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Current-Mode KHN-Equivalent Biquad Filter using Current-Mirror

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Abstract
Design of analogue integrated circuits can be done both in voltage-mode and current-mode form of signal processing. State-of-the-art analogue integrated circuits design is receiving tremendous boost from the development and application of current-mode processing, which is rapidly dominating traditional voltage-mode designs. Several techniques and circuits are available in literature for the design of various voltage-mode and current-mode signal processing circuits that are suitable for VLSI implementation both in BiCMOS and CMOS technology. Current-mode signal processing is a very attractive approach because of the ease of implementing operations such as addition/subtraction, multiplication by a constant, and the potential to operate at higher signal bandwidths than their voltage-mode analogues. The main object of this paper is to present design of current-mode continuous-time high-frequency biquadratic filters which will be suitable for implementation in VLSI technology.

Keywords: Continuous-time filters, biquadratic filters, complementary current-mirrors, Kerwin-Heulsmann-Newcomb (KHN) Biquad, current-mode signal processing.

Introduction
As compared to voltage-mode form of signal processing current-mode technique has been received considerable attention, because it offers several advantages such as: higher slew rates, lower power consumption, higher frequency range of operation, better accuracy, and improved linearity. Interest in current-mode (CM) [1,2] filters has been growing due to the fact that current-mode devices have wider dynamic range, improved linearity, and extended bandwidth as compared to voltage-mode devices. For
operation in MHz frequency range [8,15] the main challenges in the design of analog filters are: reliable high-frequency performance, automatic on-chip tuning against fabrication tolerances and changing operating conditions. So, for high frequency applications the output should be obtained in the form of current given by:

$I_{out} = g_m V_{in}$(1)

Where, g_m is the transconductance parameter provided by the active device. For application as continuous-time filters, transconductance should satisfy the following properties: circuit must be simple, linear and have a wide frequency response, must have large input and output impedances to prevent undesirable interactions and simplify circuit design, they should preferably work with low-voltage power supplies to conserve power and to be compatible with the prevalent digital technologies on the same chip. Their transconductance parameter depends on some dc bias voltage or current to facilitate electronic tuning against environmental or processing variations. Depending on the technology chosen, the frequency range of transconductance circuits [8] extends to > 50 MHz (CMOS), > 500 MHz (bipolar) or even to > 1GHz (GaAs) so that the design of high-frequency continuous-time telecommunication circuits becomes feasible. As the trans-conductance and capacitors are the only components required for realizing a filter, trans-conductance-C [21] filters can readily be implemented in fully integrated form [6], for a desired technology. In many useful active simulations of filters, we may even insist that all the trans-conductor should be identical and all the capacitors are grounded for specially simple IC layout and processing, and implementation of integrated analog filters based on analog gate arraysappears a distinct possibility. In addition, the wider useful bandwidth of trans-conductance coupled with the reduced effects of circuit and device parasitic on filter performance result in far higher operating frequencies at which the circuits can function.

Biquadratic Filters

A filter is a two-port network that shapes the spectrum of the input signal in order to obtain an output signal with the desired frequency content. Thus, a filter has pass bands where the frequency components are transmitted to the output and stop bands where they are rejected.

Current-Mode KHN-equivalent Biquad

It can be constructed by cascading two lossless integrators [9,11,13] to obtain all the five desired filter transfer functions namely:

i) Low-Pass Response.
ii) High-Pass Response.
iii) Band-Pass Response
iv) Notch Response
v) All-Pass Response.

Figure-1 gives the block diagram representation of Current-mode KHN-equivalent biquad followed by the necessary mathematical steps for deriving the current transfer functions. Figure-2 represents the circuit realization of the current-mode KHN-equivalent biquad using CMOS complementary current-mirror pairs.

