components, the SUT impedance changes gradually in the $(0, f_{CLK}/2)$ bandwidth, which allows for an insightful test-bench since both components contribute similarly to the equivalent SUT impedance, Z_{sut} . The $A(s)$

Fig. 4. a) Photograph of the measurement setup with the prototype, electrodes and physiological solution. b) Block diagram of the pen-probe used for skin impedance measurements.

Afterward, a short-circuit is made at the tip of the ports, which are connected to the SUT. As a result, an uncalibrated short-circuit, $Z_{short,uncal}(z)$, measurement is acquired, which represents the transfer function of a parallel connection between R_{stb} and $Z_{open}(z)$. The presence of R_{stb} is crucial to maintain stable operation, and therefore the phase margin of A_1 , during the short-circuit condition. Moreover, the calibrated short-circuit impedance, $Z_{short}(z)$, is obtained after de-embedding the contribution of $Z_{open}(z)$ in $Z_{short,uncal}(z)$ and adjusting the gain error in $Z_{short}(z)$ to match with R_{stb} .

Having $Z_{open}(z)$ and $Z_{short}(z)$, the transfer function of the SUT without the on-board parasitics is calculated by

$$Z_{sut}(z) = \left[\{ Z_{sut,uncal}^{-1}(z) - Z_{open}^{-1}(z) \}^{-1} - Z_{short}(z) \right], \quad (6)$$

where $Z_{sut,uncal}(z)$ represents the IIR filter obtained from the uncalibrated measurement results, with the SUT connected in the circuit, while the aforementioned $Z_{short}(z)$ is found as

$$Z_{short}(z) = \left(Z_{short,uncal}^{-1}(z) - Z_{open}^{-1}(z) \right)^{-1}. \quad (7)$$

III. EXPERIMENTAL RESULTS

The complete measurement setup with the prototype is shown in Fig. 4a. The proof-of-concept prototype, shown in Fig. 1b, is implemented on a four-layer PCB, and it occupies an area of 7.2x7.6 cm². All impedance measurements are performed using a 2-terminal electrode configuration. Before each measurement, the open- and short-circuit calibration are performed, and the offset in the $D_{out<11:0>}$ time-domain data is removed in post-processing.

A. Electrical Characterization

The measurement setup is first validated using a representative set of parallel RC values with an incremental change in capacitance from 233.2 pF to 930.9 pF. Upon collecting all the responses for various RC values, which are shown in Fig. 5, several algorithms were tested while the order of the IIR filter is maintained the same across all cases. The ARX algorithm showed slightly better estimation fit to the time-domain response ($\approx 96.86\%$) compared to others, such as autoregressive

Fig. 5. Normalized output time-domain responses obtained from the open- and short-circuit calibration measurements and various RC values.

Fig. 6. Normalized power spectral density of the open-circuit voltage waveform.

moving average with exogenous input ($\approx 96.63\%$), output-error ($\approx 95.38\%$), Box-Jenkins ($\approx 96.68\%$). The estimation fit represents in percentage the normalized root mean squared error, which is a measure of how well the response of the IIR filter fits the measurements. Hence, ARX has been chosen considering its accuracy and simplicity. A minimum of four coefficients is required (2 in the numerator and 2 in the denominator) to maintain a high estimation fit (96%). The estimation frequency range below 5 kHz is not fully utilized due to the lower power spectral density in the open-circuit waveform, as it can be seen in Fig. 6. Although the shown spectrum corresponds to the open-circuit voltage response of the SUT, the I_{exc} is identical to the open-circuit waveform after dividing with R_{bias} . Additionally, the presence of low-

Fig. 7. Calibrated magnitude and phase responses of digital IIR filters based on electrical characterization.

Fig. 8. Calibrated magnitude and phase responses of digital IIR filters based on electrochemical testing.

frequency noise (e.g. flicker noise, common-mode and power supply interference) in the measurement setup reduces the frequency range. As it can be seen in Fig. 7, the 5kHz ($0.05f_{CLK}$) to 50kHz ($0.5f_{CLK}$) frequency range has the best estimation fit to an ideal RC model. The behavior of the ideal RC model is represented by its equivalent discrete-time transfer function to appropriately compare it with the response of $Z_{sut}(z)$. Furthermore, V_{SUT+} is more sensitive to smaller SUT impedance values due to the R_{bias} shunt.

