COMPARISON OF SEVERAL MODES IN SIMPLE ARC SECOND-ORDER FILTER

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Abstract

In this paper the popular, multiple-feedback, ARC single opamp, highpass second-order filter is proposed in several types of modes, namely voltage, current and hybrid ones. These modes are studied and compared in detail. Computer experimental results are given supporting the theory.

Keywords:

filter, ARC biquad, voltage current and hybrid mode

1. Introduction

The single opamp, multiple-feedback, Huelsman biquad (ARC second-order filter) is one from well-known and popular opamp applications [1]. This ARC filter is very simple - it contains minimum number of components. There is the voltage signal processing in classical voltage mode (VM) using a conventional opamp. There the parasitic inherent features of the real opamp play here more characteristic roles, specially in high frequency region [2]. It is generally known that the frequency range of these VM filters can be only (10 - 20) times less than the unity-gain bandwidth f_T (GBP) of the opamp. Furthermore a large signal slewing effect of the opamp may restrict the operating bandwidth of these VM filters too.

For this reason a current mode (CM) of the circuits has been developed in [2], where the current signal processing is used. There an adjoint transformation is suggested as straightforward method to obtain CM filters from their VM counterparts. In this paper we shall apply the adjoint network concept to the Huelsman second-order filter. Although any type of filtering function can be realized with this class of novel circuits, we shall focus our attention on the highpass one only. The proposed modes are studied and compared in detail. Computer experimental results are given supporting the theory.

At first from principle of methodology we would recommend to distinguish more modes in which the given circuit can work. There are especially applied basic modes from [1] or [3] - namely classical voltage (VM) and adjoint current (CM) one, when all subnetworks operate either with voltage or current signals. Furthermore if certain subcircuit (usually an active element) operates in the opposite mode to basic one we will talk about hybrid modes, namely V/CM or C/VM respectively. A conventional opamp operates in VM while a current conveyor and a transimpedance amplifier in CM respectively.

2. Voltage mode in ideal case

The well-known Huelsman highpass second-order filter in the classical VM is shown in Fig. 1. Assuming an infinite gain the ideal opamp can be taken as a nullor (a pair of nullator and norator). Then, in straightforward manner, this circuit has been analyzed by program COCO.

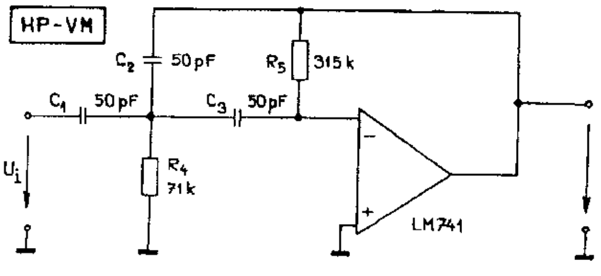

Fig. 1
The Huelsman second-order highpass filter in classical voltage mode (VM) using opamp.

A resulting voltage transfer function can be written in following symbolic form

$$K(p) = \frac{U_o}{U_i} = \frac{ap^2}{p^2 + b_1 p + b_0} = \frac{-K_0 p^2}{(p-\hat{p}_1)(p-\hat{p}_2)} \qquad (1)$$

where the coefficients are

$$a = -K_0 = -\frac{C_1}{C_2} \qquad (2)$$

$$b_1 = \frac{C_1 + C_2 + C_3}{R_5 C_2 C_3} \quad , \quad b_2 = 1 \qquad (3)$$

$$b_0 = \frac{1}{R_4 R_5 C_2 C_3} \qquad (4)$$

Furthermore the design equations are obviously

$$\omega_p = \frac{1}{\sqrt{R_4 R_5 C_2 C_3}} \tag{5}$$

$$Q = \sqrt{\frac{R_4}{R_5}}\left[\frac{C_1}{\sqrt{C_2 C_3}} + \sqrt{\frac{C_3}{C_2}} + \sqrt{\frac{C_2}{C_3}}\right]^{-1} \tag{6}$$

