A simple and reproducible capacitive electrode

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Abstract

Capacitive Electrodes (CE) allow the acquisition of biopotentials through a dielectric layer, without the use of electrolytes, just by placing them on skin or clothing, but demands front-ends with ultra-high input impedances. This must be achieved while providing a path for bias currents, calling for ultra-high value resistors and special components and construction techniques. A simple CE that uses bootstrap techniques to avoid ultra-high value components and special materials is proposed. When electrodes are placed on the skin: that is, with coupling capacitances $C_s$ of around 100 pF, they present a noise level of 3.3 $V_{\text{rms}}$ in a 0.5–100 Hz bandwidth, which is appropriate for electrocardiography (ECC) measurements. Construction details of the CE and the complete circuit, including a fast recovery feature, are presented.

Keywords: Insulating electrodes, Active electrodes, Non-contact measurements

1. Introduction

Capacitive Electrodes (CE) do not require the use of electrolytes. They acquire biopotentials through a dielectric layer by just placing them on the skin [1,2] clothing [3,4], or without any physical contact with the patient [5]. They avoid skin irritation, are simple to install, and appropriate for long-term patient monitoring.

The general scheme of a CE measurement set-up is shown in Fig. 1(a) and its equivalent circuit in Fig. 1(b) [6]. This is reduced to a simple AC-coupled amplifier [7], but for ‘coin-size’ CEs, coupling capacitance $C_s$ can be as low as a few tens of pF (10–30 pF), when biopotentials are picked up through clothing, or hundreds of pF (100–300 pF), when CEs are placed on the skin with a dielectric film [4]. In order to achieve the very low cut-off frequencies that biomedical signals require for an ECG [8], these small $C_s$ values demand bias resistors $R_B$ as high as 0.1–1 TΩ. Electronic requirements relax for large $C_s$ values, when very thin dielectric layers [2,7] or large-area CEs are used [9].

Since CEs must work with ultra-high impedances, they are vulnerable to electric-field interference and sensitive to circuit leakages, requiring high-quality Printed Circuit Board (PCB) substrates, and careful guarding and shielding techniques to keep unavoidable leakages and couplings under control [3,4]. To deal with this, a practical CE circuit includes a guard-driver, and a neutralization circuit to reduce the effects of PCB and amplifier input capacitances. Details of how these sub-circuits work can be found in [6].

Capacitive electrodes present noise levels greater than their 'wet' counterparts. The noise Power Spectral Density (PSD) $e_0$ for frequencies above the cut-off frequency $f_N$, is given approximately by [6]:

$$e_0 \approx e_{in}^2 (f_N/f)^3 + e_{in}^2 R_s^2 (f_N/f)^2 + \alpha e_{in}^2,$$

where $e_{in}$, $e_{in}$ denote the Operational Amplifier’s (OA) current and voltage noises respectively, $e_{in}$ is the thermal noise of $R_s$, factor $\alpha$ represents effects of neutralization and guarding circuits that amplify $e_{in}$, and $f_N$ denotes the cut-off frequency:

$$f_N = (2\pi R_s C_s)^{-1}.$$  

(2)

Eq. (1) shows that reducing $f_N$ decreases the electrode noise PSD. Then, the noise cut-off frequency $f_N$ must be set below the signal pass-band (as far below as possible) in order to limit the effect of low-frequency noise [6].

Expression (1) does not include noise sources outside the CE itself, such as those produced by clothing or skin layers [4]. Replacing $e_{in}^2$ by the Nyquist expression ($e_{in}^2 = 4kT R_s$) and $f_N$ by (2), results in:

$$e_0 \approx \frac{kT}{(2\pi C_s f)^2} \frac{1}{R_s} + \frac{P_{in}^2}{(2\pi C_s f)^2} + \alpha e_{in}^2.$$  

(3)

