Bandwidth and Common Mode Optimization for Current and Voltage Sources in Bioimpedance Spectroscopy

Tobias Menden¹,²,³, Jascha Matuszczyk¹,², Steffen Leonhardt² and Marian Walter²

1. Shared first authorship
2. Medical Information Technology, RWTH Aachen University, Pauwelsstr. 20, 52074 Aachen, Germany
3. Email any correspondence to: menden@hia.rwth-aachen.de

Abstract
Bioimpedance measurements use current or voltage sources to inject an excitation signal into the body. These sources require a high bandwidth, typically from 1kHz to 1MHz. Besides a low common mode, current limitation is necessary for patient safety. In this paper, we compare a symmetric enhanced Howland current source (EHCS) and a symmetric voltage source (VS) based on a non-inverting amplifier between 1kHz and 1MHz. A common mode reduction circuit has been implemented in both sources. The bandwidth of each source was optimized in simulations and achieved a stable output impedance over the whole frequency range. In laboratory measurements, the output impedance of the EHCS had its -3dB point at 400kHz. In contrast, the VS reached the +3dB point at 600kHz. On average over the observed frequency range, the active common mode compensation achieved a common mode rejection of -57.7dB and -71.8dB for the EHCS and VS, respectively. Our modifications to classical EHCS and VS circuits achieved a low common mode signal between 1kHz and 1MHz without the addition of complex circuitry, like general impedance converters. As a conclusion we found VSs to be superior to EHCSs for bioimpedance spectroscopy due to the higher bandwidth performance.

Keywords: bioimpedance spectroscopy, enhanced Howland current source, voltage source, common mode reduction, output impedance, harmonic distortion

Introduction
In bioimpedance spectroscopy (BIS), the electric conductivity of a tissue is measured via electrodes, whose impedance depends on their geometry and composition [1]. A known alternating current from a signal source is injected into the tissue under test and the resulting voltage is measured. To minimize the impact of the skin impedance, electrode impedance and the measurement system on the acquired data, often a tetra-polar measuring method is used, which requires a second set of electrodes to measure the voltage voltage drop on the bioimpedance [2, 3].

Due to its simplicity, BIS lays the foundation for more complex and medically relevant applications like electrical impedance tomography (EIT), as well as impedance pneumography and impedance cardiography. The former are used for harmless, real-time bedside pulmonary monitoring and are still subject to research [4].

The impedance of human tissue itself is frequency-dependent and also tissue-specific. Its characteristics in the frequency domain can be divided into three dispersions (α, β, and γ). Each can be described with a Cole-Cole impedance model [5, 6]. The dispersion in which the majority of cellular structures influences the tissue impedance is the β-dispersion (1kHz to 1MHz). Here, the impedance ranges from a few 10kΩ to approx. 100Ω [3, 7].

For a successful measurement, the signal of the source should not change the electrical characteristics of the probed tissue. Further, the injected signal has to be of low magnitude to prevent harm to the patient. Due to the need to inject low currents at low frequencies, galvanic coupled electrodes are used. Common mode signals easily evoke polarisation and therefore corrosion at the electrode skin interface [8, 9]. To prevent skin lesions, measurement artefacts and distortions, the common mode signal should be as low as possible [3].

In BIS, the source for the alternating excitation signal traditionally is a bidirectional “voltage-controlled current source” (VCCS). By design, the main advantage is the limited and adjustable current, with the result...
that maximum allowable current limits can inherently be achieved. Different current source topologies for floating and grounded loads exist like operational transconductance amplifier, mirrored current-conveyor [10, 11], load-in-the-loop, Tietze and Howland topologies [3].

Often, BIS applications use an Enhanced Howland Current Source (EHCS) with grounded load. Its simple layout with one op-amp and five matched resistors provides a constant current with a high output impedance independent of the load. Most simulated current sources achieve a high output impedance [3]. Above 100 kHz, however, real implementations of the EHCS have stability problems caused by mismatching and tolerances of the resistors, stray-capacitance and frequency-dependent characteristics of the active elements. As a consequence, the output impedance drops at higher frequencies [3, 12] and also the common mode signal increases [8]. In in-vivo measurements, the stability and performance of the EHCS continues to degrade further due to the capacitive configuration of the load caused by the electrode contact impedance, cable capacitance and the capacitance to ground [13].

A traditional EHCS for portable BIS applications by Xu et al. achieved an output impedance of 100 kΩ up to 100 kHz [14]. The proposed design became unstable for higher frequencies. The stability of EHCSs can be increased by compensation capacitors. For example, Nouri et al. achieved up to 2.8 MΩ at 1 MHz [15]. Recently, Saulnier et al. presented an FPGA-based adaptive algorithm to adjust the EHCS output current achieving an output impedance of 7 MΩ at 1 MHz [16]. Another approach to increase the bandwidth and reduce the common mode of EHCS topologies is the usage of symmetric current injections. Sirtoli et al. presented a symmetric VCCS achieving a constant 1 MΩ output impedance up to 300 kHz. Also, the output impedance shortened to 150 kΩ at 1 MHz [8]. The reduced performance at higher frequencies can be also challenged by more complex designs, like the general impedance converter.

Instead of a current source, a tetra-polar BIS measurement can also be performed with a ‘voltage-controlled voltage source’ (VCVS), if the resulting current is measured and used for the impedance calculation. A VS does not limit the maximum output current and does not limit itself to a maximum allowable current by design. Further modifications or actions have to be taken to ensure patient safety. Qureshi et al. presented a symmetric VCVS, which addresses both issues with an output impedance ranging from 7.2 Ω to 13.2 Ω up to 20 MHz [17].

This work compares a modified EHCS from Sirtoli et al. and a modified VS for BIS measurements from Qureshi et al. for floating loads in the frequency range from 1 kHz to 1 MHz. The key challenge of bandwidth improvement and common mode reduction has been addressed by simple and easy to implement circuitry modifications. Furthermore, the complexity, benchmarks and tolerances of both sources were chosen similar to allow a fair comparison.

**Materials and methods**

**Requirements for BIS measurements**

A typical BIS measurement system consists of a front-end, a data acquisition unit and a data processing unit. The front-end generates the excitation signal, typically by a VCCS, and measures the resulting voltage. Measurements are mainly performed in the beta-dispersion range from 1 kHz to 1 MHz and should have a measurement accuracy of 0.1% or higher [18]. The output impedance of a source determines the stability of the generated signal with respect to the connected load. To achieve a measurement accuracy of 0.1% for loads between 100 Ω and 10 kΩ [7], a minimum output impedance of at least 1 MΩ in the mentioned frequency band is necessary. To obtain the same accuracy for a voltage source (VS), the output impedance should be at a maximum of 1 Ω.

