



Araştırma Makalesi / Research Article
DESIGN OF UNIVERSAL CURRENT-MODE FILTER EMPLOYING CCCII+s

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ABSTRACT

In this paper, a universal current-mode filter for simultaneously providing all the general responses such as low-pass, band-pass, high-pass, notch and all-pass responses, is proposed. The proposed biquad employs only two grounded capacitors and three identical plus-type current-controlled conveyors (CCCII+s) thus it can easily be realized in integrated circuit (IC) process. Nevertheless, it requires a resistive component matching condition for all-pass response. Also, signal limitation and stability problems in analog filters are discussed. Finally, simulation results with SPICE program are given to confirm the theory.

Keywords: Current-mode, Universal filters, CCCIIs.

CCCII+ LAR KULLANILARAK AKIM-MODLU GENEL FİLTRE TASARIMI

ÖZET

Bu makalede, alçak-geçiren, bant-geçiren, yüksek-geçiren, bant-söndüren ve faz kaydıran cevaplar gibi genel filtre cevaplarını aynı anda sağlayabilen akım-modlu bir filtre önerilmiştir. Önerilen ikinci derece filtre sadece topraklanmış iki kapasitör ve üç tane benzer artı-tipli kontrollü akım taşıyıcı içerir. Böylece önerilen filtre tümleşik devre (IC) olarak kolaylıkla tasarlanabilir. Fakat, önerilen filtre faz-kaydırımı için bir dirençsel eleman uyumuna ihtiyaç duyar. Ayrıca, analog filtrelerdeki işaret sınırlamaları ve kararlılık problemleri tartışılmıştır. En sonunda, simülasyon sonuçları SPICE program kullanılarak teoriyi sağlamlaştırmak için verilmiştir.

Anahtar Sözcükler: Akım-modu, Genel filtreler, CCCIIs.

1. INTRODUCTION

Filters can be divided into two categories, analog and digital filters. In this paper, we deal with only analog filters which can be defined as; they let or prevent the passage of analog signals at some frequencies or frequency ranges. Filters can also be classified as universal or generic filters and multi-function filters. The former give all the standard filter responses such as low-pass, band-pass, high-pass, notch (band-reject) and all-pass responses but the latter can only realize more than one but less than five standard filter responses. There are also four types of filters depending on the type of the signals at their inputs and outputs, current-mode filters, transadmittance-mode filters, transimpedance-mode filters and voltage-mode filters. Filters can also be categorized as single input-single output, single input-multiple outputs, multiple inputs-

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single output and multiple inputs-multiple outputs where the last one, multiple inputs- multiple outputs filter is under consideration. In this study, we are only interested in current-mode filters due to the fact that the use of current-mode processes offers certain important advantages such as greater linearity, wider bandwidth, larger dynamic range, less power consumption, less sensitivity, better accuracy and more simple realization [1-3] than that of the voltage-mode ones.

The floating capacitors [4] and floating resistors [5] can be realized in current technologies easily. However, as explained previously in [6-7] that employing only grounded passive components in integrated circuit (IC) design provides better properties than employing floating ones.

In the literature, quite a great number of circuits based on second-generation current controlled conveyors (CCCIIs) have been devoted to design second-order current-mode analog filters [8-20] but without simultaneously realizing all the universal filter responses, due to their versatility and convenience for practical applications in many areas such as analog communication systems and signal processing. Nonetheless, there is still a need to introduce an analog filter having much more advantages than previously reported ones.

For this purpose, a new current-mode analog filter derived from the filters in [8] is proposed, which simultaneously satisfies all of the following essential properties; **(i)** tuned electronically, **(ii)** uses a canonical number of only grounded capacitors, which is important in IC realization from the parasitic impedance effect reduction/ cancellation points of view, **(iii)** simultaneously realizes all standard filter responses, **(iv)** has high output impedances, **(v)** operates at high frequencies because only resistive components but no capacitors are connected in series to the X terminals of the plus-type CCCII+s (CCCII+s) [21], **(vi)** enjoys low component sensitivities, **(vii)** employs only identical plus-type CCCII+s thus it occupies less area and it can easily be realized in IC technology. However, it needs identical input currents and a single passive component matching constraint for all-pass filter response. On the other hand, signal limitations and stability analysis for the developed filter are given.

