

Bio-potential amplifier for potential gradient measurements

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Abstract. This work proposes a bio-potential amplifier suitable for measurements from an electric potential gradient sensor, in electro-encephalography (EEG). The sensor is an array made by three electrodes placed on the vertices of an equilateral triangle of reduced size. Measuring the gradient requires small separation between electrodes hence, very low amplitude signals, of a few μV , are obtained. Therefore, it is important to minimize amplifier noise and electromagnetic interference effects. In the proposed scheme, the first stage is a passive and balanced ac-coupling network adapted to the gradient configuration and the second stage is an 80 dB gain amplifier. The implementation requires a reduced number of components. Therefore, the circuit can be mounted just above the electrodes (active electrodes). The proposed amplifier was built and tested. It achieves a CMRR of 125dB at 50 Hz and an equivalent input noise voltage of $0,3\mu\text{V}$ RMS in the band 0.5 – 500 Hz. Finally, some preliminary results in the detection of occipital alpha rhythm are presented.

1. Introduction

Electroencephalography techniques (EEG) are used to obtain the electric potential distribution on the scalp. Solving the inverse problem, it is possible to estimate the bioelectric sources that produced the observed potential. Within this context, gradient sensors present some advantages against classical configurations of EEG electrode: they do not require a reference potential and they are less sensitive to artifacts and spontaneous brain noise [1].

An electrode array to implement a gradient sensor is proposed in [1]. It consists on three electrodes composing an equilateral triangle (figure 1). The reduced size of the array, 1 cm side length, results in low amplitude EEG gradient signals of about a few μV , requiring a low-noise amplifier and a high rejection of electromagnetic interference (EMI). For these reasons, the front-end was mounted on the electrode itself [2], composing an active electrode array. In order to reduce the size of amplifier, an implementation with a reduced number of components was preferred [3].

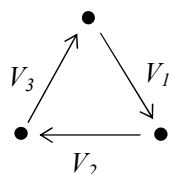


Figure 1. Electrodes distribution for the electrical potential gradient sensors configuration.

A high-gain front-end is desirable to design low noise bio-potential amplifiers. In order to achieve this, ac coupling is needed to avoid amplifier saturation due to dc electrode offset potentials. Spinelli *et al.* [4] proposed a passive network that provides ac coupling for differential mode signals preserving a high common mode rejection ratio (CMRR). This circuit could be used for each one of the gradient sensor channels, but a considerable number of components are required. This work generalizes the circuit proposed in [4] and presents an ac coupling network especially intended for gradient sensors, hence reducing the number the components while preserving a high CMRR.

2. Methods and Materials

The proposed scheme involves two stages; the first one is a passive and balanced ac-coupling network adapted to the gradient configuration, and is followed by an amplifier stage with a gain of 80 dB.

2.1. Ac-coupling stage

Figure 2 (a) shows the ac-coupling network presented by Spinelli *et al.* [4] for bipolar channels applied to the gradient sensor scheme. The proposed ac-coupling network for gradient sensors, which is shown in Fig2 (b), can be seen as a generalization of the previous circuit. It provides ac-coupling for the three input channels and reduces the number of components.

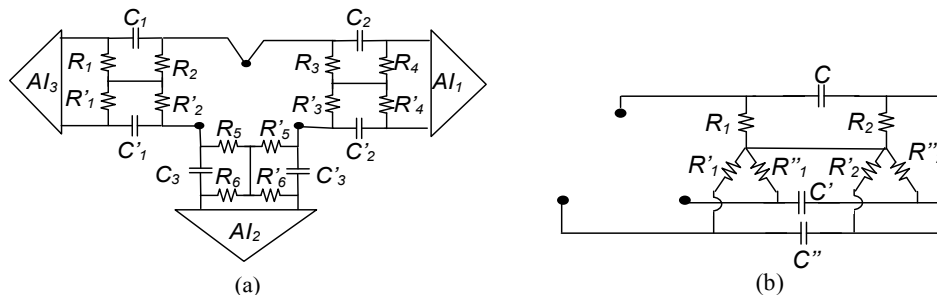


Figure 2. (a) Ac-coupling network applied to the gradient sensor scheme. (b) Proposed ac-coupling network adapted to the gradient configuration.