For safe and continuous operation on biological tissue, the standard EN 60601-1-1 defines the following maximum allowable currents

\[
I_{\text{max}}(f) = \begin{cases} 
100 \mu A & f \leq 1 \text{kHz}, \\
100 \mu A \cdot \frac{1}{1 kHz} & 1 \text{kHz} < f < 100 \text{kHz}, \\
10 mA & f \geq 100 \text{kHz}. 
\end{cases}
\]  

(1)

In addition, the common mode current has to be below 10 μA to prevent electrode polarisation and harm to the patient. In literature, this particular demand often is neglected. However, Pliquett et al. presented an active common mode rejection circuit for a symmetrical EHCS and achieved a common mode rejection below -80 dB [9].

**Symmetric enhanced Howland current source**

As mentioned before, in BIS most commonly the EHCS is employed, along with current mirror circuits and multiple feedback operational amplifier circuitry [19]. Theoretically, an EHCS can achieve a very high output impedance $Z_{\text{out}}$. In reality, this source suffers from low bandwidth, the ability to drive capacitive loads and low common mode rejection. To overcome such limitations, we chose an improved EHCS by Sirtoli et al. [8] as a basis for further modifications.

The basic functionality of the circuitry, depicted in Figure 1, is an EHCS consisting of $R_1$ to $R_4$ and $R_x$.

The resistor $R_x$ defines and limits the output current. Assuming a perfect symmetric current injection

\[
|U_{L+}| = |U_{L-}| = \frac{|U_{\text{load}}|}{2},
\]  

(2)

we can apply the traditional balancing conditions of the EHCS as given in [8] to achieve a high $Z_{\text{out}}$:

\[
R_1 = R_3 + R_x,
\]  

(3)

\[
R_2 = R_3 = R_4.
\]  

(4)
We note that the advantage of the circuitry as depicted in Figure 1 is the symmetric current injection from a fully differential amplifier (OP1).

The load is driven from \( U_{L+} \), to \( U_{L-} \), where the signal at \( U_{L-} \) is a phase shifted copy of \( U_{L+} \). This behavior is equivalent to two current sources working in a push-pull configuration as source or sink. As a result, each source has to generate only half of the voltage drop across the load. Theoretically, this doubles the output impedance and the maximum output voltage (compliance) of the current source [8, 20]. Additionally, the circuitry in Figure 1 drives a quasi-floating load, due to the resistor \( R_5 \) connected to ground. The connection to ground could increase the common mode voltage and reduce patient safety. This issue could be addressed by a separate monitoring circuit for \( I_{\text{out}} \).

Other drawbacks of the traditional EHCS include instabilities due to the positive and negative feedback. Such oscillations may occur only at high frequencies and can be compensated by additional capacitors in parallel with \( R_1 \) and \( R_4 \) [21]. However, in this paper we used a capacitor \( C_c \) in parallel with \( R_3 \) to increase the bandwidth of the EHCS. To simplify further calculations, \( C_c \) will be neglected initially. The symmetric current injection requires an equivalent load at \( U_{L+} \) and \( U_{L-} \), independent of \( Z_{\text{load}} \). Therefore, the resistor \( R_5 \) has to mimic the voltage drop across \( R_1 \) and \( R_3 \). Furthermore, the voltage drop over \( R_1 \) has to be identical to the one over \( R_x \). This leads to the additional balancing conditions:

\[
R_5 = R_3 + R_1 = R_x + 2R_3. \quad (5)
\]

Because of the connection of \( R_5 \) to ground, the symmetrical current source is quasi floating and creates a load dependency of \( R_1 \) as presented in [8]. To accomplish the identical voltage drop over both current generating resistors \( R_x \) as well as \( R_1 \), \( R_1 \) has to be

\[
R_1 = \frac{(Z_{\text{load}} + 2R_5)R_x}{2R_5}. \quad (6)
\]

According to eq. (5), large values of \( R_x \) and \( R_3 \) lead to an increased value of \( R_5 \) and thus, \( R_1 \) becomes less dependent on \( Z_{\text{load}} \).

Let \( A_{\text{OL}} \) be the open loop gain. Due to the above conditions, the output current is load-independent and calculated as follows

\[
I_{\text{out}} = \frac{A_{\text{OL}}U_{\text{in}}}{(A_{\text{OL}} + 2)R_x} A_{\text{OL}} \gg 2 \frac{U_{\text{in}}}{R_x}. \quad (7)
\]

In case of \( A_{\text{OL}} \gg 2 \), \( I_{\text{out}} \) can be determined by the input voltage \( U_{\text{in}} \) and \( R_x \). As a result of the balancing condition eqs. (3), (4) and (5), the output impedance is solely dependent on \( R_3 \) and \( R_x \) as well as \( A_{\text{OL}} \) and can be calculated as followed [8]:

\[
Z_{\text{out}} = \frac{R_x^3 + (2A_{\text{OL}} + 7)R_xR_3^2 + (2A_{\text{OL}} + 6)R_3^2R_x}{3R_x^2 + 6R_3R_x + 3R_3^2}. \quad (8)
\]

The output impedance is dominated by the open loop gain value of OP1. By choosing \( R_x < R_3 \), the output impedance is determined mainly by \( R_3 \), whereas the output current is set by \( R_x \), according to eq. (7).

The output voltage swing or the saturation voltage \( U_{\text{sat}} \) of the operational amplifier limits the maximum operable load \( Z_{\text{load, max}} \). According to Figure 1, the output voltage \( U_{\text{out}} \) of OP1 is given as the voltage drop across \( R_x \) and \( Z_{\text{load}} \), which is always smaller than the saturation voltage of OP1 [8]:

\[
U_{\text{out}} = I_{\text{out}}R_x \left[ 1 + \frac{Z_{\text{load}}}{2R_x} \left( 1 + \frac{R_x}{R_3} \right) \right] \leq U_{\text{sat}}. \quad (9)
\]

Thus, the theoretical maximum operable value of \( Z_{\text{load}} \) becomes

\[
Z_{\text{load, max}} = \left( \frac{U_{\text{sat}}}{I_{\text{out}}R_x} - 1 \right) \frac{R_x}{R_3 + 2R_3}. \quad (10)
\]

The maximum load increases for smaller values of \( R_x \). Additionally, the term \( A \) from eq. (10) increases the maximum operable load for \( R_x < R_3 \). Thus, \( R_3 \) has the potential to double \( Z_{\text{load, max}} \) without changing \( I_{\text{out}} \) [8].

**Parameter selection for the EHCS**

In our application we have chosen the fully differential amplifier THS4151 (Texas Instruments, Dallas, USA) to generate the symmetrical output signal. The THS4151 has an open-loop gain \( A_{\text{ol}} \) of 67 dB, which is sufficient to determine the output current by \( U_{\text{in}} \) and \( R_x \), according to eq. (7).