2. CIRCUIT DESCRIPTION

Using standard notation, as a versatile active device, a CCCII+ can be represented by the following equations:

$$I_y = 0 \quad (1a)$$

$$I_{zij+} = \alpha_{ij} I_x \quad (1b)$$

$$V_x = V_y + R_x I_x \quad (1c)$$

Electrical symbol of the CCCII+ is given in Fig. 1 in which the directions of the terminal currents of the CCCII+ are simultaneously taken into it.

For the non-ideal current gain α_{ij} in (1b), ($i = 1, 2, 3$) represents the i -th CCCII+ of the filter in Fig. 2 and ($j = 1, 2, 3$) exhibits the j -th Z terminal of the CCCII+. α_{ij} can be expressed as $\alpha_{ij} = 1 + \varepsilon_{ij}$ where ε_{ij} is called as current tracking error and $|\varepsilon_{ij}| \ll 1$.

The intrinsic resistance of the CCCII+, R_x in (1c) can be defined for bipolar realization as in the following [22]:

$$R_x = \frac{V_T}{2I_o} \quad (2)$$

Here, V_T is the thermal voltage and I_o is the bias current of the CCCII+.

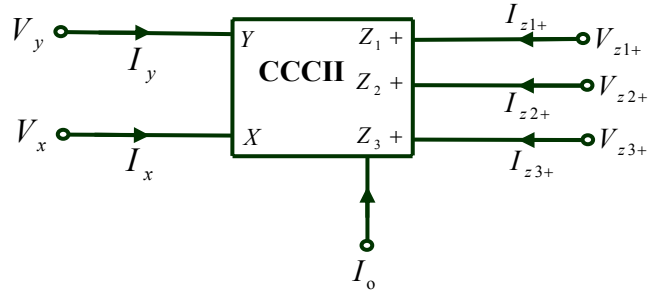


Figure 1. Electrical symbol of the CCCII+

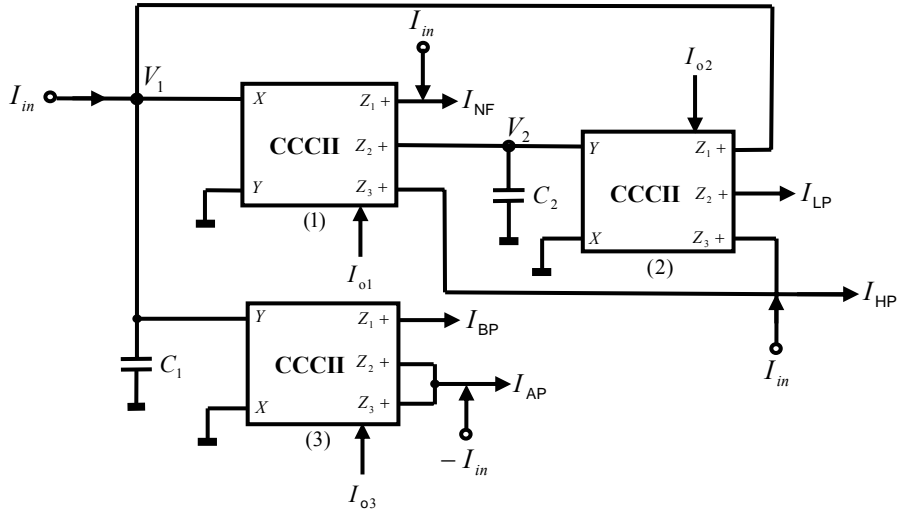


Figure 2. The presented filter for simultaneously realizing all universal responses