To compare performances of several modes this highpass VM biquad (Fig. 1) has been design for following specifications: f_c = 20 kHz, K_0= 0 dB and Butterworth approximation. The resulting passive component values are: $C_1 = C_2 = C_3 = 50pF$, $R_4 = 71k\Omega$, $R_5 = 315\,\Omega$. Then the voltage transfer function (1) has following numerical form

$$K_p = \frac{-1.00p^2}{1.00p^2 + 1.904\cdot 10^5 p + 1.788\cdot 10^{10}} =$$

$$= \frac{-1.00(p-0)(p-0)}{[p-(-9.52+j9.39)\cdot 10^4][p-(-9.52-j9.39)\cdot 10^4]} \tag{7}$$

and poles and zeros location is shown in Fig. 3a - double zero at initial and two complex conjugate poles in left half-plane. The magnitude and phase frequency responses for this ideal case are shown in Fig. 4, the curves (a).

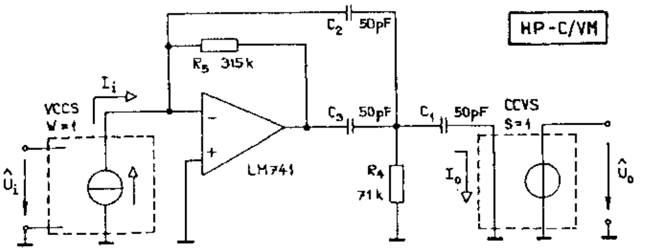

Fig. 2
The proposed biquad in the hybrid current-voltage mode (C/VM) using opamp.

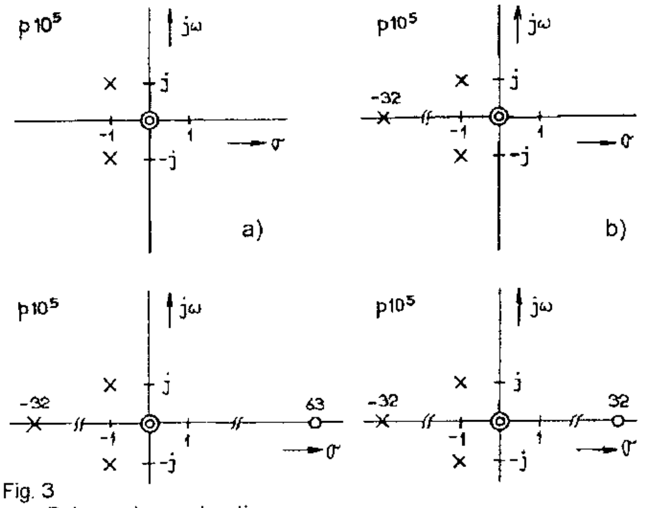

Fig. 3
Poles and zeros location
a) for voltage mode and an ideal opamp (nullor),
b) for voltage mode and the real opamp LM 741,
c) for the current-voltage hybrid mode using the opamp LM 741 - partial zero-pole compensation,
d) for the full zero-pole compensation in the hybrid mode.

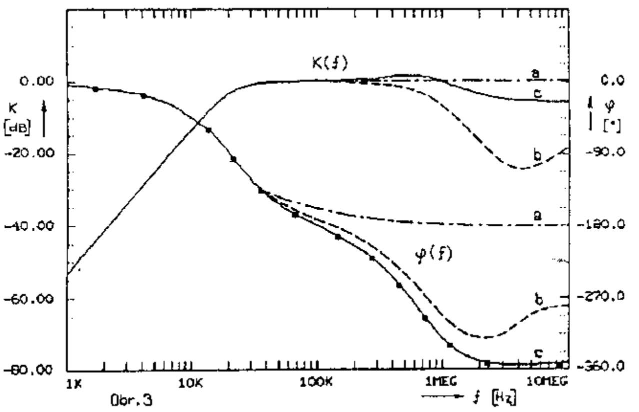

Fig. 4
Comparing of experimentally observed frequency magnitude responses for the Huelsman second-order highpass filter
a) in the ideal case using nullor (VM and CM is identical), and approximately for the real current mode (CM) using a current opamp,
b) in the real voltage mode (Fig. 1) using the opamp LM 741,
c) in the hybrid current-voltage mode (Fig. 2) using the opamp LM 741.