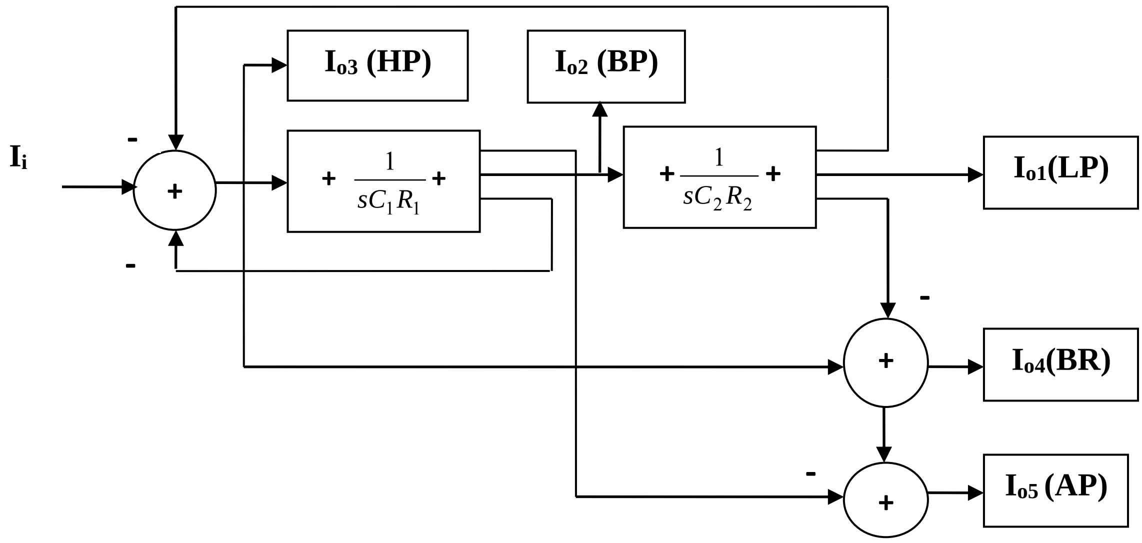

Figure1: Current-Mode KHN-equivalent biquad

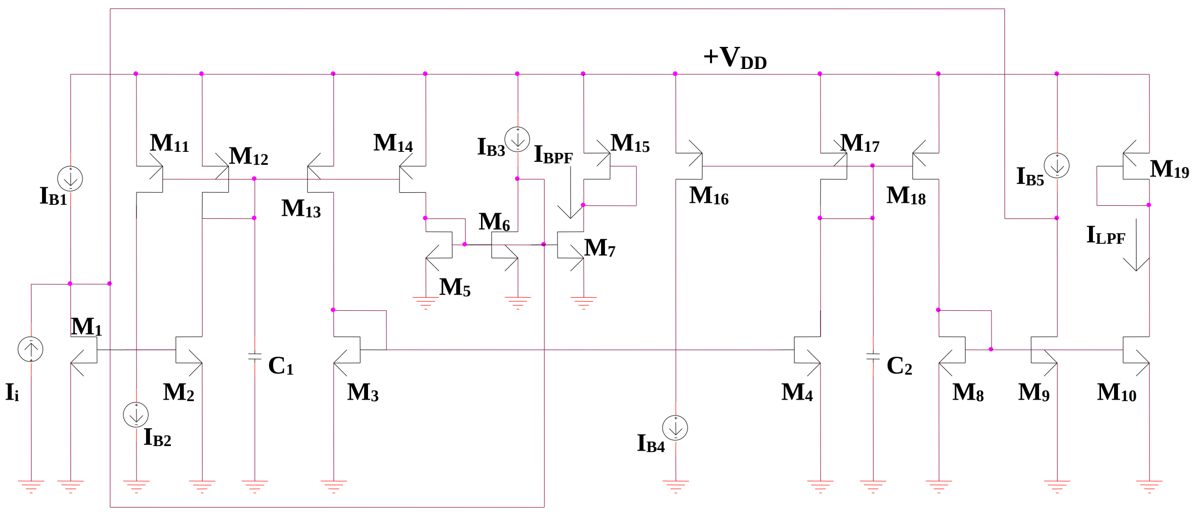

Figure2: Circuit realization of Current-Mode KHN-equivalent biquad

Mathematical Analysis

From Figure-1, we can write the following three equations:

$$I_{o3} = I_i - I_{o2} - I_{o1} \quad (2)$$

$$I_{o1} = \frac{I_{o2}}{sC_2R_2} \quad (3)$$

$$I_{o2} = \frac{I_{o3}}{sC_1R_1} \quad (4)$$

From equation-(3) and (4) we have,

$$I_{o1} = \frac{I_{o3}}{s^2C_1C_2R_1R_2} \quad (5)$$

Now substituting the value of I_{o1} and I_{o2} from equation (4) and (2.4) in equation-(2.1) we obtain the expression for the current I_{o3} in terms of input current I_i as:

$$\frac{I_{03}}{I_i} = \frac{s^2}{s^2 + s\left(\frac{1}{C_1R_1}\right) + \left(\frac{1}{C_1C_2R_1R_2}\right)} \quad (6)$$

Thus, we can say that the equation (6) represents a high-pass response. Similarly, when the value of I_{o3} is substituted from equation (6) in equation (4) and (5) respectively we obtain the expression for the currents I_{o1} and I_{o2} in terms of the input current I_i as follows:

$$\frac{I_{o2}}{I_i} = \frac{s\left(\frac{1}{C_1R_1}\right)}{s^2 + s\left(\frac{1}{C_1R_1}\right) + \left(\frac{1}{C_1C_2R_1R_2}\right)} \quad (7)$$

$$\frac{I_{o1}}{I_i} = \frac{\left(\frac{1}{C_1C_2R_1R_2}\right)}{s^2 + s\left(\frac{1}{C_1R_1}\right) + \left(\frac{1}{C_1C_2R_1R_2}\right)} \quad (8)$$