Routine analysis of the filter in Fig. 2 with $R_{x3} = R_{x1}$ and considering only non-ideal gain effects, gives the following filter transfer functions (TFs):

$$H_{A1}(s) = \frac{I_{LP}}{I_{in}} = -\frac{\omega_{oi}^2 \alpha_{12} \alpha_{22}}{D_n(s)} \quad (3a)$$

$$H_{A2}(s) = \frac{I_{BP}}{I_{in}} = \frac{s \frac{\omega_{oi}}{Q_i} \alpha_{31}}{D_n(s)} \quad (3b)$$

$$H_{A3}(s) = \frac{I_{HP}}{I_{in}} = \frac{s^2 + s \frac{\omega_{oi}}{Q_i} (1 - \alpha_{13}) + \omega_o^2 \alpha_{12} (\alpha_{21} - \alpha_{23})}{D_n(s)} \quad (3c)$$

$$H_{A4}(s) = \frac{I_{NF}}{I_{in}} = \frac{s^2 + s \frac{\omega_{oi}}{Q_i} (1 - \alpha_{11}) + \omega_o^2 \alpha_{12} \alpha_{21}}{D_n(s)} \quad (3d)$$

$$H_{A5}(s) = \frac{I_{AP}}{I_{in}} = - \frac{s^2 - s \frac{\omega_{oi}}{Q_i} (\alpha_{32} + \alpha_{33} - 1) + \omega_o^2 \alpha_{12} \alpha_{21}}{D_n(s)} \quad (3e)$$

Here, $H_{Ai}(s)$ for $i = 1, 2, 3, 4$ and 5 corresponds to low-pass, band-pass, high-pass, notch and all-pass analog filter TFs, respectively. Apart from this, the effects of non-ideal voltage gains of the CCCII+s of the filter in Fig. 2 are ignored because their pole frequencies are much greater than those of current gains [8].

The denominator $D_n(s)$ in equation (3) of the filter in Fig. 2 are calculated as

$$D_n(s) = s^2 + s \frac{\omega_{oi}}{Q_i} + \omega_{oi}^2 \alpha_{12} \alpha_{21} \quad (4)$$

The ideal parameters, angular resonance frequency ω_{oi} and quality factor Q_i for the equation (4) are computed as

$$\omega_{oi} = \sqrt{\frac{1}{C_1 C_2 R_{x1} R_{x2}}} \quad (5a)$$

$$Q_i = \sqrt{\frac{C_1 R_{x1}}{C_2 R_{x2}}} \quad (5b)$$

If non-ideal gains are taken into account, the parameters in equation (5) respectively turn to

$$\omega_{on} = \omega_{oi} \sqrt{\alpha_{12} \alpha_{21}} \quad (6a)$$

$$Q_n = Q_i \sqrt{\alpha_{12} \alpha_{21}} \quad (6b)$$

Note that all of the active and passive component sensitivities in equations (5) and (6) equal to one half in magnitudes. Further, the current gains of the filter TFs in equations (3a)-(3e) can be set to unity by using extra resistors and electronically tunable current conveyors (ECCII)s with analog control proposed in [23] and digital control developed in [24]. For this reason, the undesired terms of the TFs in equations (3c)-(3e) will vanish along with the gains of all the TFs in equation (3) become unity in magnitudes.

3. SIGNAL LIMITATIONS

If non-ideal gain and parasitic impedance effects are not taken into account, the node voltages of the developed filter are calculated as follows:

$$V_1(s) = \frac{s \frac{\omega_{oi}}{Q_i} R_{x1}}{s^2 + s \frac{\omega_{oi}}{Q_i} + \omega_{oi}^2} I_{in} \quad (7a)$$

$$V_2(s) = -\frac{R_{x2}}{s^2 + s \frac{\omega_{oi}}{Q_i} + \omega_{oi}^2} I_{in} \quad (7b)$$

The magnitudes of the above node voltages in $j\omega$ domain at ω_{oi} are found to be

$$|V_1(j\omega_{oi})| = R_{x1} |I_{in}| \quad (8a)$$

$$|V_2(j\omega_{oi})| = Q_i R_{x1} |I_{in}| \quad (8b)$$

It is observed from equation (8b) that the magnitude of the node voltage V_2 is proportional to Q_i . In other words, V_2 may have distortion for high Q_i values. As a consequence, the values of Q_i , R_{x1} and $|I_{in}|$ in (8) should be chosen small enough.

The node voltages in (8) should be less than their rated voltages which are less than DC symmetrical power supply voltages in magnitudes as formulized below

$$-V_{SS} \leq V_{rk-} \leq V_k(j\omega_{oi}) \leq V_{rk+} \leq V_{DD} \quad (9)$$

where V_k ($k = 1, 2$) and V_{rk} represent the k -th node voltage and rated voltage of the k -th node, respectively. Also, the rated voltages V_{rk-} and V_{rk+} can respectively be defined as negative and positive rated voltages. The voltages V_{DD} and V_{SS} respectively denote the DC positive and negative power supply voltages.