The choice of \( R_x \) and \( R_3 \) determine the output impedance according to eq. (8). Figure 2 shows the
The maximum operable load can be used to determine $R_x$. We further assume a saturation voltage of 12.7 V (THS4151) and an output current of 2.08 mA. To ensure that the current source can drive a load of up to 6.5 kΩ, $R_x$ can be up to 2.2 kΩ according to eq. (10). In the following, we use $R_x = R_1 = 2.2$ kΩ and $R_3 = R_2 = R_4 = 4.9$ kΩ. $R_1$ and $R_3$ are set according to the balancing eq. (3) and (5) to 7.1 kΩ and 12 kΩ, respectively. Apart from $R_3$ and $R_x$, the output impedance depends on the open-loop-gain $A_{ol}$ of the THS4151, which is 67 dB. These parameters achieve a theoretical output impedance of 2.27 MΩ. However, the $A_{ol}$ has a low-pass characteristic, which lowers the output impedance at higher frequencies. For the THS4151, the frequency-dependency of $A_{ol}$ is not specified in the datasheet. However, the operational amplifier can practically maintain gain values of 40 dB for a bandwidth of 1 MHz. Under the conservative approximation of 40 dB, the output impedance drops down to 104.5 kΩ at 1 MHz.

**Symmetric voltage source**

The basis of the VS, used in this work, is a non-inverting amplifier with a current sensing resistor $R_s$ added between the output of OP1 and the feedback path, similar to the VS presented by Qureshi et al. [17]. The single-ended VS is depicted in Figure 3.

At first, the capacitor $C_v$ will be neglected to simplify the calculation of $U_{out}$ given by

$$U_{out} = U_{in} \left( \frac{Z_{load} + Z_{OA} + R_s}{A_{OL} Z_{load}} + \frac{R_3}{R_1 + R_2} \right)^{-1}.$$  \hspace{1cm} (11)

The maximum operable load can be used to determine $R_s$. We further assume a saturation voltage of 12.7 V (THS4151) and an output current of 2.08 mA. To ensure that the current source can drive a load of up to 6.5 kΩ, $R_s$ can be up to 2.2 kΩ according to eq. (10). In the following, we use $R_x = R_1 = 2.2$ kΩ and $R_3 = R_2 = R_4 = 4.9$ kΩ. $R_1$ and $R_3$ are set according to the balancing eq. (3) and (5) to 7.1 kΩ and 12 kΩ, respectively. Apart from $R_3$ and $R_x$, the output impedance depends on the open-loop-gain $A_{ol}$ of the THS4151, which is 67 dB. These parameters achieve a theoretical output impedance of 2.27 MΩ. However, the $A_{ol}$ has a low-pass characteristic, which lowers the output impedance at higher frequencies. For the THS4151, the frequency-dependency of $A_{ol}$ is not specified in the datasheet. However, the operational amplifier can practically maintain gain values of 40 dB for a bandwidth of 1 MHz. Under the conservative approximation of 40 dB, the output impedance drops down to 104.5 kΩ at 1 MHz.

The maximum operable load can be used to determine $R_s$. We further assume a saturation voltage of 12.7 V (THS4151) and an output current of 2.08 mA. To ensure that the current source can drive a load of up to 6.5 kΩ, $R_s$ can be up to 2.2 kΩ according to eq. (10). In the following, we use $R_x = R_1 = 2.2$ kΩ and $R_3 = R_2 = R_4 = 4.9$ kΩ. $R_1$ and $R_3$ are set according to the balancing eq. (3) and (5) to 7.1 kΩ and 12 kΩ, respectively. Apart from $R_3$ and $R_x$, the output impedance depends on the open-loop-gain $A_{ol}$ of the THS4151, which is 67 dB. These parameters achieve a theoretical output impedance of 2.27 MΩ. However, the $A_{ol}$ has a low-pass characteristic, which lowers the output impedance at higher frequencies. For the THS4151, the frequency-dependency of $A_{ol}$ is not specified in the datasheet. However, the operational amplifier can practically maintain gain values of 40 dB for a bandwidth of 1 MHz. Under the conservative approximation of 40 dB, the output impedance drops down to 104.5 kΩ at 1 MHz.

**Symmetric voltage source**

The basis of the VS, used in this work, is a non-inverting amplifier with a current sensing resistor $R_s$ added between the output of OP1 and the feedback path, similar to the VS presented by Qureshi et al. [17]. The single-ended VS is depicted in Figure 3.

At first, the capacitor $C_v$ will be neglected to simplify the calculation of $U_{out}$ given by

$$U_{out} = U_{in} \left( \frac{Z_{load} + Z_{OA} + R_s}{A_{OL} Z_{load}} + \frac{R_3}{R_1 + R_2} \right)^{-1}.$$  \hspace{1cm} (11)

The maximum operable load can be used to determine $R_s$. We further assume a saturation voltage of 12.7 V (THS4151) and an output current of 2.08 mA. To ensure that the current source can drive a load of up to 6.5 kΩ, $R_s$ can be up to 2.2 kΩ according to eq. (10). In the following, we use $R_x = R_1 = 2.2$ kΩ and $R_3 = R_2 = R_4 = 4.9$ kΩ. $R_1$ and $R_3$ are set according to the balancing eq. (3) and (5) to 7.1 kΩ and 12 kΩ, respectively. Apart from $R_3$ and $R_x$, the output impedance depends on the open-loop-gain $A_{ol}$ of the THS4151, which is 67 dB. These parameters achieve a theoretical output impedance of 2.27 MΩ. However, the $A_{ol}$ has a low-pass characteristic, which lowers the output impedance at higher frequencies. For the THS4151, the frequency-dependency of $A_{ol}$ is not specified in the datasheet. However, the operational amplifier can practically maintain gain values of 40 dB for a bandwidth of 1 MHz. Under the conservative approximation of 40 dB, the output impedance drops down to 104.5 kΩ at 1 MHz.

**Symmetric voltage source**

The basis of the VS, used in this work, is a non-inverting amplifier with a current sensing resistor $R_s$ added between the output of OP1 and the feedback path, similar to the VS presented by Qureshi et al. [17]. The single-ended VS is depicted in Figure 3.

At first, the capacitor $C_v$ will be neglected to simplify the calculation of $U_{out}$ given by

$$U_{out} = U_{in} \left( \frac{Z_{load} + Z_{OA} + R_s}{A_{OL} Z_{load}} + \frac{R_3}{R_1 + R_2} \right)^{-1}.$$  \hspace{1cm} (11)

The maximum operable load can be used to determine $R_s$. We further assume a saturation voltage of 12.7 V (THS4151) and an output current of 2.08 mA. To ensure that the current source can drive a load of up to 6.5 kΩ, $R_s$ can be up to 2.2 kΩ according to eq. (10). In the following, we use $R_x = R_1 = 2.2$ kΩ and $R_3 = R_2 = R_4 = 4.9$ kΩ. $R_1$ and $R_3$ are set according to the balancing eq. (3) and (5) to 7.1 kΩ and 12 kΩ, respectively. Apart from $R_3$ and $R_x$, the output impedance depends on the open-loop-gain $A_{ol}$ of the THS4151, which is 67 dB. These parameters achieve a theoretical output impedance of 2.27 MΩ. However, the $A_{ol}$ has a low-pass characteristic, which lowers the output impedance at higher frequencies. For the THS4151, the frequency-dependency of $A_{ol}$ is not specified in the datasheet. However, the operational amplifier can practically maintain gain values of 40 dB for a bandwidth of 1 MHz. Under the conservative approximation of 40 dB, the output impedance drops down to 104.5 kΩ at 1 MHz.