In both circuits the ac-coupling networks are not grounded; if a common mode input voltage is applied, no currents flow through the network and all nodes of the circuit achieve the same potential. This absence of potential difference due to common mode inputs implies an ideally infinite CMRR regardless of component tolerances. In practice, however, there are some grounded impedances (mainly due to op amps' input capacitances) and the CMRR becomes finite due to the *potential divider effect* [2]. Regarding an electrode impedance unbalance ΔZ_e and common mode input impedances Z_{ic} , the CMRR is approximated by:

$$CMRR \cong \frac{Z_{ic}}{2 \cdot \Delta Z_e} \quad (1)$$

If the network unbalances and the electrodes impedances are smaller than R_1 and R_2 , the differential mode (DM) transfer is the same than the DM transfer of the circuits in the figure 2 (a), and it is:

$$G_{DD} \cong \frac{\tau_{DD} s}{1 + s \cdot \tau_{DD}} \quad (2)$$

where $\tau_{DD} \cong R_2 C$.

The ac-coupled circuit introduces thermal noise due to the resistor included in the network. In this type of ac-coupled circuit, the signal transfer function and the noise transfer function are different [5].

The differential mode circuit, including noise sources of resistors, is shown in figure 3. The voltage sources e_{Rd1} and e_{Rd2} represent the differential-mode voltage noise corresponding to resistors R_1 and R_2 respectively. Assuming uncorrelated resistors noise, the root-mean-square (RMS) values of these sources are:

$$e_{Rd1}^2 \cong 2 \cdot e_{R1}^2 \quad y \quad e_{Rd2}^2 \cong 2 \cdot e_{R2}^2 \quad (3)$$

The noise transfer functions for e_{Rd1} and e_{Rd2} are given by:

$$G_{\text{Ruido.Rd1}} = A_{R1} \frac{s}{1 + s \cdot \tau_{1p}} \quad y \quad G_{\text{Ruido.Rd2}} = \frac{1 + s \cdot \tau_{2z}}{1 + s \cdot \tau_{2p}} \quad (4)$$

where

$$A_{R1} \cong CR_e; \quad \tau_{2z} \cong CR_e; \quad \tau_{1p} = \tau_{2p} = R_2C \quad (5)$$

and R_e represents the electrode impedance.

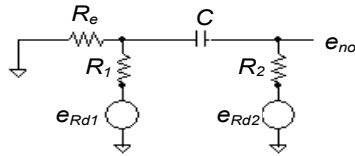


Figure 3. Differential-Mode circuit of the proposed ac-coupling network, including noise sources of resistors.

Given that, in order to avoid load effects, the resistor R_1 is made much larger than R_e , the contribution of R_1 is not significant and the total noise is dominated by R_2 . The noise spectral density of this resistor was denoted as η_{R2} and it is given by the Nyquist formula:

$$\eta_{R2} \cong 4KTR_2 \quad [V_{\text{RMS}}^2/\text{Hz}] \quad (6)$$

where K is the Boltzman constant ($K = 1,38 \cdot 10^{-23} J / K$) and T the absolute temperature.

The RMS value of the output noise in the bandwidth $[\omega_{\text{inf}}; \omega_{\text{sup}}]$ results:

$$e_{no}^2 \cong \frac{\eta_{R2}}{\pi} \left[\frac{(1/\omega_{\text{inf}} - CR_e)}{(R_1C)^2} + (R_e/R_1)^2 (\omega_{\text{sup}} - 1/(CR_e)) \right] \quad (7)$$

Considering that the low cut-of frequency is $1/(CR_1)$ and supposing $R_e \ll R_1$, the second term in (7) can be neglected and (7) becomes:

$$e_{no}^2 \cong \frac{4KT}{\pi C} [V^2] \quad (8)$$

This expression shows that the total noise depends exclusively on the capacitor C . So, this capacitance should be designed as high as possible.

2.2. Amplification Stage

The second stage of the proposed bio-potential amplifier (BA) provides 80dB gain by using three instrumentation amplifiers (IA) INA111 of *Texas Instruments*. This IA was selected because of its low current and voltage noise. The complete amplifier scheme is shown in figure 4.