4. STABILITY TEST

We use a single-pole model for each current gain to compute the stability of the introduced filter. Hence, $\alpha(s)$ can be written as

$$\alpha(s) = \frac{\alpha_o}{1 + \tau s} \quad (10)$$

Replacing the model in (10) instead of the current gains, $D(s)$ in (4) turns to

$$D(s) = s^2 C_1 C_2 R_{x1} R_{x2} + s C_2 R_{x2} + \frac{\alpha_{1o}}{1 + \tau_1 s} \cdot \frac{\alpha_{2o}}{1 + \tau_2 s} \quad (11)$$

Therefore, $D(s)=0$ in equation (11) converts to

$$P(s) = \{(1 + \tau_1 s)(1 + \tau_2 s)\} \left\{ s^2 + s \frac{\omega_o}{Q} \right\} + \omega_o^2 \alpha_{1o} \alpha_{2o} = 0 \quad (12a)$$

Similarly, $P(s)$ in (12a) can be rewritten as

$$P(s) = a_4 s^4 + a_3 s^3 + a_2 s^2 + a_1 s + a_0 = 0 \quad (12b)$$

We assume identical conveyors such that $\tau_1 = \tau_2 = \tau$ and $\alpha_{10} = \alpha_{20} = \alpha_0$. As a result, Routh's stability criterion can be applied to $P(s)$ in (12b) as follows:

$$\begin{cases} s^4 & a_4 & a_2 & a_0 \\ s^3 & a_3 & a_1 & \\ s^2 & b & a_0 & \\ s^1 & c & & \\ s^0 & a_0 & & \end{cases} \quad (13)$$

Here,

$$a_4 = \tau^2 \quad (14a)$$

$$a_3 = 2\tau + \frac{\omega_0}{Q} \tau^2 \quad (14b)$$

$$a_2 = 1 + \frac{\omega_0}{Q} 2\tau \quad (14c)$$

$$a_1 = \frac{\omega_0}{Q} \quad (14d)$$

$$a_0 = \omega_0^2 \alpha_0^2 \quad (14e)$$

It is known that the terms b and c in (13) must be positive real numbers for stable filter responses. Thus,

$$b = \frac{a_2 a_3 - a_1 a_4}{a_3} > 0 \quad (15a)$$

$$c = \frac{a_1 b - a_0 a_3}{b} > 0 \quad (15b)$$

Note that ω_0 and Q values put restriction on possible element selection C_1 , C_2 , R_{x1} , R_{x2} and R_{x3} .

5. SIMULATION RESULTS

The CCCII+s of the proposed four inputs five outputs filter in Fig. 2 are simulated using the schematic implementation exhibited in Fig. 3 with $\pm 2.5V$ DC power supply voltages. The two extra Z^+ terminals are added to each of the CCCII+ in Fig. 3. The PNP and NPN transistors employed in the CCCII+ implementation are simulated using the parameters in [25]. The following settings are selected with resonance frequency $f_0 = 318.3$ kHz and $Q = 1$: capacitors $C_1 = C_2 = 1nF$ are taken and bias currents $I_{o1} = I_{o2} = I_{o3} = 26\mu A$ are chosen to obtain the intrinsic resistors $R_{x1} = R_{x2} = R_{x3} = 500\Omega$.