As can be observed in (3), to reduce $e_0$, a low noise OA should be used, and — less obviously — the value of $R_s$ should be the highest possible [3]. Resistors $R_s$ of the order of TΩ are desirable, but they are not easy to obtain and handle. Some techniques to achieve ultra-high value resistors have been proposed, such as using reverse polarized diodes [2,10] and ‘gimmick’ resistors implemented from insulated cables’ leakages [6]. Moreover, these high-value $R_s$ values impose large time constants to discharge $C_s$ when
times greater than that of a ‘real resistor’ of the same value. Replacing in (1) \( e_{RB} \) by the \( e_{RBQ} \) expression given by (5), and
\[
\frac{f_n}{(2\pi R_{BEQ}C_S)} \text{, Eq. (1) becomes:}
\]
\[
e_n^2 \approx \frac{kT}{(\pi C_S)^2 R_1} + \frac{i_0^2}{(2\pi \nu f)^2} + \alpha e_n^2;
\]
which corresponds exactly to (1) with \( R_0 = R_1 \). Hence, the bootstrapping multiplies \( R_1 \) allowing to achieve a very low cut-off frequency \( f_n \) and proper transient responses, but it also amplifies the noise of \( R_1 \). As a result, the CE noise is the same as using \( R_1 \) in place of \( R_0 \). However, an \( R_1 \) of a few \( \Omega \) (a high but accessible value) is high enough to acquire good-quality ECG signals, even picking them up through cotton clothes. The circuit herein proposed implements \( R_{BEQ} = 100 \, \Omega \) by \( R_1 = 10 \, \Omega \), and a bootstrap ratio \( R_2/R_3 = 10 \). This \( R_0 \) value is enough to achieve time constants of a few seconds and allows building the CE with standard FR4 PCB material, which has a superficial resistivity (SR) of around 50 \( \Omega \)/square, instead of using more expensive substrates materials such as Teflon\textsuperscript{TM}, that present SR values of 1 \( \Omega \)/square and more.

2.1. Description of the proposed circuit

The complete circuit shown in Fig. 3(a) includes a guard driven by the output of OA1, and an input capacitance neutralization circuit implemented by OA2 through capacitor \( C_4 \), according to [6]. The corresponding PCB design is shown in Fig. 3(b), where it can be observed that the neutralization capacitance \( C_4 \) is implemented by a PCB area. A dual low-bias current operational amplifier OPA2320 (by Texas Instruments\textsuperscript{TM}) was used. The non-inverting input of OA1 – the most vulnerable node of the circuit – is not soldered to the PCB, but bent upwards, and capacitor \( C_4 \) and \( R_1 \) are soldered directly to it [13]. No solder-mask was used, in order to reduce superficial leakages.

The proposed CE itself, without additional elements, provides a ‘fast recovery’ mechanism to restore the baseline when high-amplitude artefacts saturate the amplifier. The circuit time constant in normal operation is \( R_{BEQ} C_4 \), but short-circuiting the output of OA1 (it must be output-protected), reduces it to \( (R_1 + R_3)C_4 \), thus providing a way to discharge \( C_4 \). The recovery of the baseline is not as fast as using the circuit proposed in [14], but is much simpler to implement.

Using resistors \( R_1 \) of around 10 \( \Omega \) and a low current noise OA as the OPA320, the noise PSD \( e_0^2 \) is dominated by the first term in (6), decreasing with frequency \( f \) according to:
\[
e_0^2 \approx \frac{kT}{(\pi C_S)^2 R_1};
\]
Fig. 3. (a) Complete circuit of the proposed CE and (b) printed circuit board and photo of the built prototype. Note that the proposed CE requires a reduced number of parts and includes a guard ring, plus a shield (ground) ring.

Fig. 4. Noise Power Spectral Density (PSD) of the proposed capacitive electrode for \( C_S = 10 \) pF and \( C_S = 100 \) pF. The expected noise due to resistor \( R_N \), given by (8) for each \( C_S \) value, is indicated in dashed line.

The total noise that is obtained by integrating (7) in a bandwidth from \( f_1 \) to \( f_2 \); for \( f_2 > f_1 \) becomes independent of \( f_2 \) [6] and is given by:

\[
E_{\text{noise}}^2 \approx \frac{kT}{(\pi C_S)^2 R_1 f_1}.
\]  \( (8) \)

Considering \( R_1 = 10 \) k\( \Omega \), \( T = 300 \) K, \( C_S = 100 \) pF and \( f_1 = 0.05 \) Hz, the total noise \( E_{\text{noise}} \) results in 9 \( \mu \)V\(_{\text{RMS}} \), which is a reasonable value for ECG signals. If \( f_1 = 0.5 \) Hz is considered, the noise is limited to 2.9 \( \mu \)V\(_{\text{RMS}} \).