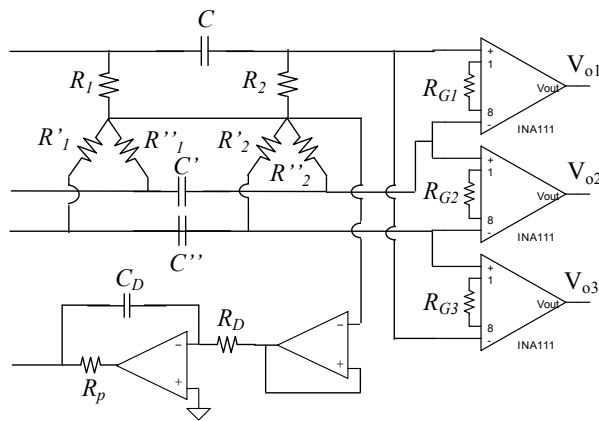


Figure 4. AI complete circuit for electric potential gradient sensor. The components values are: $R_1 = R'_1 = R''_1 = R_2 = R'_2 = R''_2 = 10\text{M}\Omega$; $C = C' = C'' = 0.2\mu\text{F}$, all at 5%; $R_{G1} = R_{G2} = R_{G3} = 5.6\Omega$; $R_p = 910\text{K}\Omega$; $R_D = 4.9\text{K}\Omega$ y $C_D = 22\text{nF}$. A_1 y A_2 : LF353 (dual operational); $A11$, $A12$ y $A13$: INA111 (instrumentation amplifier).

3. Experimental Results

As a validation of the proposed scheme, a prototype was built and tested. In order to avoid power-line interference, all the measurements were made shielding the amplifier and the array of electrodes with a metallic grounded box.

3.1. Frequency Response

The frequency response was evaluated on a bench using a signal generator. The experimental measurements, presented in figure5, show a good correspondence with the theoretical results. They also show that for high frequencies, the amplifier bandwidth is limited by the IA gain-bandwidth product (GBP). The lower cutoff frequency is approximately 80 mHz, the upper one is 6 KHz, the gain in the band-pass is 79dB and the input of amplifier range is $\pm 680\mu\text{V AC}$.

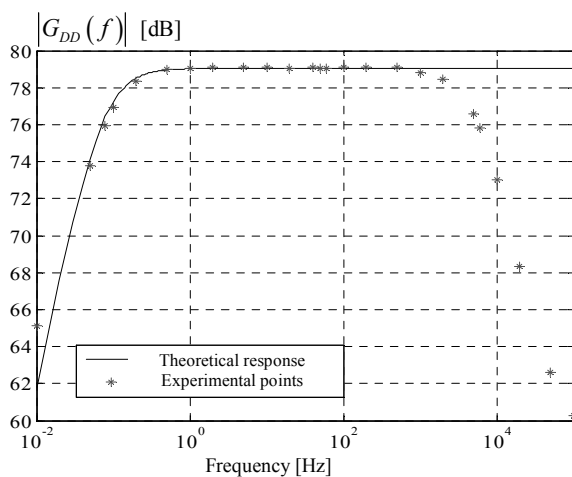


Figure 5. Proposed amplifier frequency response, theoretical and experimental results.

3.2. Common Mode Rejection Ratio (CMRR)

The CMRR was measured including resistors at the amplifier inputs, to simulate an electrode impedance unbalance of $9\text{K}\Omega$. In this condition, the CMRR at 50 Hz was of 95dB.

3.3. Equivalent Noise

Noise measurements were performed on the three channels with their inputs short-circuited. The obtained spectral density is shown in figure 6 and the total noise in the band of 0.5 to 500 Hz was always less than $0.3\mu\text{V}$.