**Symmetric voltage source**

The basis of the VS, used in this work, is a non-inverting amplifier with a current sensing resistor $R_s$ added between the output of OP1 and the feedback path, similar to the VS presented by Qureshi et al. [17]. The single-ended VS is depicted in Figure 3.

At first, the capacitor $C_v$ will be neglected to simplify the calculation of $U_{out}$ given by

$$U_{out} = U_{in} \left( \frac{Z_{load} + Z_{OA} + R_s}{A_{OL} Z_{load}} + \frac{R_3}{R_1 + R_2} \right)^{-1}.$$  \hspace{1cm} (11)

The maximum operable load can be used to determine $R_s$. We further assume a saturation voltage of 12.7 V (THS4151) and an output current of 2.08 mA. To ensure that the current source can drive a load of up to 6.5 kΩ, $R_s$ can be up to 2.2 kΩ according to eq. (10). In the following, we use $R_x = R_1 = 2.2$ kΩ and $R_3 = R_2 = R_4 = 4.9$ kΩ. $R_1$ and $R_3$ are set according to the balancing eq. (3) and (5) to 7.1 kΩ and 12 kΩ, respectively. Apart from $R_3$ and $R_x$, the output impedance depends on the open-loop-gain $A_{ol}$ of the THS4151, which is 67 dB. These parameters achieve a theoretical output impedance of 2.27 MΩ. However, the $A_{ol}$ has a low-pass characteristic, which lowers the output impedance at higher frequencies. For the THS4151, the frequency-dependency of $A_{ol}$ is not specified in the datasheet. However, the operational amplifier can practically maintain gain values of 40 dB for a bandwidth of 1 MHz. Under the conservative approximation of 40 dB, the output impedance drops down to 104.5 kΩ at 1 MHz.

**Symmetric voltage source**

The basis of the VS, used in this work, is a non-inverting amplifier with a current sensing resistor $R_s$ added between the output of OP1 and the feedback path, similar to the VS presented by Qureshi et al. [17]. The single-ended VS is depicted in Figure 3.

At first, the capacitor $C_v$ will be neglected to simplify the calculation of $U_{out}$ given by

$$U_{out} = U_{in} \left( \frac{Z_{load} + Z_{OA} + R_s}{A_{OL} Z_{load}} + \frac{R_3}{R_1 + R_2} \right)^{-1}.$$  \hspace{1cm} (11)
The symmetric output signals. However, the design from Pliquett uses additional OA in the feedback path which results in a comparable amount of active components. Future work could focus on the reduction of components by omitting the active feedback OAs, which lead to a more miniaturized realization.

**Parameter selection for the symmetric voltage source**

The operating point is a load current of 2.93 mA across a 1 kΩ resistor. Thus, the source should achieve an output voltage of 2.93 V over the entire frequency range and a preferably low output impedance. The THS4631 operational amplifier is used for OA1 and has a saturation voltage of ±13 V and an open loop gain of $A_{OL}$ of 80 dB. The gain of the two VS set by $R_1$ and $R_2$ should be kept as small as possible to reduce the influence of output imbalances. In addition, the current through the resistors $R_1$ and $R_2$ is kept as small as possible to reduce the influence on the connected load. Consequently, the resistor pair $R_1 = 100k\Omega$ and $R_2 = 2k\Omega$ results in an amplification of $A = 1.02$ and has a low bypass current of approx. 28.7 µA. To achieve an $I_{load} = 2.93$ mA, the VS requires an input voltage of $U_{in} = 2.885$ V. For frequencies above 100 kHz, the maximum load-current is 10 mA. Applying eq. (12) and (14), the minimum operable load is 208.8 $\Omega$ with an $R_s$ of 1 kΩ. The connected load consists of the contact impedance and the body impedance inbetween the electrodes. The electrode skin impedance is highly dependent on the type of electrodes used for the measurement. However, the contact impedance is usually in the range of a few kΩ for low frequencies and drops down to around 200$\Omega$ for frequencies above 100 kHz [1]. The resulting load from contact and body impedance is typically in the range of 200$\Omega$ or higher. However, the load might undercut the minimum operable load of the VS in specific applications. Future work should consider the expected load of the application in the design such that $R_s$ or $U_{out}$ do not lead to a compromising minimum operable load. This ensures that the permitted patient auxiliary current cannot be exceeded. Nevertheless, for an increased patient safety the auxiliary current can be monitored via $R_s$. If necessary the load current can be reduced by adjusting $U_{in}$.

The fully differential amplifier $OA2$ uses a resistor ratio of 1:3 with $R_3 = R_5 = 3R_4 = 3 \times R_6 = 1k\Omega$. This results in a gain of 1.5 and the overall $U_{in}$ is set to 0.963 V to satisfy a load current of 2.93 mA. Similar to the EHCS, the output impedance of the VS depends on the open-loop-gain $A_{OL}$ of the THS4631, which is 80 dB. According to eq. (13), the theoretical output impedance $Z_{out}$ is 204 mΩ, which is 104 mΩ higher than the output impedance of the THS4631. At 1 MHz, the $A_{OL}$ of the THS4631 decreases to 46 dB. Thus, the output impedance increases to 10.2 $\Omega$.

**Common mode rejection**

As both sources are symmetrical, the output signal consists of two alternating signals ($U_{L+}$ and $U_{L-}$), which are out of phase by 180°. Their difference creates the symmetrical signal

$$U_{load} = U_{L+} - U_{L-}.$$  \hspace{1cm} (15)

Eq. (16) describes the common mode signal $U_{cm}$ over the load $Z_{load}$, which is equally shared in both individual signals of the symmetrical signal and is their mean value.

$$U_{cm} = \frac{U_{L+} + U_{L-}}{2}.$$  \hspace{1cm} (16)

In an ideal case ($U_{L+} = -U_{L-}$), common mode components are equally shifted in the positive and negative parts of the signal and cancel each other. Thereby asymmetric effects of electrode double layers and artefacts are partly compensated [8, 9]. Indeed, inequalities of the positive and negative output create a difference between the generated currents. Due to the finite output impedance, a high common mode signal may occur [8]. This can be compensated with the help of an active feedback circuit, as presented in [9]. The compensation method by Pliquett et al. is only applicable to EHCS topologies. Here, we propose a modified circuit suitable for both types of sources (Figure 5).

According to eq. (16), $U_{L+}$ and $U_{L-}$ are picked up at their respective outputs. For the EHCS, this causes instabilities. Thus, the signals for the feedback loop are derived from the outputs of the differential amplifier of the EHCS. OP5 forms an inverting summing amplifier and is fed back to the offset pin of the input stage of the source. The negative feedback of the offset is an active compensation of the current common mode over $Z_{load}$. To avoid unwanted high-frequency oscillations of $I_{out}$, a first-order active low-pass filter is added to the summing stage through $C_{cm} = 2.2$ pF. This results in a cutoff frequency of 723 kHz.

