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A digitally controlled universal second-order filter block for biosignals with CMOS D/A converters

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A universal second-order low-pass filter block with a digitally adjustable cutoff frequency is described. The filter section replaces conventional manually adjustable low-pass filters and is suitable for a wide range of biosignals such as EEG, EMG, and the recording of evoked potentials (ERPs). The advantages of digital control of filter parameters are briefly outlined and construction hints are given for a number of applications. The universal state-variable filter section may be programmed by any laboratory computer and does not require a complex installation procedure.

In the earlier days of psychophysiological measurements, the assessment of biosignals was accomplished solely by means of analog circuitry (Goldstein & Free, 1979). The documentation of recorded biodata was carried out with multichannel mechanical pen driving circuits, which were much too inert to follow high frequencies appropriately. The data was stored on an analog tape for further off-line analysis; thereafter, it was evaluated and measured by visual inspection. In the course of the development of modern laboratory computers, many stages of manually adjustable data acquisition equipment were replaced by digitally controlled units. Now, at least for the time being, on-line data assessment and appropriate preprocessing of biosignals are standard techniques in psychophysiological laboratories (Ruchkin, 1988).

In routine clinical work, a constant hardware configuration is essential, whereas biomedical, psychophysiological, and pharmacological research activities demand more flexibility. Since experimental settings frequently have to be changed because of particular questions of interest, a manually adjustable circuitry increases the probability of operator errors. This is particularly true, whenever important parameters such as the sample rate of the data acquisition system are subject to change due to different recording requirements. It is well known that aliasing effects during digitizing of analog signals may irreversibly distort spectral information contained in the data. Therefore, a digitally controlled antialiasing filter, set automatically to the designated Nyquist frequency of at least twice the highest frequency in the recorded data, is considered mandatory (Coppola & Morgan, 1987; Ruchkin, 1988).

Since different biosignals occupy varying frequency ranges—ranging, for example, from 0.1-70 Hz in electroencephalographic recordings to a required bandwidth of 0-10 kHz for the recording of intracellular voltages (Regan, 1972; Thompson & Patterson, 1973)—it is more reliable and elegant to control variations of filter adjustments by means of sophisticated software. This allows for individual programming of different experimental designs without one's having to check hardware adjustment controls to see that they are correct along the data acquisition path.

This paper centers on a universal filter section with preset gain and circuit Q-factor and digitally selectable corner frequency. The dimensioning of the frequency-dependent components is based on the specific requirements for qualitatively different types of low-pass filters. Since the versatile state-variable filter arrangement may be easily configured as a high-pass, low-pass, resonant, or notch filter, the calculations may serve as a model for other implementations. The basic filter circuit is designed to provide an attenuation of −12 dB/octave or −40 dB/decade, respectively, which is appropriate for standard applications. The realization of a fourth-order filter with a much steeper gradient of −24 dB/octave or −60 dB/decade necessitates a serial configuration of two single units, which in turn requires a somewhat more refined dimensioning of frequency-dependent components.

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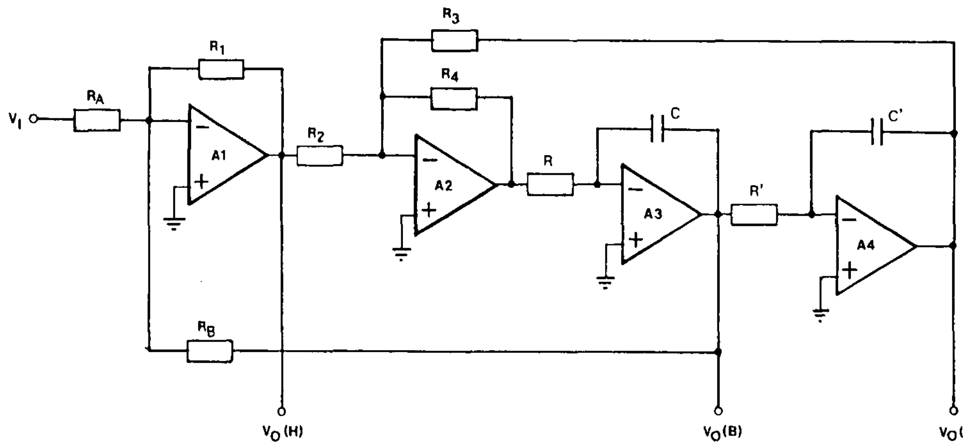


Figure 1. A second-order state-variable filter circuit.