3. Real voltage mode

Assuming a limited frequency response, the real opamp can be modeled by single parasitic pole (a simple integrator model) described by

$$A(p) = \frac{A_0}{1+\frac{p}{\omega_0}} \tag{8}$$

where $A_0 w_0$ is a gain-bandwith product.

Taking this model in circuit of Fig. 1 the symbolic analysis results now as follows

$$K_v(p) = \frac{U_{v0}}{U_{vi}} = \frac{a_v p^2}{b_{3v}p^3 + b_{2v}p^2 + b_{1v}p + b_{0v}} =$$

$$= \frac{-K_0 p^2}{(p-\hat{p}_1)(p-\hat{p}_2)(p-\hat{p}_3)} \tag{9}$$

$$b_{3v} = \frac{1}{A_0\omega_0}\left(1+\frac{C_1}{C_2}\right) \tag{10}$$

$$b_{2v} = 1 + \frac{1}{A_0}\left(1+\frac{C_1}{C_2}\right) +$$

$$+ \frac{1}{A_0\omega_0}\left(\frac{1}{C_2 R_4} + \frac{C_1+C_2+C_3}{C_2 C_3 R_5}\right) \tag{11}$$

$$b_{1v} = \frac{C_1+C_2+C_3}{C_2 C_3 R_5} + \frac{1}{A_0}\left(\frac{C_1+C_2+C_3}{C_2 C_3 R_5} + \frac{1}{R_4 C_2}\right) +$$

$$+ \frac{1}{A_0\omega_0 R_4 R_5 C_2 C_3} \tag{12}$$

$$b_{0v} = \frac{1+\frac{1}{A_0}}{R_4 R_5 C_2 C_3} \tag{13}$$

Furthermore evaluating these eq`s for our example, the voltage transfer function becomes now

$$K_V(p) = \frac{-3.14 \cdot 10^6 (p-0)(p-0)}{[p-(-9.24+j9.53)\cdot 10^4][p+(-9.24-j9.53)\cdot 10^4]} \cdot \frac{1}{[p-(-3.19)\cdot 10^6]} \tag{14}$$

and an additional parasitic pole turns up (Fig. 3b). The frequency responses for real VM are shown in Fig. 4, namely the curves (b). There is a greater distortion in passband of the magnitude response. It is an effect of two parasitic poles modeling of the opamp LM 741 in program MICROCAP.

4. Ideal current mode

To obtain the corresponding current version of the circuit from Fig. 1 we will apply the adjoint transformation given in [2]. Assuming ideal case - the nullor model of the opamp, the current mode can be obtained simply by interchanging the nullator and norator and also reversing the input and output of the network [2]. The resulting CM filter is shown in Fig. 2. The current transfer function of the ideal circuit $K(p) = I_{out}/I_{inp}$ has the identical expression of (1) and the given design equations are valid too. Also the frequency responses have the ideal form (a) in the Fig. 4. Note that using nullors both identical modes VM and CM can be obtained. Nevertheless it is not truth for the real opamp.

5. Hybrid mode

The real opamp, with the finite voltage gain, operates separately in the VM, while the filter in Fig. 2 in the CM respectively. Therefore this mode is called a hybrid current-voltage one. Symbolical analysis of the current transfer function for this C/VM now results in

$$K_c(p) = \frac{I_{co}}{I_{ci}} = \frac{a_{3c}p^3 + a_{2c}p^2}{b_{3c}p^3 + b_{2c}p^2 + b_{1c}p + b_{0c}} = \frac{-K_0 p^2 (p - \hat{n}_3)}{(p-\hat{p}_1)(p-\hat{p}_2)(p-\hat{p}_3)} \tag{15}$$

The denominator $D_C(p)$ is identical with $D_V(p)$ in (9) for the real VM

$$D_C(p) = D_V(p) \ , \qquad b_{ic} = b_{iv} \ , \tag{16}$$

but the numerator has an additional third zero in the right half-plane (Fig. 3c). Note that the double one stays in initial too. In our example

$$K_c(p) = \frac{-5 \cdot 10^{-1}(p-0)(p-0)(p-6.28\cdot 10^6)}{[p-(-9.23+j9.52)\cdot 10^4]} \cdot \frac{1}{[p-(-9.23-j9.52)\cdot 10^4][p-(-3.19)\cdot 10^6]} \tag{17}$$

The third zero can partially compensate the parasitic pole introduce by the opamp. It produces a flatter magnitude response (c) at high frequencies (Fig. 4) for the hybrid mode C/VM using the same LM 741 as in the VM.