$R_{ocm}$ creates a symmetric bypass from $U_{L+}$ and $U_{L-}$ to ground. $R_{ocm}$ should be at least one magnitude above the
Figure 5: The active offset compensation circuit sums up $U_{L+}$, $U_{L-}$ and an adjustable $U_{\text{offset}}$. The common-mode signal is fed back over a 1st order low pass filter to the $U_{\text{ocm}}$ input of the source.

maximum load to minimize the influence on the injected current. Thus, $R_{\text{ocm}}$ is set to 100 kΩ. $U_{\text{offset}}$ can be manually set to provide a frequency-independent offset voltage taking the inherent offset voltage of OP5 and slight miss-matches of $R_{\text{ocm}}$ into account.

For the evaluation, different offset compensation strategies can be realized by the switch $S_1$ and $U_{\text{offset}}$:

- Off: No compensation with $U_{\text{ocm}} = 0$ V ($S_1$ open, $U_{\text{offset}} = 0$ V).
- Static: $U_{\text{ocm}}$ is manually set ($S_1$ open, $U_{\text{offset}} \neq 0$ V).
- Active: $U_{\text{ocm}}$ in feedback control ($S_1$ closed, $U_{\text{offset}} = 0$ V).
- Active & Static: $U_{\text{ocm}}$ in feedback control & offset ($S_1$ closed, $U_{\text{offset}} \neq 0$ V).

The static offset compensation adds a constant offset voltage $U_{\text{offset}}$ to the feedback signal. $U_{\text{offset}}$ was trimmed manually, so that the common mode signal is minimized. We acknowledge that trimming values might be specific to the measurement frequency. However, for the sake of practicability we used 1 kHz to trim the constant offset voltage to a low common mode signal.

**Measurement setup**

The output current, output impedance, harmonic distortions and common mode rejection were measured with the setup depicted in Figure 6. All signals were sampled by a Tektronix MSO2024 along with a passive test probe (1 MΩ || 11.5 pF) having a total bandwidth of 100 MHz. Both sources drive floating loads, which requires differential voltage measurements. Thus, an instrumentation amplifier (AD8428, not depicted in Figure 6) measured the current through $R_{\text{shunt}} = 0.39$ Ω. By Ohm’s law the voltage drop of two different loads ($Z_1, Z_2$) is directly related to the output impedance. For the symmetrical current sources (Figure 6, left), the output impedance is

$$Z_{\text{out, CS}} = \frac{U_{d, open} - U_{d, load}}{U_{d, load}}. \quad (17)$$

where $Z_1$ simulates the voltage drop across a small and $Z_2$ across a larger load impedance. To not invoke the probeloading effect due to the sources high output impedance, $U_d$ of the current source was measured with an instrumentation amplifier (AD8429).

A similar principle applies for the VS (Figure 6, right), where the output voltage with open clamps $U_{d, open}$ and with an applied load $U_{d, Z_2}$ are measured. Therewith, the output impedance of the VS holds

$$Z_{\text{out, VS}} = Z_2 \cdot \frac{U_{d, open} - U_{d, load}}{U_{d, load}}. \quad (18)$$

Impedance measurements with a Keysight LCR meter E4980A yielded $Z_1 = 100.1$ Ω and $Z_2 = 1012.3$ Ω within the frequency range between 1 kHz and 1 MHz.

**Ethical approval**

The conducted research is not related to either human or animal use.

**Results**

The output impedance of both sources has been optimized with respect to the bandwidth requirements. Therefore, $C_v$ was swept from 0 pF to 4.4 pF and $C_v$ from 0 pF to 3.3 nF. All resistors were chosen with a tolerance of 0.1%. All simulations were performed with LTSpice XVII (Linear Technology. 17.0.0.11).

**Simulative output impedance of the EHCS**

First, we consider the output impedance of the current source without the influence of the capacitance $C_v$ and ideal resistor properties, which exactly fulfill the balancing condition given in eqs. (3), (4) and (5). The simulation
achieved an output impedance of 2.15 MΩ, which is 150 kΩ below the theoretical value. Second, as this is only valid for ideal resistor conditions, we evaluated the mean output impedance for Gaussian resistor value distributions with tolerances of 0.1% (Figure 7). Here, the simulated output impedance of the EHCS is 230 kΩ. For $C_c = 0$ pF, the output impedance has a low-pass character and drops to 100 kΩ at 1 MHz. The phase drops from 0° to −62.3°. A performance loss at higher frequencies is a typical phenomenon for current sources based on Howland topologies. It occurs due to inequalities at the inputs and outputs of the operational amplifier. The $f_{3dB}$ point lies slightly above 1 MHz.

Lastly, adding the capacitor $C_c$ parallel to $R_3$ in Figure 1 shifts the $f_{3dB}$ point to even higher frequencies. The additional capacitor is a common modification for unbalanced Howland Current Sources [22]. The output impedance has been simulated for 0 pF < $C_c$ < 4.4 pF between 1 kHz and 1 MHz. Figure 7 reveals an overshoot of $Z_{\text{out}}$ at 1 MHz with $C_c = 2.1$ pF. The range of $C_c$ is on an interval of 0.6 pF. For the evaluation in hardware, this value will be also influenced by the amplifiers input- and wire-capacitances on the PCB. Thus, a trim-capacitor will be used to optimize the behavior of the EHCS in the hardware realisation.

**Simulative output impedance of the voltage source**

Qureshi et al. proposed a capacitance (here: $C_v$) in parallel to $R_3$ to reduce oscillations above a few MHz [17, 23, 24]. Figure 8 shows the simulations $Z_{\text{out}}$ of the symmetrical VS with common values for $C_v$ between 0 pF and 3.3 nF. Without the additional capacitance $C_v$, the output impedance increases from 145 mΩ up to 14.5 Ω with 20 dB per decade over the whole frequency range. By introducing $C_v$, the output impedance turns flat for higher frequencies, due to an additional pole in the transfer function. For a capacitance of 3.3 nF, $Z_{\text{out}}$ drops down to 713.5 mΩ at 1 MHz and for $C_v > 2.2$ nF, the output impedance stays below 1.05 Ω. However, such high values of $C_v$ reduce the resulting current limiting impedance of $R_s$ from 1 kΩ to 67.5 Ω at 1 MHz. This results in higher maximum currents then 10 mA which should be prevented. This limitation could be overcome by selecting an appropriate $Z_t = R_s||C_v$, which can also fulfill the current limits due to the EN 60601-1 for lower frequencies.

**Output impedance of the EHCS hardware realization**

The output impedances were calculated from the measured voltage differences across the loads $Z_1$ and $Z_2$ by eq. (17). As the differences of the measured voltages are within a few milivolts, we ensured a higher accuracy by using a moving average over 64 values of $U_d$.