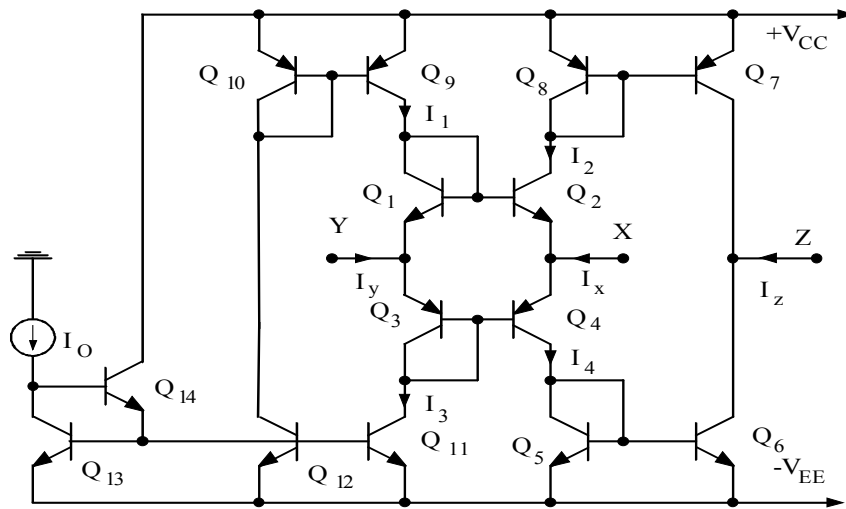


Figure 3. Internal structure of the CCCII+ with BJT transistors

Both ideal and simulated low-pass, high-pass and notch responses are exhibited in Fig. 4. In addition, phase and magnitude responses of the band-pass and all-pass filter TFs are given in Figs. 5 and 6, respectively.

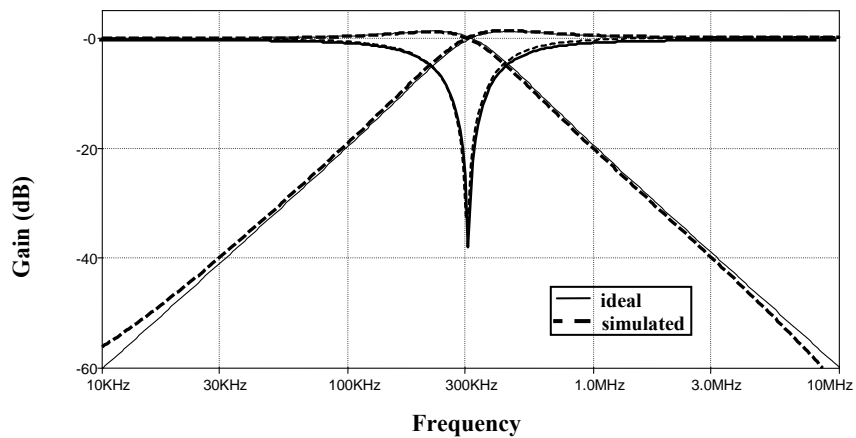


Figure 4. Ideal and simulated low-pass, high-pass and notch responses of the filter in Fig. 2