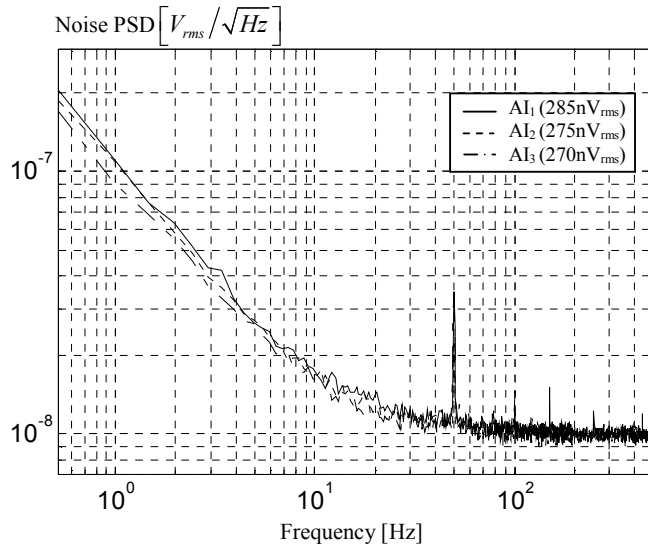


Figure 6. Voltage noise spectral density measured in each BA's output and their RMS values for the bandwidth of 0.5 to 500Hz.

3.4. Biopotential Acquisition

The proposed amplifier was tested acquiring real EEG signals using three standard EEG disk electrodes composing an equilateral triangle of 2 cm size, which was placed on the occipital zone of the scalp (Figure 7). In this test, a *Driven Right Leg* (DRL) circuit was included; this circuit improves the CMRR on about 30dB at 50 Hz increasing the total CMRR to 125dB for an electrode impedance unbalance of $9\text{K}\Omega$.

The amplifier was battery-powered and mounted on the electrodes themselves (active electrodes) to minimize capacitive coupling to the leads. Three isolation amplifiers (ISO122V) were used to ensure subject safety and to provide grounded signals for non-isolated instruments. The outputs of the three channels were measured simultaneously by using two synchronized digital oscilloscopes. The bandwidths of the signals were limited to 32Hz to minimize noise power.

Figure 8 shows an example of alpha rhythm detection. This rhythm presents components around 10 Hz and arises when the subject closes his eyes. The acquisition time was 4 seconds and the sampling frequency 2.5 KHz.

The sum of the three channels should be zero, but in practice it did not happen exactly, due to noise and gain unbalances of amplifiers. The measured sum of these channels was of $0.8\mu\text{VRMS}$.

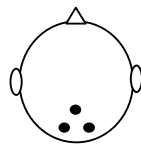


Figure 7. Electrodes distributions in the detection of occipital alpha rhythm.

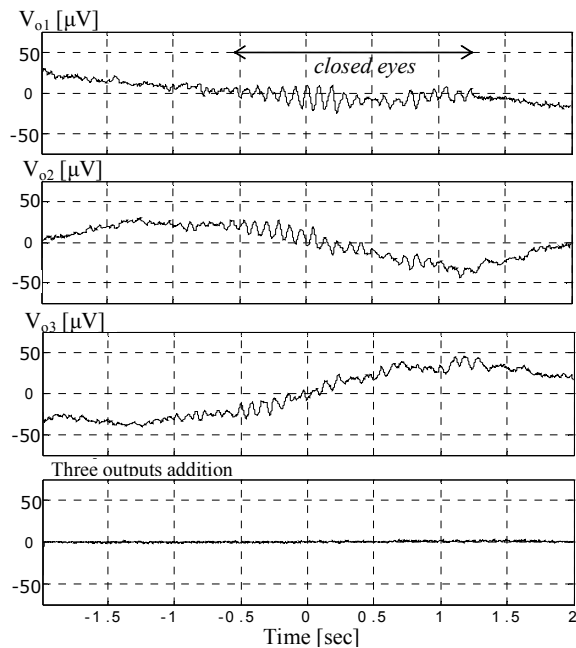


Figure 8. Signals measured in alpha rhythm detection with the gradient potential sensor, and the sum of the three differential signals.

4. Conclusions

A bio-potential Amplifier suitable to measure electrical gradient potential on the scalp was proposed. It includes an ac-coupled network that was specially adapted for the gradient sensor, which is composed by three electrodes placed on the vertices of an equilateral triangle.

The resulting amplifier was built and tested. It achieves a 79 dB gain, a CMRR of 125dB at 50 Hz and an equivalent input noise voltage of 0.3 μ V RMS in the bandwidth of 0.5–500 Hz. The prototype was also tested with real EEG signals, acquiring and detecting occipital alpha rhythm.

Acknowledgments

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