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Technical note

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 μV_{RMS} in a 0.5–100 Hz bandwidth, which is appropriate for electrocardiography (ECG) measurements. Construction details of the CE and the complete circuit, including a fast recovery feature, are presented.
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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_O for frequencies above the cut-off frequency f_N, is given approximately by [6]:

$$e_O^2 \approx e_{R_B}^2 (f_N/f)^2 + i_n^2 {R_B}^2 (f_N/f)^2 + \alpha e_n^2, \tag{1}$$

where i_n, e_n denote the Operational Amplifier's (OA) current and voltage noises respectively, e_{RB} is the thermal noise of R_B, factor α represents effects of neutralization and guarding circuits that amplify e_n, and f_N denotes the cut-off frequency:

$$f_N = (2\pi R_B C_S)^{-1}. \tag{2}$$

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 $e_{R_B}^2$ by the Nyquist expression ($e_{R_B}^2 = 4kTR_B$) and f_N by (2), results in:

$$e_O^2 \approx \frac{kT}{(\pi C_S f)^2}\frac{1}{R_B} + \frac{i_n^2}{(2\pi C_S f)^2} + \alpha e_n^2. \tag{3}$$

As can be observed in (3), to reduce e_O a low noise OA should be used, and — less obviously — the value of R_B should be the highest possible [3]. Resistors R_B 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_B values impose large time constants to discharge C_S when

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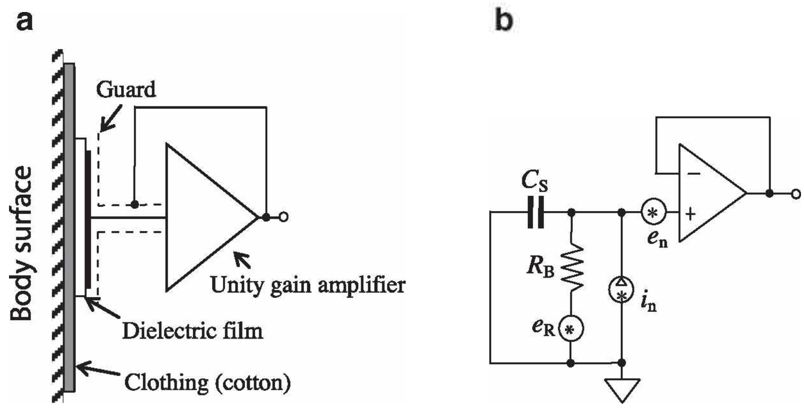

Fig. 1. (a) General scheme of a capacitive electrode and (c) its equivalent circuit including intrinsic noise sources.

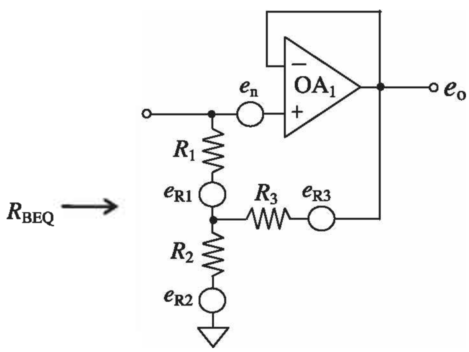

Fig. 2. Bootstrap circuit to high-input impedances by 'multiplying' resistor R_1 value. This figure also includes circuit's noise sources related to resistors and to the operational amplifier.

a high-amplitude artefact saturates the front-end, thus requiring additional circuits elements to recover the baseline in reasonable times.

It is possible to 'simulate' ultra-high resistance values by moderate value resistors, using bootstrapping techniques. A typical circuit, shown in Fig. 2, works as an equivalent resistor R_{BEQ} given by:

$$R_{BEQ} = R_1 + R_2 + R_1R_2/R_3. \tag{4}$$

In general, the last term in (4) is dominant and $R_{BEQ} \approx R_1R_2/R_3$: the bootstrap circuit increases R_1 value by R_2/R_3 times. However, this technique increases resistor R_1 value at the expense of increasing OA's noise and offset voltage effects [11]. Voltage offset amplification can be avoided by replacing R_3 with a capacitor [3], but it introduces a singularity in the transfer function. Thus, the input impedance becomes inductive [12], and it is difficult to fulfil the strict transient response that biomedical standards demand [8]. Nowadays, OAs suitable for CEs with low offset input voltages (i.e. the OPA320 of Texas Instruments) are available, and the fully resistive circuit of Fig. 2 is feasible. Herein, a very simple CE based on this circuit, is proposed.