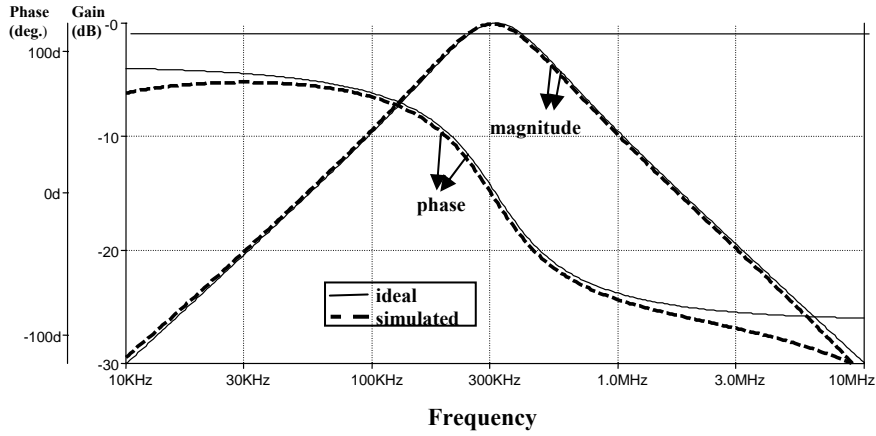


Figure 5. Phase and magnitude responses of the band-pass filter in Fig. 2

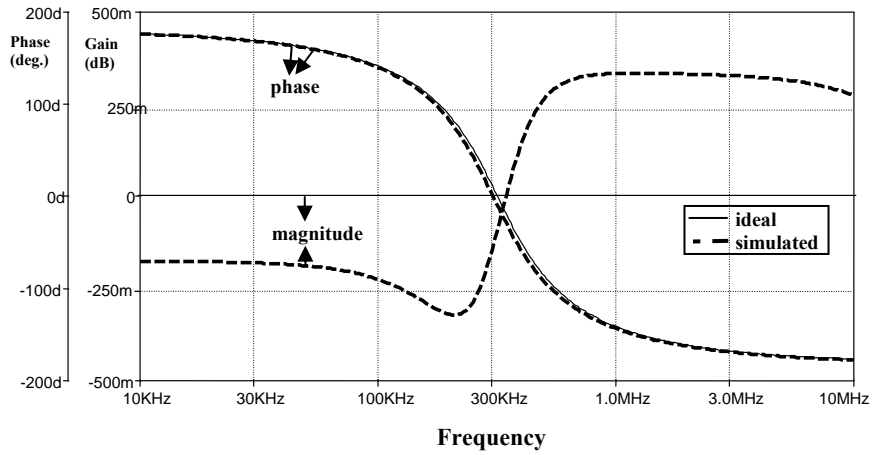


Figure 6. Phase and magnitude responses of the all-pass filter in Fig. 2

It is observed from Figs. 4-6 that the ideal and simulated responses obtained using SPICE program are obviously in good agreement but there are some differences between them. These discrepancies mainly stem from non-ideal gain and parasitic impedance effects of the CCCII+s as well as signal limitations of the developed filter as explained in the text.

Using Routh's stability criterion, the $P(s) = 0$ in (12) is tested with respect to the parameters α_0 , k and Q . Using MATLAB program, the stable and unstable regions of the developed filter with respect to two combinations of three parameters k , α_0 and Q are depicted in Figs. 7-9 in which $k = \omega_{c1}/\omega_0 = 1/(\tau\omega_0)$, $\alpha_{10} = \alpha_{20} = \alpha_0$, $\omega_{c1} = \omega_{c2} = \omega_c$ or $\tau_1 = \tau_2 = \tau$ and $\omega_0 = 5 \times 10^5$ rad/s are chosen.