The State-Variable Filter

The filter section in Figure 1, which appears in numerous minor variations in technical publications, is known as the "state-variable-filter" or "universal filter," since it provides simultaneously high-pass, band-pass, and low-pass outputs (Lancaster, 1981; Tietze & Schenk, 1986). The parameters gain, bandwidth, and resonant frequency are independently adjustable over a wide range. At the expense of an additional operational amplifier, the summation of the high-pass and low-pass output gives a notch filter characteristic. The gain of the filter circuit in Figure 1 at the resonant frequency is given by Equation 1:

$$A_o = -\frac{R_B}{R_A} \qquad (1)$$

The circuit Q-factor for the band-pass output $V_{out}(B)$ in Figure 1, defined as the quotient of the resonant frequency divided by the 3-dB bandwidth, can be derived by using the Equation 2:

$$Q = \frac{R_2}{R_4} \cdot \frac{R_B}{R_1} \cdot \sqrt{\frac{R_4}{R_3}} \qquad (2)$$

The resonant frequency of the circuit is given by Equation 3:

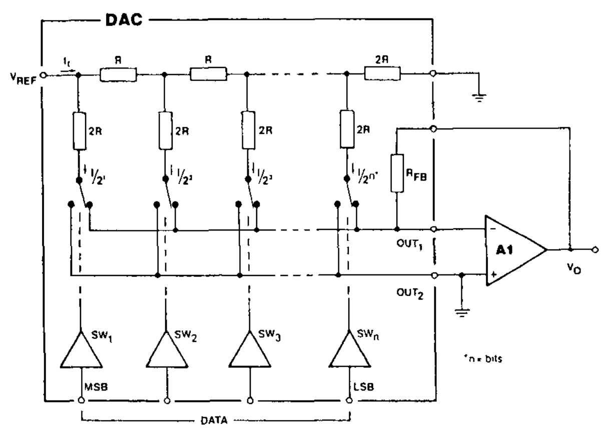
$$f_o = \frac{1}{2\pi RC} \cdot \frac{R_4}{R_3} \qquad (3)$$

This circuit meets the requirements for the filtering of biodata from EEG, ECG, and EMG to high frequency intercellular signals and brainstem responses. A next step is to add a computer-filter interface to the basic filter design in order to provide a convenient software-control of the cutoff frequency. This is accomplished by substituting the frequency-dependent resistors R in Figure 1 by a switched resistor ladder network. CMOS D/A converters are ideally suited for this purpose.

Employing CMOS D/A converters (DACs)

CMOS D/A converters are used as multipliers and attenuators for high-level signals because they provide very low distortion, and minimal zero offset and noise. A simplified sketch of the internal structure of a DAC based on a switched 2R/R ladder network and the graphical symbol for circuit diagrams is delineated in Figure 2 (A&B). The output voltage V_{out} is the inverted product of the digital attenuation code and the voltage V_{ref} applied to the input. The input current I is divided into binary-weighted fractional currents steering the inverting input of the operational amplifier A_1 via the digitally controllable switches $SW_1 \ldots SW_n$. The maximum attenuation of the ladder

(A)



(B)

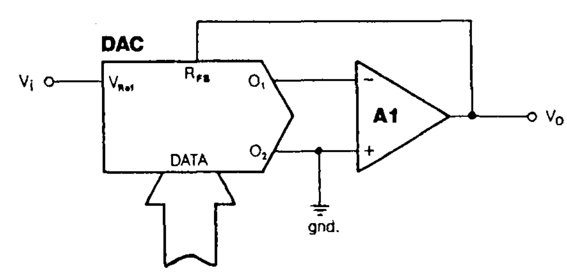


Figure 2. (A) The simplified principle of a CMOS DAC multiplier. (B) Circuit schematic of the circuit in (A).

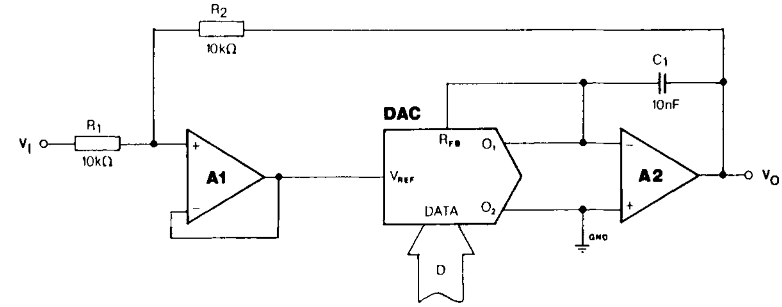


Figure 3. A simple first-order low-pass filter with digitally selectable corner frequencies.

network is achieved when all digital data inputs are set to zero.