2. Material and methods

Bootstrapping allows increasing resistor R_1 to R_{BEQ}, but R_{BEQ} presents a noise voltage e_{RBEQ} greater than that of a real resistor of the same value. As is depicted in the Appendix, the noise of the equivalent resistor R_{BEQ} is:

$$e^2_{R_{BEQ}} \approx 4kTR_{BEQ}(R_2/R_3). \tag{5}$$

The bootstrap circuit in Fig. 2 increases R_1 by the factor (R_2/R_3), thus reducing f_N to $(2\pi R_{BEQ}C_S)^{-1}$, but the noise of R_{BEQ} is $\sqrt{R_2/R_3}$ times greater than that of a 'real resistor' of the same value. Replacing in (1) e_{RB} by the e_{RBEQ} expression given by (5), and $f_N = (2\pi R_{BEQ}C_S)^{-1}$, Eq. (1) becomes:

$$e^2_O \approx \frac{kT}{(\pi C_S f)^2}\frac{1}{R_1} + \frac{i^2_n}{(2\pi C_S f)^2} + \alpha e^2_n; \tag{6}$$

which corresponds exactly to (1) with $R_B = R_1$. 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_B. However, an R_1 of a few GΩ (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 GΩ by R_1=10 GΩ, and a bootstrap ratio R_2/R_3=10. This R_B 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 GΩ/square, instead of using more expensive substrates materials such as Teflon™, that present SR values of 1 TΩ/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 OA_1, and an input capacitance neutralization circuit implemented by OA_2 through capacitor C_C, according to [6]. The corresponding PCB design is shown in Fig. 3(b), where it can be observed that the neutralization capacitance C_C is implemented by a PCB area. A dual low-bias current operational amplifier OPA2320 (by Texas Instruments™) was used. The non-inverting input of OA_1 — the most vulnerable node of the circuit — is not soldered to the PCB, but bent upwards, and capacitor C_A 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 $R_{BEQ}C_S$, but short-circuiting the output of OA_1 (it must be output-protected), reduces it to $(R_1+R_3)C_S$, thus providing a way to discharge C_S. 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 GΩ and a low current noise OA as the OPA320, the noise PSD e_O is dominated by the first term in (6), decreasing with frequency f according to:

$$e^2_O \approx \frac{kT}{(\pi C_S f)^2}\frac{1}{R_1}; \tag{7}$$

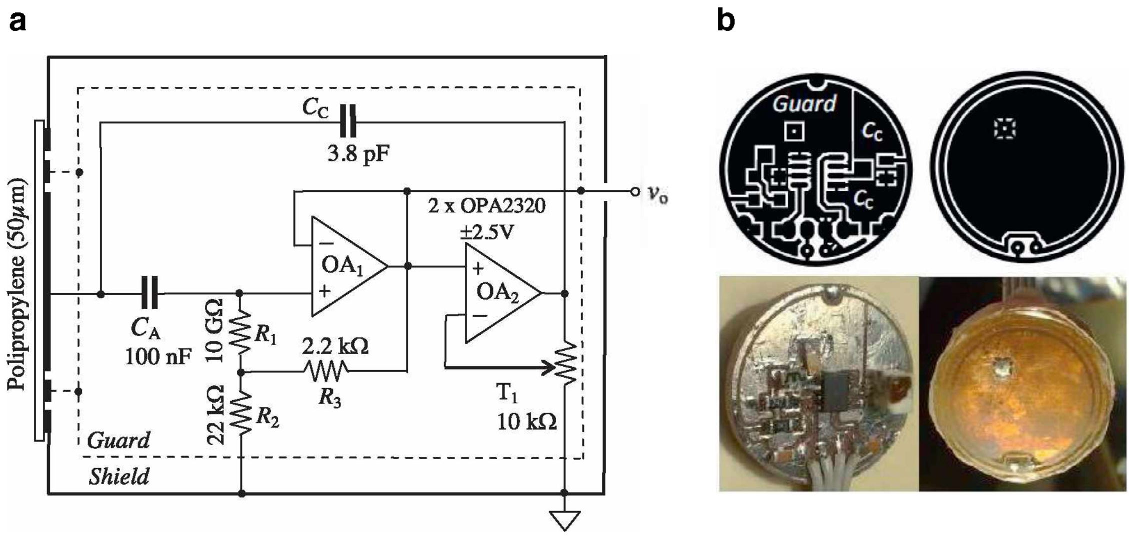

Fig. 3. (a) Complete circuit of the proposed CE and (b) printed circuit board and photo of the built prototype. Note that the proposed CE requires a reduced number of parts and includes a guard ring, plus a shield (ground) ring.