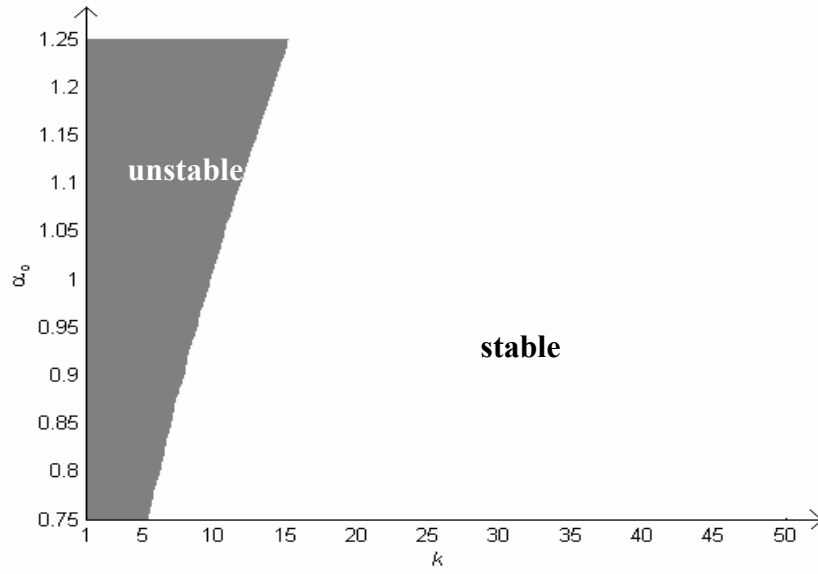


Figure 7. Stable region with respect to variations of k and α_0 where $Q=5$

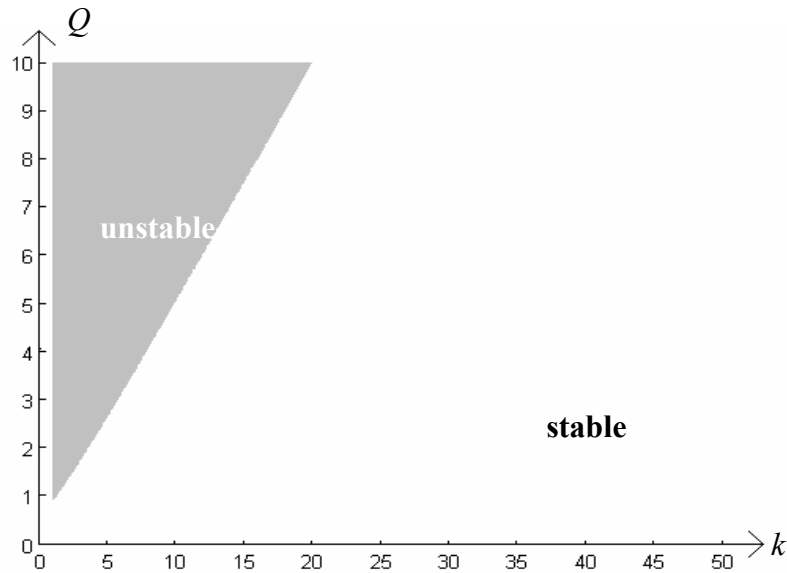


Figure 8. Stable region with respect to variations of k and Q where $\alpha_0=1$

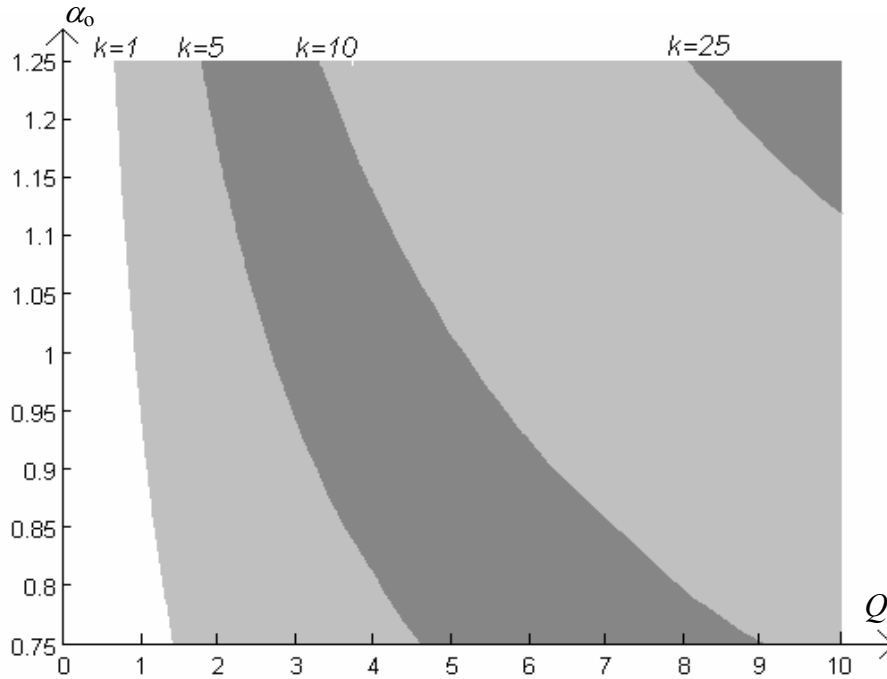


Figure 9. Stable region with respect to variations of Q and α_0 where $k=1, 5, 10$ and 25

From Fig. 7, assuming a straight line between the stable and unstable regions, a linear relation could be given as $k > 20\alpha_0 - 10$ in order to satisfy stability.

From Fig. 8, the stable and unstable regions are nearly separated with a line and the linear relation could be written as $k > 2Q$ to provide stability.

From Fig. 9, for the given intervals, the relationship among k , Q and α_0 that define the stable regions seems to be highly non-linear. For simplicity, by taking a linear relationship, it may be approximated as $k > 4\alpha_0 Q - 2Q$ for stability.

It is observed from Figs.7-9 that the larger k and the smaller α_0 and Q , the more stable filter responses can be obtained.

6. CONCLUSION

The proposed filter offers some important advantages as outlined in the text. Further, the introduced filter scheme exhibits the property of cascability and insensitivity making it much more attractive. In conclusion, the major purpose of this study is to propose a configuration, based on only CCCII+s and grounded capacitors, for the design of electronically tunable universal current-mode filter. In addition to this, signal limitation and stability test based on proposed filter are given in this study to take attention of the analog circuit designers to these restrictions.

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