Full compensation occurs as long as the modules of the third zero and the parasitic pole are equal (Fig. 3d). Then the magnitude response turns to the ideal curve (a) in Fig. 4. The third zero can be moved by proper design using following formula

$$\hat{n}_3 = \omega_0 A_0 \frac{C_3}{C_2} \tag{18}$$

Varying the ratio C_3/C_2 there is unfortunately an undesirable location of poles in the denominator of (15) too. Nevertheless we have tried to investigate experimentally this problem using the program MC-3. Results are shown in Fig. 5.

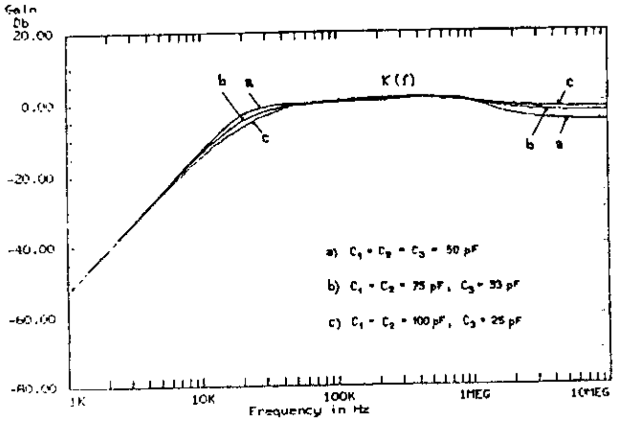

Fig.5
Frequency magnitude responses for the pole-zero compensation changing the values of capacitances.

6. Real current mode

A practical CM implementation of this filter will be discussed now. Namely the nullor model from the ideal CM (paragraph 4) has been implemented by the classical opamp in last section (paragraph 5), to obtain the hybrid mode C/VM. Nevertheless for the full current mode the grounded nullor can be subsequently implemented by a current conveyor CC II [2], a transimpedance amplifier [6] and current amplifier [2] respectively. There is consider the last one in Fig. 6. An acceptable current amplifier in monolithic IC form is shown in Fig. 6b [5]. The biquad in the adjoint full current mode HP-CM using this current amplifier is proposed in Fig. 6a. The frequency responses have an almost ideal forms (Fig. 4a).

7. Transimpedance amplifiers

Commercially available transimpedance opamps (also technologycally classified as current-feedback opamps), with compensation pin (AD 844 or AD846) are surely the right choice to implement high performance biquads in the current or hybrid V/CM modes [6].

The subnetwork in Fig. 1, consisting of the opamp and the resistor R_5, can be ingeniously replaced by the transimpedance opamp (TOA) using a compensation pin (Z), where the resistor R_5 is connected [6]. In such a way, the resulting circuit HP-V/CM has been obtained in Fig. 7a. Similarly from Fig. 2 the full current mode of HP-CM is proposed in Fig. 7b. Note that the frequency responses have also the almost ideal forms (Fig. 4a).