By the addition of $C_c$, a great bandwidth improvement of the symmetrical EHCS circuit was observed in the simulations. This capacitance was realised with a trim-capacitor and could be set between 1.5 pF and 5 pF. In order to demonstrate the influence of $C_c$ on the circuit, Figure 9 shows $Z_{\text{out},CS}$ for different positions of the trimming capacitor. Below 50 kHz, for all $C_c$ the output impedance has values between 100 kΩ and 200 kΩ. For $C_c \approx 2.6$ pF and $C_c \approx 3.2$ pF, $Z_{\text{out},CS}$ drops to 83 kΩ at 5 kHz. Between 50 kHz and 100 kHz, the output impedance of all capacitance values fans out, whereupon they all drop to values below 10 kΩ at 1 MHz. The largest differences in the output impedance for the different capacitors occur at 100 kHz. Here, the impedance increases up to 400 kΩ for $C_c = 1.5$ pF and $C_c = 2.6$ pF, but decreases for the measurements with the capacitor value in between ($C_c = 2.1$ pF) to 92 kΩ at 100 kHz. Overall, the output impedance at 100 kHz declines for the majority of capacitor values compared to its value at 50 kHz. From 200 kHz to 1 MHz, all measurement series generate a declining output impedance, with the same characteristics and no blatant aberrations. $C_c \approx 2.6$ pF was chosen for further considerations, as this value has
a significant higher value for $Z_{\text{out,CS,meas}}(f = 100 \text{ kHz})$ than most other capacities.

The theoretical maximum load of the proposed current source is $R_{\text{load,max,calc}} = 3.5 \text{ kΩ}$. To determine the maximum operable load $R_{\text{load,max,meas}}$ at 1 MHz, four resistive loads ($R_{\text{load}} = 2.47 \text{ kΩ}, 3 \text{ kΩ}, 3.47 \text{ kΩ}$ and $3.74 \text{ kΩ}$) were selected for current measurements. The magnitude and phase of the output current $I_{\text{out,crit,meas}}(R_{\text{load}} = 2.47 \text{ kΩ})$ in Figure 10 are similar to the values of the previous measurements below 2 kΩ. For higher loads, the amplitude and phase decrease further. Thus, it can be concluded that the maximum operable load of the symmetric EHCS is about 2.5 kΩ in the frequency band of the β-dispersion.

Output impedance of the VS hardware realization

In simulations, the bandwidth of the VS’s output impedance increased for higher values of $C_v$. However, the VS showed instabilities for $C_v > 2.2 \text{ nF}$. This behavior might be based on imbalances of the capacitors $C_v$ of the two VS (Figure 4). Higher values of $C_v$ have larger absolute differences due to capacitor tolerance’s and influence different cut-off frequencies and phases of the two VSs. To avoid such instabilities, we chose $C_v = 82 \text{ pF}$ for all further measurements. Besides, with this value, the VS still fulfills the current limits of EN 60601-1 for higher frequencies.

The measured output impedance of the balanced VS was calculated using eq. (18) and is shown in Figure 11. Up to 60 kHz, $Z_{\text{out,VS,meas}}$ has a flat profile with a mean value of $Z_{\text{out,VS,meas}} < 1 \text{ Ω}$. From this point up to a frequency of 1 MHz, the output impedance fluctuates much more and has a maximum value of 15.5 Ω at 700 kHz. Overall, the balanced VS has measured impedance values in the range of the simulated values.

In order to measure the critical lower load range for the VS, the output voltages were measured across 270 Ω, 470 Ω and 740 Ω. The measurement results are shown in Figure 12. The theoretical minimum value $R_{\text{load,min}}$ is 208.8 Ω (eq. (14)). Magnitude and phase of the output voltage are very similar for all values of $R_{\text{load}}$. However, the waveform of the signal slightly distorts for $R_{\text{load}} = 470 \text{ Ω}$ at a frequency of 1 MHz, resulting in a steeper drop of the phase. For $R_{\text{load}} = 270 \text{ Ω}$, the distortion already appears at 500 kHz. The actual critical load value of the VS is between 470 Ω and 740 Ω and more than twice the theoretical value of 208.8 Ω.
Simulative common mode rejection

The common mode rejection (CMR) circuit from Figure 5 enables us to compare the resulting common mode using no compensation, static offset correction and active compensation with the proposed feedback loop. The simulated CMR of all compensation strategies, presented in Table 1, has a nearly constant value between 1 kHz and 1 MHz, due to the low variance $\sigma$.

Table 1: The mean CMR without, with static and active compensation methods. The variances $\sigma$ of the VS ($C_v = 82$ pF) and current source ($C_c = 2.6$ pF) are given for the frequency range 1 kHz to 1 MHz.

| Compensation        | $U_{cm, VS} \text{ [dB]}$ mean | $U_{cm, VS} \text{ [dB]}$ $\sigma$ | $U_{cm, EHCS} \text{ [dB]}$ mean | $U_{cm, EHCS} \text{ [dB]}$ $\sigma$ |
|---------------------|--------------------------------|---------------------------------|---------------------------------|---------------------------------|
| Off                 | -29.76                        | 0                               | -30.71                          | 0                               |
| Static              | -79.04                        | 0.004                           | -130.08                         | 9.03                            |
| Active              | -39.43                        | 0                               | -38.65                          | 0.04                            |
| Active & Static     | -128.93                       | 0.5                             | -139.76                         | 10.91                           |

The passive method with an $U_{off} = 0$ V produces a common mode voltage of approx. -30 dB or 29.15 mV with an rms value of 648 $\mu$V for both sources. The static compensation technique achieved a CMR of approx. -80 dB for the VS and approx. -130 dB for the EHCS, respectively. In contrast, the active compensation reduces the common mode signal of approx. -40 dB for both VS and EHCS. To further improve the CMR we investigated the combined application of active and static compensation. The combination reduces the common mode voltage up to $-129$ dB for the VS and up to $-139$ dB for the current source, which is equivalent to 0.35 nA and 0.1 nA offset current, respectively.

Common mode reduction of the hardware realisation

The common mode rejection was measured at 1 k$\Omega$ load impedance for the three presented compensation methods. Without any compensation method, the EHCS achieved a CMR of almost -30 dB, shown in Figure 13 (blue). The CMR of the EHCS reduces from -72 dB to -43 dB with higher frequencies for the static compensation method. In contrast, the active compensation method has a fairly low frequency-dependent behavior and achieved a mean common mode reduction of -46.5 dB. The combined compensation has a mean CMR of -57.7 dB with a slightly lower reduction value for higher frequencies.

The VS showed a similar behavior as the EHCS without any compensation. The static common mode reduction of the VS also has a frequency-dependency (Figure 14). In contrast to the EHCS, the highest reduction of -80 dB is achieved for 1 kHz and 1 MHz and increases to -56.5 dB at 10 kHz. The active compensation method slightly improves for higher frequencies from -65.2 dB to -72.8 dB. Both compensation methods combined achieved a mean CMR of -71.8 dB, which is slightly below the active compensation for lower frequencies and above for higher frequencies.