B. Electrochemical Testing

For the electrochemical testing, two gold rectangular electrodes (ENIG Ni/AU process on a 0.75 mm hydrocarbon ceramic substrate Rogers4000) were used [11]. The rectangular electrodes (with an area of $3 \times 1 \text{ mm}^2$) are first fully submerged in a glass container filled with sodium-chloride physiological solution (0.9% NaCl, $\sigma \approx 1.584 \text{ S/m}$, $\epsilon_r \approx 78$). The impedance magnitude of the electrodes and physiological solution is then measured using an LCR meter (HP4284A) across the 5 kHz to 50 kHz bandwidth in 5 kHz increments to compare it with the data obtained with the proposed setup. The LCR results are then fitted to a second-order continuous-time model, which is then converted in MATLAB to an equivalent

discrete-time model. In this way, the response of IIR filters, which are obtained with the system identification technique, can be appropriately compared with the LCR measurements. The number of IIR filter coefficients in the numerator and denominator are doubled compared to the test-bench with passive RC devices in order to obtain a better fit with the response of the physiological solution. The results of the electrochemical testing for all approaches are shown in Fig. 8. In the system identification approach, the measurements are repeated four times in succession to test for repeatability, while in the case of the LCR meter, four times averaging and integration is applied. The magnitude responses of the IIR filter follows a similar trend, however the phase response diverges from the behavior of the physiological solution, indicating a bottleneck in the measurement method.

C. Skin impedance measurements

The skin impedance measurements are performed by using a pen shaped probe with a spring mechanism as shown in Fig. 4b. The two gold electrodes are glued to the tip of the pen-probe. For the applied external force, F_{ext} , the elasticity of the spring, k , will resist with the same force intensity, F_e , for the same displacement, Δx . To reduce the variability between

Fig. 9. Calibrated magnitude and phase responses of digital IIR filters based on non-invasive skin impedance measurements.

experiments, the probe is always pressed to its maximum displacement.

The skin is first cleaned with alcohol and dried before applying the sodium-chloride solution. After approximately 4 minutes, a steady-state contact is observed and multiple measurements are taken, first with the LCR meter followed by the proposed measurement method without breaking the electrode-skin contact. Similarly, as in the electrochemical test, the number of IIR filter coefficients is doubled to better model the response of the skin. As it can be seen in Fig. 9, the magnitude responses are matching much better than in the previous test though they do not exactly overlap. As it was the case in the electrochemical test-bench, the adaptive algorithm fails to capture a proper phase response, which indicates that the phase of the bio-impedance may be too complex to be measured with the proposed method.

IV. DISCUSSION

The proof-of-concept prototype has been used to demonstrate and validate an impedance measurement method in a focused frequency range. The experimental data shows that the main bottleneck in the measurement accuracy is the estimation fit of the adaptive algorithm. Therefore, in future work, the ADC resolution can be relaxed to 10-bit, and considering the achieved frequency range, a low-speed ADC architecture can be employed. A Nyquist rate ADC is more advantageous for system identification compared to an oversampling ADC, since the decimator is omitted from the transfer function of the measurement setup. The maximum ADC sampling frequency is limited by the parasitic pole in the pseudo-differential voltage amplifier. Specifically, an additional pole was observed at 140 kHz in the open-circuit measurements, which corresponds to the contribution of C_{p2} and R_{bias} . As a result, the complexity of the complete measurement setup, shown in Fig. 1a, should not exceed a second-order IIR filter transfer function up to $f_{CLK}/2$, otherwise the algorithm's estimation fit deteriorates. To mitigate this issue, the SUT can be placed in an open-loop configuration using a differential current-steering DAC [15], [16] and a wide-band instrumentation amplifier [9], [25]. In this way, the stability issue can be avoided. On the other hand, the lower frequency value is limited by the period and the shape of the excitation waveform,

which has a low power spectral density at low-frequencies. Currently, the characterization of the SUT is not influenced by the measurement latency introduced by the post-processing in MATLAB. However, a dedicated FPGA will be required for running the adaptive algorithm in a closed-loop system. Lastly, the biomedical test-benches showed that the phase response can not be accurately captured. Nevertheless, the phase sensitivity could be potentially improved with further design optimization since the measurement method demonstrated that it can capture the phase response of simple RC components.

V. CONCLUSION

In this paper, a measurement method for impedance spectroscopy based on adaptive filters, specifically linear system identification technique, has been presented. A proof-of-concept prototype was designed with off-the-shelf components taking into account their limitations. In addition, guidelines for ensuring stable operation were provided. The measurement data has been post-processed using the System Identification Toolbox in MATLAB, which enabled the design space exploration of various algorithms from which the ARX topology was chosen. An open/short calibration procedure was employed to mitigate the impact of non-idealities in the prototype. In the case of discrete RC components, an equivalent SUT response was achieved with a second-order IIR filter with 4 coefficients, which provided the best trade-off between model complexity and estimation fit. Throughout the electrical characterization, the selected ARX filter topology achieves a 96% estimation fit over a (5, 50) kHz frequency range. In the case of bio-impedance, the complexity of the IIR transfer function was increased to 8 coefficients to improve the fitting. However, the electrochemical and skin impedance results showed that the proposed method could not match the phase characteristic. Therefore, the presented measurement method can provide a low-complexity circuit implementation for acquiring the magnitude response of a biological SUT.

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