Thus, we can say that the equation (7) and (8) respectively represents a band-pass response and a low-pass response. Similarly, the currents I_{o4} can be written as:

$$I_{o4} = I_{o3} + I_{o1} \quad (9)$$

Now, substituting the value of the currents I_{o3} and I_{o1} from equation (6) and (8) respectively in equation (9) we obtain the expression for the current I_{o4} in terms of the input current I_i.

$$\frac{I_{o4}}{I_i} = \frac{s^2 + \left(\frac{1}{C_1C_2R_1R_2}\right)}{s^2 + s\left(\frac{1}{C_1R_1}\right) + \left(\frac{1}{C_1C_2R_1R_2}\right)} \quad (10)$$

Thus, we can say that the equation (10) represents a notch response. Similarly, for the current I_{o5} we have the following expression:

$$I_{o5} = I_{o4} - I_{o2} \quad (11)$$

Now, substituting the value of the currents I_{o4} and I_{o2} from equation (10) and (6) respectively in equation (11) we obtain the expression for the current I_{o5} in terms of the input current I_i.

$$\frac{I_{o5}}{I_i} = \frac{s^2 - s\left(\frac{1}{C_1R_1}\right) + \left(\frac{1}{C_1C_2R_1R_2}\right)}{s^2 + s\left(\frac{1}{C_1R_1}\right) + \left(\frac{1}{C_1C_2R_1R_2}\right)} \quad (12)$$

Thus, we can say that the equation (12) represents an all-pass response. The expressions for the filter cut-off frequency (ω_o) and the quality factor (Q) can be obtained from the above current transfer function and they are given by:

$$\omega_o = \frac{1}{\sqrt{C_1C_2R_1R_2}} \quad (13)$$

$$Q = \sqrt{\frac{C_1R_1}{C_2R_2}} \quad (14)$$

where,

$$R_1 = \frac{1}{g_{m1}} \text{ and } R_2 = \frac{1}{g_{m2}} \quad (15)$$

and g_{m1}, g_{m2} are the trans-conductance of the diode connected transistors Q_{12}, Q_{17} respectively. Also, I_{B1} and I_{B2} are the DC bias currents shown as I_{B2} and I_{B4} in Figure-2.

But the trans-conductance g_{m1} and g_{m2} are directly related to the square root of the bias currents I_{B2} and I_{B4} and are given as:

$$g_{m1} = \sqrt{2\mu_n C_{ox}\left(\frac{W}{L}\right)_{12} I_{B2}} \quad (16)$$

$$g_{m2} = \sqrt{2\mu_n C_{ox}\left(\frac{W}{L}\right)_{17} I_{B4}} \quad (17)$$

Therefore, the expression for filter parameters cut-off frequency (ω_o) and the quality factor (Q) becomes:

$$\omega_o = \sqrt{\frac{I_{B2}I_{B4}}{C_1C_2}} \quad (18)$$

$$Q = \sqrt{\frac{C_1I_{B4}}{C_2I_{B2}}} \quad (19)$$

If $C_1 = C_2 = C$, then the equation (18) and (19) becomes:

$$\omega_o = \frac{1}{C}\sqrt{I_{B2} I_{B4}} \quad (20)$$

$$Q = \sqrt{\frac{I_{B4}}{I_{B2}}} \quad (21)$$

Also, if $I_{B2} = I_{B4} = I_B$ then the equation (20) and (21) becomes:

$$\omega_o = \frac{I_B}{C} \quad (22)$$

$$Q = 1 \quad (23)$$

Simulation Results

The workability of the proposed circuit shown in Figure-2 was tested and verified in SPICE using 0.5µm CMOS process parameters provided by MOSIS (AGILENT) as listed in Table-1.