Signal-to-noise ratio and harmonic distortion of the hardware realisation

The signal quality is also dependent on the spectral components aside from the measurement frequency. The relation between the signal energy at a specific measurement frequency and the energy of the remaining frequency bandwidth is called signal-to-noise ratio (SNR). The SNR of the source limits the dynamic range of the resulting bioimpedance measurements. Thus, we evaluated the resulting SNR of both sources at 10 kHz and 100 kHz. The current was measured through a 1 k$\Omega$ load with a NI-USB 6259 DAQ board (National Instruments, Austin, USA). The EHCS and the VS achieved an SNR of approx. 47 dB and 46 dB, respectively (Table 2).

Besides, the sinusoidal shape of the current is another aspect of signal quality. Due to the finite slew rate of operational amplifier, the injected sinusoidal current is biased by higher harmonics of the base frequency, resulting in a deformation of the signal shape. This behavior is
Table 2: SNR and THD of the hardware realization.

| f [kHz] | EHCS    | VS      |
|---------|---------|---------|
| SNR [dB] |         |         |
| 10      | 47.53 dB | 46.22 dB |
| 100     | 47.14 dB | 46.25 dB |
| THD [dBc] |        |         |
| 10      | -42.31  | -44.15  |
| 100     | -35.17  | -35.74  |

quantified by the total harmonic distortion (THD) and the harmonic distortion (HD):  

\[ THD = \sqrt{\sum_{i=2}^{n} \frac{V_i^2}{V_1}} \]  

(19)

\[ HD_i = \frac{V_i}{V_1} \]  

(20)

where \( V_1 \) is the base frequency and \( V_i \) with \( i \geq 2 \) are the higher harmonics. THD and HD is a relative measure of the harmonic's signal amplitude in relation to the injected signal amplitude. The amplitude difference is given in dBc (dB carrier).

Measurements showed that the VS had a 1.85 dBc lower THD compared to the EHCS at 10 kHz. This effect diminished at 100 kHz, where both sources achieved a THD of less than -35 dBc.

A more detailed investigation of the harmonic distortion has been measured with the Tektronix MSO2024. The second harmonic distortions (HD_2, solid lines) of the EHCS and the VS are constant at -68 dBc up to 100 kHz, which is equal to the noise floor of the oscilloscope (Figure 15). Above 100 kHz, HD_2 of the VS rises constantly with 20 dBc per decade. For the EHCS, HD_2 rises at 300 kHz with 60 dBc per decade. Both sources reach approx. -30 dBc at 1 MHz. Typical for differential signals, the third harmonics of both sources (HD_3, dashed lines) are above the second harmonics. On average, the EHCS had a HD_3 of -50 dBc and the VS has -53 dBc. The third order harmonics shows no frequency-dependent behavior in contrast to the second order harmonics. The values of the fourth as well as the fifth harmonic distortion are below the noise floor of the used oscilloscope (-64 dB).

Discussion and conclusion

Output impedance

The parallel capacitors \( C_V \) and \( C_C \) have been introduced to improve the bandwidth of the sources. In simulations, the added capacitor could achieve a flat spectrum of \( Z_{out} \) for both sources. Moreover, the simulated output impedance of the EHCS was 230 k\( \Omega \) for resistor tolerances of 0.1%. The measured output impedance achieved similar results in the range of 100 - 200 k\( \Omega \) for frequencies below 100 kHz. However, \( C_C \) increased \( Z_{out} \) up to 400 k\( \Omega \) at 100 kHz. For frequencies above 200 kHz, the output impedance dropped down to 10 k\( \Omega \), independent of \( C_C \). Overall, the different values of \( C_C \) have a lower influence on the measured \( Z_{out} \) compared to the simulated \( Z_{out} \). While the output impedance of the current source obtained from Sirtoli et al. featured a small difference between simulated and measured values, our implementation of the VCCS has in this regard a greater performance loss. The difference to the simulations might be caused by typical problems of the EHCS such as tolerances of component values and stray capacitance influencing the balancing conditions. Also, a too low open loop gain could be the reason for the reduced output impedance at higher frequencies. In future studies, the usage of resistors with lower tolerances or laser trimmable resistors could achieve higher output impedances, as the theoretical value reaches 2.27 M\( \Omega \). An improved version with a higher output impedance might reveal if \( C_C \) has the ability to increase the output impedance of the VCCS in real implementations.

For the VS, \( Z_{out} \) matched the expected values of the simulations for frequencies above 100 kHz. In the lower frequency range, the measured output impedance is slightly higher, but still mostly below the pursued 1 \( \Omega \) output impedance. The measured output impedances were evaluated for \( C_V = 82 \) pF. In future work, also higher values for \( C_V \) should be considered to further reduce the output impedance in the frequency range above 100 kHz.

The acceptable load of the EHCS has good characteristics in the lower load range. The difficulties of the EHCS occur at loads above 2.47 k\( \Omega \) at higher frequencies (above 500 kHz). Such large biom impedances do not usually occur at these frequencies, hence this impairment is negligible. The voltage source exhibits phase shifts and distortions above 470 \( \Omega \). Yet, amplitude attenuation is not affected.

Common mode reduction

The common mode signals of voltage and current sources are influenced by offset voltages of the operational amplifiers and miss-matches of resistors. Different common mode reduction strategies were evaluated. We note that the measured common mode of the EHCS might be influenced by the probe loading effect especially for the high impedance positive output. The usage of an instrumentation amplifier, similar to the measurement setup of the output impedance, is not applicable due to
the common mode measurement principle. However, the probe loading effect applies to all compensation strategies and should not compromise the comparability. We could significantly reduce the common mode of both sources in simulations as well as in experimental evaluation by the application of a static offset compensation. Additionally, an active feedback compensation could also reduce the common mode signal, but to a minor extent. However, the common mode signal increases for higher frequencies. This behavior can be reduced by the usage of active compensation and results in a more constant common mode reduction over the whole frequency range. The combination of both methods achieved a mean common mode reduction by -71.8 dB for the voltage source, which is similar to the passive compensation but shows significantly less frequency-dependency. For the EHCS, the combined usage of both methods is equivalent to the purely static compensation with -57.7 dB, but has a more constant frequency behavior. According to the EN 60601-1, a maximum current of 10 μA is tolerable. The measured absolute common mode using the combined compensation strategy achieved a mean value 1.7 μA and 0.06 μA for the EHCS and VS, respectively. Thus, the proposed compensation technique is an adequate principle to fulfill the strict patient safety requirements of bioimpedance measurements. The proposed offset compensation strategy is easy to implement and achieved a significant common mode reduction, which makes them suitable for many low-cost bioimpedance applications.