A relatively simple first-order low-pass filter using a CMOS DAC is shown in Figure 3. The cutoff frequency of the circuit in Figure 3 is given by applying Equation 4 to calculate the radian frequency ω, where D is the fractional binary number of the digital code sent to the DAC:

$$\omega = \frac{R_1}{(R_1+R_2)} \cdot \frac{D}{(C \cdot R_{DAC})} \tag{4}$$

Assuming a DAC ladder resistance of about 10 kΩ and a digital code of FF HEX (1111 1111) applied to the DAC, the filter frequency will be set to about 5000 Hz, whereas a digital code of 1 sets the corner frequency to approximately 20 Hz. However, the circuit in Figure 3 reveals a number of disadvantages. First, the maximum rejection in the stop band of −6 dB/octave does not necessarily meet the requirements for an effective filtering of biodata. Moreover, the radian frequency ω in Equation 4 is largely dependent on the ladder resistance of the DAC, which may vary considerably between the devices (Burton, 1986). Unfortunately, the ladder resistance R varies between 0.8 R and 2 R with identical digital input codes, where R is typically 11 kΩ. The circuitry in Figure 4 avoids that problem and provides a maximum attenuation of −12 dB/octave beyond the corner frequency.

Here, the input signal is fed into the inverting input of amplifier A_1, which is the summing node of the signals delivered by the integrators A_3 and A_5 via resistors R_3 and R_4. The implementation of the amplifiers A_2 and A_4 in Figure 4 makes the control of cutoff frequency independent of device mismatches due to variations of individual ladder resistances. Both independent sections of the dual 8-bit CMOS DAC AD7528 (Analog Devices, Inc.) are matched, providing an accuracy of 1%. In the potentiometric configuration as a divider network, the precision of attenuation via R_D and R_{FB} is determined by the accuracy of the internal resistors on the chip. The differential nonlinearity is in the range of ±1 LSB over the entire operating temperature range and guarantees monotonicity. The digital control code is applied to Pin 7 (MSB, D1) through Pin 14 (LSB, D0), in the same manner as already described in Equation 4. Pin 6 (DAC A/B select) and Pin 15 (chip select) control the internal data

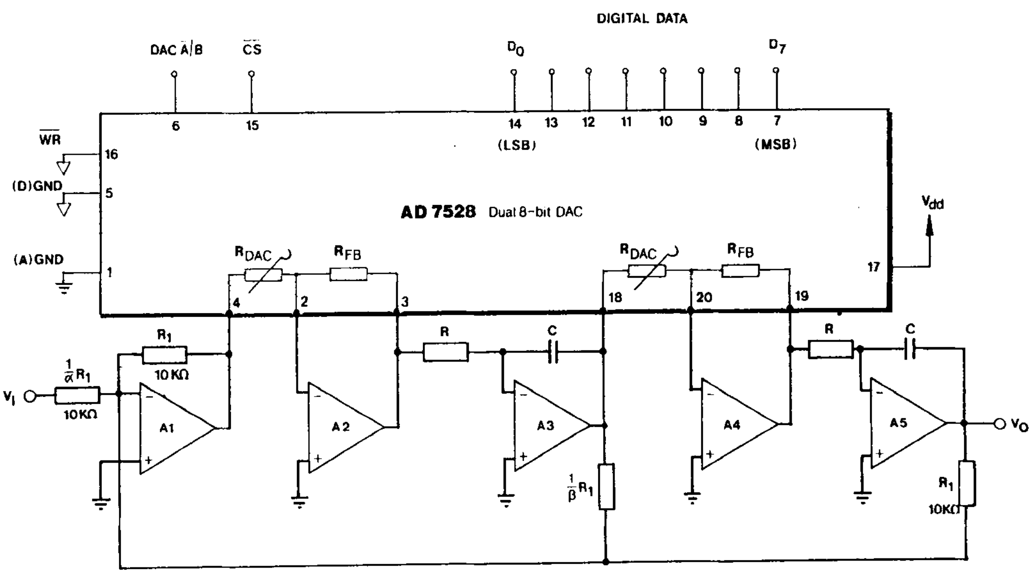


Figure 4. A second-order low-pass filter section with a dual 8-bit DAC.

latches. The digital code has to be applied to both units in a two-step sequential procedure in order to set the filter correctly.

The −3 dB cutoff frequency of the circuitry in Figure 4 is given by Equation 5:

$$f_o = \frac{1}{2\pi RC} \cdot \frac{D}{D_{max}+1} \tag{5}$$

Assuming C to be 10 nF and R 16 kΩ (Figure 4), a digital code of FF HEX (1111 1111) switches the filter frequency to about 995 Hz, whereas a code of 1 shifts the −3 dB cutoff point to approximately 4 Hz. Note that a zero code is undefined, since it causes the multiplier to deliver a zero output, which in turn disables the closed dc-loop of the op-amps and forces the amplifiers A_3 and A_5 into saturation. The integration resistor R_i (attenuation via DAC and R) varies with the digital control code, covering a range from about 20 kΩ up to over 5 MΩ and is calculated with Equation 6:

$$R_i = R\,\frac{D_{max}+1}{D} \tag{6}$$

The resistor $1/\alpha R_1$ in Figure 4 determines the overall gain of the filter. The resistor $1/\beta R_1$ is responsible for the filter characteristic. Both values are set independently of each other, i.e. when $1/\beta R_1$ is held constant, cutoff frequency and the overall gain can be set separately to the desired values. The coefficient β equals the reciprocal of the circuit-Q-factor Q at the band-pass output node (output of A_3), which is determined by means of the terms in Equation 7:

$$Q_i = \sqrt{\frac{b_i}{a_i}},\ \beta = \frac{a_i}{\sqrt{b_i}} = \frac{1}{Q_i},\ \frac{1}{\beta} = Q_i \tag{7}$$