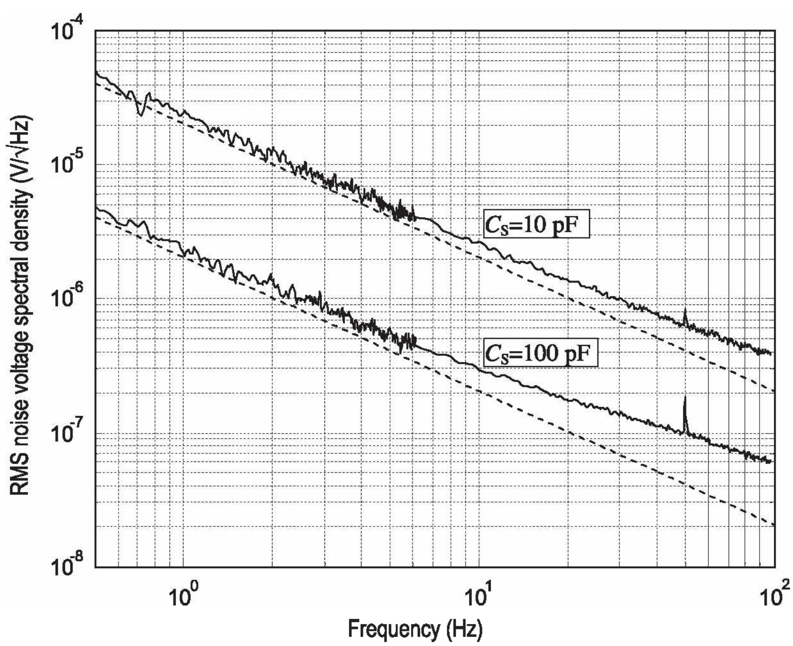

Fig. 4. Noise Power Spectral Density (PSD) of the proposed capacitive electrode for C_S=10 pF and C_S=100 pF. The expected noise due to resistor R_1, given by (8) for each C_S value, is indicated in dashed line.

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:

$$E_O^2 \approx \frac{kT}{(\pi C_S)^2} \frac{1}{R_1 f_1}. \tag{8}$$

Considering R_1=10 GΩ, T=300 K, C_S=100 pF and f_1=0.05 Hz, the total noise E_O results in 9 μV_{RMS}, which is a reasonable value for ECG signals. If f_1=0.5 Hz is considered, the noise is limited to 2.9 μV_{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 ± 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 μV_{RMS}. This value is a little higher than the 2.9 μV_{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 μm 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.

Fig. 5. ECG signals picked: (a) by standard disposable wet electrodes; (b) by the built capacitive electrodes placed on the skin; and (c) through a cotton T-shirt. Records in (a) and (b) were acquired simultaneously, and (c) corresponds to a different trial. Signal bandwidth was limited to 0.05–100 Hz and linear trends subtracted.

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 capacitive 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_{BEQ} 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_n 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 that i_n 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:

$$e_o^2 \approx e_n^2(1+R_2/R_3)^2 + e_{R1}^2(1+R_2/R_3)^2 + e_{R2}^2 + e_{R3}^2(1+R_2/R_3)^2; \tag{9}$$

and replacing resistors' noise PSDs by the Johnson–Nyquist formula (e_R=4kTR):

$$e_o^2 \approx e_n^2(1+R_2/R_3)^2 + 4kTR_1(1+R_2/R_3)^2 + 4kTR_2 + 4kTR_3(1+R_2/R_3)^2. \tag{10}$$

Given that $R_1 \gg R_2, R_3$; $R_2/R_3 \gg 1$ and $e_n \ll e_{R1}$; Eq. (10) can be approximated by:

$$e_{R_{BEQ}}^2 \approx 4kTR_1(R_2/R_3)^2; \tag{11}$$

which can be written as:

$$e_{R_{BEQ}}^2 \approx 4kTR_{BEQ}(R_2/R_3). \tag{12}$$

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