**Signal-to-noise ratio and harmonic distortion**

For the signals at 10 kHz and 100 kHz, the SNRs are constant and almost identical with approximately 47 dB. The THD, on the other hand, increases from -42.3 dBc by 7.1 dBc for the EHCS, and from -44.2 dBc by 8.4 dBc for the VS. Both sources show the same frequency-dependent behavior, which might be caused by the THD of the differential amplifier THS4151.

The usage of symmetric signals suppresses HDs. The measured HD2 were below the noise floor of -68 dBc for both sources up to 100 kHz. For higher frequencies, HD2 increased, which might be caused by higher frequency poles of the sources.

**Conflict of interest**

Authors state no conflict of interest.

**References**

1. Nguyen DT, Kosobrodov R, Barry MA, Chik W, Jin C, Oh TI, Thiagalingam A, and McEwan A. Electrode-skin contact impedance: In vivo measurements on an ovine model. Journal of Physics: Conference Series 2013; 434. DOI: 10.1088/1742-6596/434/1/012023

2. Ackmann JJ. Complex bioelectric impedance measurement system for the frequency range from 5 Hz to 1 MHz. Annals of biomedical engineering 1993; 21:135–46. DOI: 10.1007/BF02367609

3. Bouchala D. Investigation of Current Excitation for Personal Health and Biological Tissues Monitoring. Vol. Volume 1. Scientific Reports on Measurement and Sensor Technology. Chemnitz: Universitätssverlag Chemnitz, 2016. Available from: https://nbn-resolving. urn:nbn:de:bsz:ch1-qucosa-204801

4. Naranjo-Hernández D, Reina-Tosina J, and Min M. Fundamentals, recent advances, and future challenges in bioimpedance devices for healthcare applications. Journal of Sensors 2019; 2019:42. DOI: 10.1155 / 2019 / 9210258

5. Cole KS and Cole RH. Dispersion and Absorption in Dielectrics I. Alternating Current Characteristics. The Journal of Chemical Physics 1941; 9:341–51. DOI: 10.1063/1.1750906. arXiv: arXiv:1011.1669v3

6. Schwan HP. Electrical Properties of Tissue and Cell Suspensions. Advances in Biological and Medical Physics. Elsevier Inc., 1957 :147–209. DOI: 10.1016/B978-1-4832-3111-2.50008-0

7. Rosell J, Colominas J, Riu P, Palas-Areny R, and Webster JG. Skin impedance from 1 Hz to 1 MHz. IEEE transactions on bio-medical engineering 1988; 35:649–51. DOI: 10.1109/10.4599

8. Sirtoli VG, Morcelles KF, and Vincente VC. Design of current sources for load common mode optimization. Journal of Electrical Biomeadance 2018; 9:59–71. DOI: 10.2478/joeb-2018-0011

9. Pliquet U, Schönfeldt M, Barthel A, Frense D, Nacke T, and Beckmann D. Front end with offset-free symmetrical current source optimized for time domain impedance spectroscopy. Physiological Measurement 2011; 32:927–44. DOI: 10.1088/0967-3334/32/7/S15

10. Zhang F, Teng Z, Zhong H, Yang Y, Li J, and Sang J. Wideband mirrored current source design based on differential difference amplifier for electrical bioimpedance spectroscopy. Biomedical Physics & Engineering Express 2018; 4:025032. DOI: 10.1088/2057-1976/aaa9cd

11. Rao AJ, Murphy EK, Shahghasemi M, and Odame KM. Current-conveyor-based wide-band current driver for electrical impedance tomography. Physiological Measurement 2019; 40:034005. DOI: 10.1088/1361-6579/ab0c3c

12. Hong H, Rahal M, Demosthenous A, and Bayford RH. Comparison of a new integrated current source with the modified Howland circuit for EIT applications. Physiological Measurement 2009; 30:999. DOI: 10.1088/0967-3334/30/10/001

13. Mohamadou Y, Oh TI, Wh H, Sohal H, Farooq A, Woo EJ, and McEwan AL. Performance evaluation of wideband bio-impedance spectroscopy using constant voltage source and constant current source. Measurement Science and Technology 2012; 23:105703. DOI: 10.1088/0957-0233/23/10/105703

14. Xu Z, Yao J, Wang Z, Liu Y, Wang H, Chen B, and Wu H. Development of a Portable Electrical Impedance Tomography System for Biomedical Applications. IEEE Sensors Journal 2018; 18:8117–24. DOI: 10.1109 / JSEN.2018.2864539
15. Nouri H, Bouchaala D, Gargouri R, and Kanoun O. Stability Analysis for Howland Current Source for Bioimpedance Measurement. 2021 IEEE International Symposium on Medical Measurements and Applications (MeMeA). 2021 :1–6. DOI: 10.1109/MeMeA52024.2021.9478601

16. Saulnier GJ, Abdelwahab A, and Shishvan OR. DSP-based current source for electrical impedance tomography. Physiological Measurement 2020; 41:064002. DOI: 10.1088/1361-6579/ab8f74

17. Qureshi TR, Chatwin CR, Zhou Z, Li N, and Wang W. Investigation of voltage source design’s for Electrical Impedance Mammography (EIM) Systems. Annual International Conference of the IEEE Engineering in Medicine and Biology Society. 2012 :1582–5. DOI: 10.1109/EMBC.2012.6346246

18. Brown BH and Seagar AD. The Sheffield data collection system. Clinical physics and Physiological Measurement 1987; 8 Suppl A:91–7. DOI: 10.1088/0143-0815/8/4a/012

19. Chen CY, Lu YY, Huang WL, and Cheng KS. The simulation of current generator design for multi-frequency electrical impedance tomograph. 2006 International Conference of the IEEE Engineering in Medicine and Biology Society. 2006 :6072–5. DOI: 10.1109/IEMBS.2006.259576

20. Qureshi TR, Chatwin CR, Huber N, Zarafshani A, Tunstall B, and Wang W. Comparison of Howland and General Impedance Converter (GIC) circuit based current sources for bio-impedance measurements. Journal of Physics: Conference Series 2010; 224:012167. DOI: 10.1088/1742-6596/224/1/012167

21. Franco S. Design with Operational Amplifiers and Analog Integrated Circuits. 3rd ed. USA: McGraw-Hill, Inc., 2001

22. Rafiei Naeini M and McCann H. Low-noise current excitation sub-system for medical EIT. Physiological Measurement 2008; 29:S173. DOI: 10.1088/0967-3334/29/6/S15

23. Qureshi T, Mehboob B, Chatwin CR, and Wang W. Optimal device for a low output impedance voltage source for Electrical Impedance Tomography (EIT) systems. 2012 IEEE International Conference on Electro/Information Technology. IEEE, 2012 :1–6. DOI: 10.1109/EIT.2012.6220742

24. Qureshi TR, Chatwin C, and Wang W. Bio-impedance Excitation System: A Comparison of Voltage Source and Current Source Designs. APCBEE Procedia 2013; 7. The 3rd International Conference on Biomedical Engineering and Technology - ICBET 2013:42–7. DOI: 10.1016/j.apcbee.2013.08.010