For the low-pass output, the corner frequency is calculated by means of Equation 8:

$$f_g = \frac{\sqrt{b_i}}{2\pi RC} \cdot \frac{D_{max}+1}{D} \tag{8}$$

The coefficient α is defined by dividing the overall gain by the reciprocal circuit-Q-factor as shown in Equation 9:

$$\alpha = \frac{A_o}{\beta} \tag{9}$$

The value of resistor $1/\beta R_1$ determines the filter characteristic. In the relevant literature on the design of filters (Lancaster, 1981; Tietze & Schenk, 1986) calculated values are given for the coefficients a_i, b_i, and Q_i dependent on the characteristic of Bessel, Butterworth, and Chebychev filters. For the filtering of biosignals, either filters of the Bessel or the Butterworth response are employed. However, the Bessel characteristic provides a linear phase delay and an optimized transient response, which is obligatory in event-related potential analysis (Coppola & Morgan, 1987). For a second-order Bessel filter the value of a_i is 1.3617 and b_i equals 0.6180.

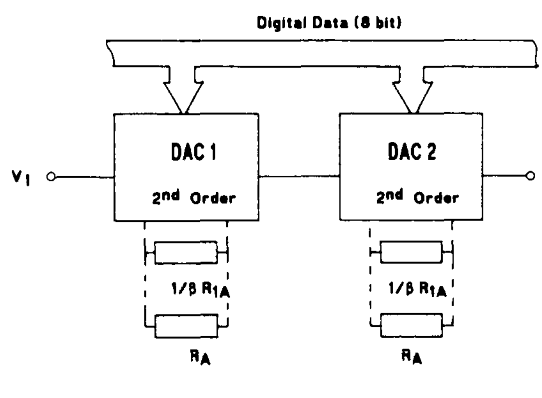
Thus, taking Equation 7 into consideration, the resistor is dimensioned as $Q_i \cdot R_1 = \approx 0.58 \cdot 10$ kΩ = 5.8 kΩ in order to give a Bessel characteristic. The Q-factor Q_i for a Butterworth response is about ≈ 0.71, giving a value of 7.1 kΩ for $1/\beta R_1$.

The procedure for the design of filter sections of higher order employs the same design strategy. Assuming a fourth-order low-pass filter with a −24 dB attenuation, the overall arrangement is briefly delineated in Figure 5. Here, two sections of Figure 4 are assembled in serial configuration. Assuming a Bessel response, Stage A should have a Q_i of 0.52 and Stage B should have a Q_i of 0.81, respectively, so that the values of the resistors $1/\alpha R_1$ and $1/\beta R_1$ are 5.2 kΩ and 8.1 kΩ. Regarding Equation 8, it is evident that different coefficients b_i cause minute shifts in the cutoff frequency of the single-filter subsections constituting a filter of higher order.

With component values of C = 10 nF and R = 16 kΩ (Figure 4), the frequency range of the filter may be adjusted from 4-995 Hz. This is a good choice for filtering auditory and visually evoked potential data. Changing the value of the condenser C to 1 nF causes the cutoff frequency to vary from 40-9950 Hz, which is ideal for a flexible antialiasing filter and for the filtering of brainstem recordings, somatosensoric potentials, and intracellular signals.

Conclusion

A digital filter design using CMOS DACs is described. It is very well-suited for computer-controlled biosignal amplifier systems. Depending on the type of DACs, almost any type of filter resolution, attenuation, and overall transfer characteristic may be realized. In digitally controlled *analog* data systems, a digitally controlled filter design offers the advantage of quickly changing the filter settings to predefined sets. In digitally controlled *digital* data systems with less computer power, it is possible to implement rather complex and adaptive filter designs to



Q_i	0.52	0.81
b_i	0.49	0.39

Figure 5. A fourth-order filter block with Bessel response.

unburden the CPU from digital filtering tasks. Even if digital signal processing and filtering should become available at considerable speed and affordable prices, it will still be necessary to provide antialiasing low-pass filtering before digitizing the analog data. If various sample rates have to be used, it will be necessary to adapt the corner frequency of the cutoff filter to the given Nyquist frequency. Thus, a software-controlled, state-variable filter design employing CMOS DACs is a good choice: it meets the requirements of various filter applications in psychophysiological and medical research.

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