3. Experimental results

The circuit of Fig. 3(a) was built with a diameter of 25 mm and shielded. At first, a sinusoidal signal of \( \pm 100 \) mV, 1 kHz was applied through a 10 pF capacitor working as \( C_S \), and neutralization was adjusted by the trimpot \( T_1 \) to a unity gain. Then, the noise PSD was measured for \( C_S = 10 \) pF and \( C_S = 100 \) pF, resulting in the curves presented in Fig. 4. The CE total noise in the bandwidth 0.5–100 Hz for \( C_S = 100 \) pF, obtained by integrating the respective PSD, is 3.3 \( \mu \)V\(_{\text{RMS}} \). This value is a little higher than the 2.9 \( \mu \)V\(_{\text{RMS}} \) predicted by (8), because of additional noise sources not considered in this equation, such as OA current noise and other effects that amplify the voltage noise of the OA [6].

Finally, the CE was insulated with a 50 \( \mu \)m auto-adhesive polypropylene film, and real ECG signals were acquired from a volunteer. Records were performed simultaneously by two CEs placed on the subject’s chest, using a pair of standard disposable wet electrodes (3M™ 2223). Signals were acquired using an 8-channel biopotential acquisition system, based on the IC ADS1298 of Texas Instruments™. Monopolar channels were used, in order to verify that each electrode worked properly, and bipolar (differential) signals shown in Fig. 5 were obtained digitally by subtraction.
4. Conclusions

By using bootstrapping it is possible to implement CE avoiding the use of ultra-high bias resistors, simulating them with moderate value ones, but the CE noise is the same as when using the circuit’s higher value resistor as a bias path. As can be observed in Fig. 5, the proposed capacitorate electrode allows ECG signals to be acquired with a good signal-to-noise ratio, even picking them up through clothing.

The proposed CE does not require the special substrates and fabrication techniques needed in [3]. It does not demand ultra-high value resistors as the one presented in [6], and provides a fast recovery feature with a simpler circuit than those in [14] or [15]. The noise level of the CE is slightly higher than that in its previous version [6], but it is easier to build and replicate. Complete circuits and construction details were provided, thus placing this work within a reproducible research framework.

Conflict of interest

No conflict of interest.

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Appendix. Noise analysis of the bootstrap circuit

The simulated resistor $R_{EQ}$ presents a noise voltage greater than that of a real resistor of the same value. To estimate this, the noise of each resistor and OA composing the circuit in Fig. 2(b) must be considered. The voltage source $e_i$ represents the OA voltage noise and $e_{R1}$, $e_{R2}$, $e_{R3}$ are the noise of resistors $R_1$, $R_2$ and $R_3$, respectively. The amplifier current noise $i_n$ is not included, because the effect it produces on $R_{EQ}$ is the same as that it produces on a real $R_B$. This is already considered in the CE noise analysis that yields (1). Solving the circuit of Fig. 2, the overall noise at the output results:

$$
V_n = e_i^2 = e_i^2 + 2e_i^2 + e_i^2 + e_i^2 + e_i^2 = (1 + R_2/R_3)^2;
$$

and replacing resistors’ noise PSDs by the Johnson–Nyquist formula ($e_i = 4kT$):

$$
V_n = e_i^2 = e_i^2 + 4kT R_1(1 + R_2/R_3)^2
+ 4kT R_2 + 4kT R_3(1 + R_2/R_3)^2.
$$

Given that $R_1 > R_2$, $R_3 > R_2/R_3$ and $e_i << e_{R1}$: Eq. (10) can be approximated by:

$$
V_{R2} = 4kT R_1 (R_2/R_3)^2;
$$

which can be written as:

$$
V_{R2} = 4kT R_{EQ} (R_2/R_3)^2.
$$

References