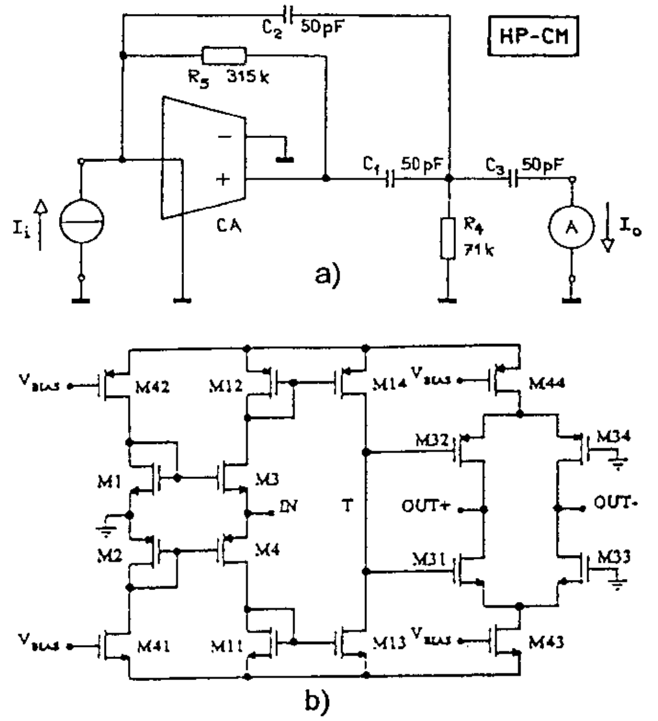

Fig. 6
The proposed biquad in the adjoint current mode (CM) using the current amp (a), simplified schematic circuit of the current amp (b).

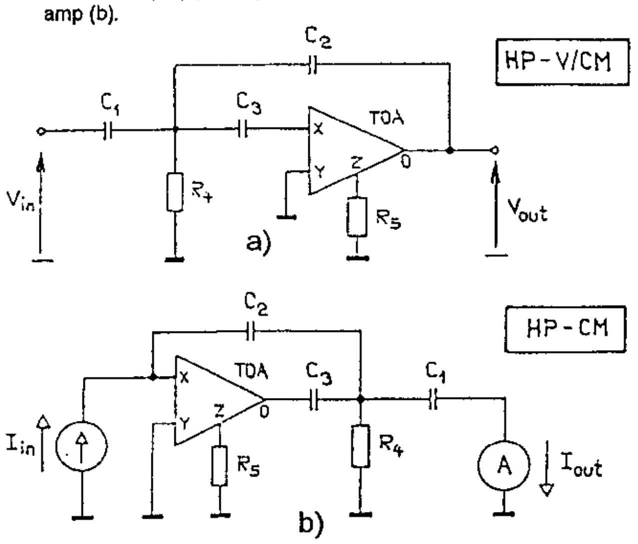

Fig. 7
The proposed second-order highpass filter using transimpedance amplifier
a) in the hybrid voltage-current mode HP-V/CM,
b) in the adjoint full current mode HP-CM.

8. Influence of opamp parasitic impedances

Many efforts have been made to extend the opamp use at higher frequencies (up to f_T), but with limited successes. Nevertheless the output resistance of the opamp plays a positive dominant role at higher frequencies. It is tested for the HP-VM in Fig. 8.

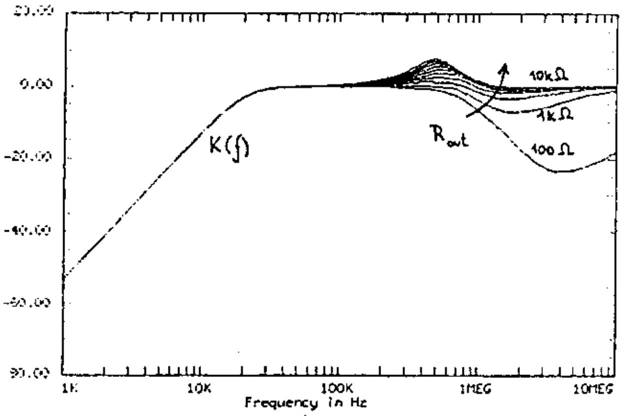

Fig. 8
Influence of the output resistance of the opamp on the magnitude response in the HP-VM filter.

Other possibility to increase the useful bandwith of the response at high frequencies is an additional series admittance or a RC cell at the input of the amplifier. Using this way a desired phase margin and stability can be obtained too.

9. Conclusion

Voltage, current and hybrid modes of simple ARC biquad have been analyzed, compared and experimentally tested. In any case an important conclusion may be generally pointed out as follows. The hybrid current-voltage mode and specially full current one are closer to the ideal and over a much wider bandwith than the corresponding voltage mode.

It is evident that a previously designed and optimized VM filter can be directly transformed into the CM filter, without losing its optimum characteristics. Due to the adjoint transformation which is used, the CM filter has identical network functions and their sensitivities, as the VM prototype.

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