Table1: Cmosprocess parameters

TRANSISTOR	PROCESS PARAMETERS
Nmos	LEVEL=3 UO=460.5 TOX=1.0E-8 TPG=1 VTO=0.62 JS=1.08E-6 XJ=0.15U RS=417 RSH=2.73 LD=0.04U VMAX=130E3 NSUB=1.17E17 PB=0.761 ETA=0.00 THETA=0.129 PHI=0.905
pMOS	LEVEL=3 UO=100 TOX=1.0E-8 TPG=1 VTO=0.58 JS=0.38E-6 XJ=0.10U RS=886 RSH=1.81 LD=0.03U VMAX=113E3 NSUB=2.08E17 PB=0.911 ETA=0.00 THETA=0.120 PHI=0.905 GAMMA=0.76 KAPPA=2 CJ=85E-5 MJ=0.429 CJSW=4.67E-10 MJSW=0.631 CGSO=1.38E-10 CGDO=1.38E-10 CGBO=3.45E-10 KF=1.08E-29 AF=1 WD=0.14U DELTA=0.81 NFS=0.52E11

For the circuit shown in Figure-2 the ac analysis were carried out with the value of dc bias current $I_{B1} = I_{B3} = I_{B5} = 24\mu A$, $I_{B2} = I_{B4} = 15\mu A$, $C_1 = 0.01pF$, $C_2 = 0.1pF$, $(W/L)_P$ ratio = 1µm/1µm, $(W/L)_N$ ratio = 1µm/1µm and supply voltage $V_{DD} = 5V$. The simulated value of the cut-off frequencies for low-pass, high-pass and band-pass response is found to be $(f_O)_{LPF}$ = 51.67MHz, $(f_O)_{HPF}$ = 95.404MHz and $(f_O)_{BPF}$ = 88.444MHz respectively. The bandwidth of the band-pass filter is found to be 186.894MHz. These simulated results are very well in agreement with the calculated theoretical values which were respectively found to be of $(f_O)_{LPF}$ = 52MHz, $(f_O)_{HPF}$ = 95MHz, $(f_O)_{BPF}$ = 88MHz and BW = 185MHz. The SPICE simulation result for the proposed circuit is shown in Figure-3.

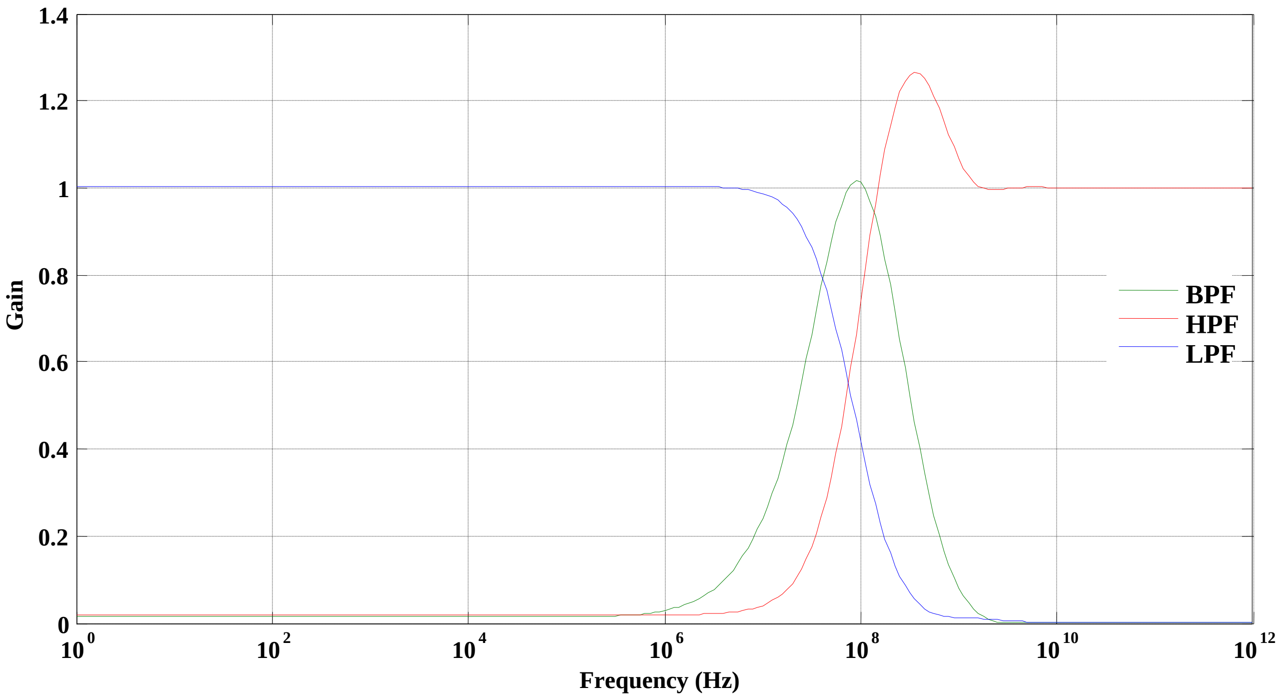

Figure3: Simulated response of Current-Mode KHN-equivalent biquad

Figure-4 and Figure-5 respectively represents the variation in the cut-off frequency and the gain of the band-pass and the low-pass responses of the current-mode KHN equivalent biquad circuit shown in Figure-2.

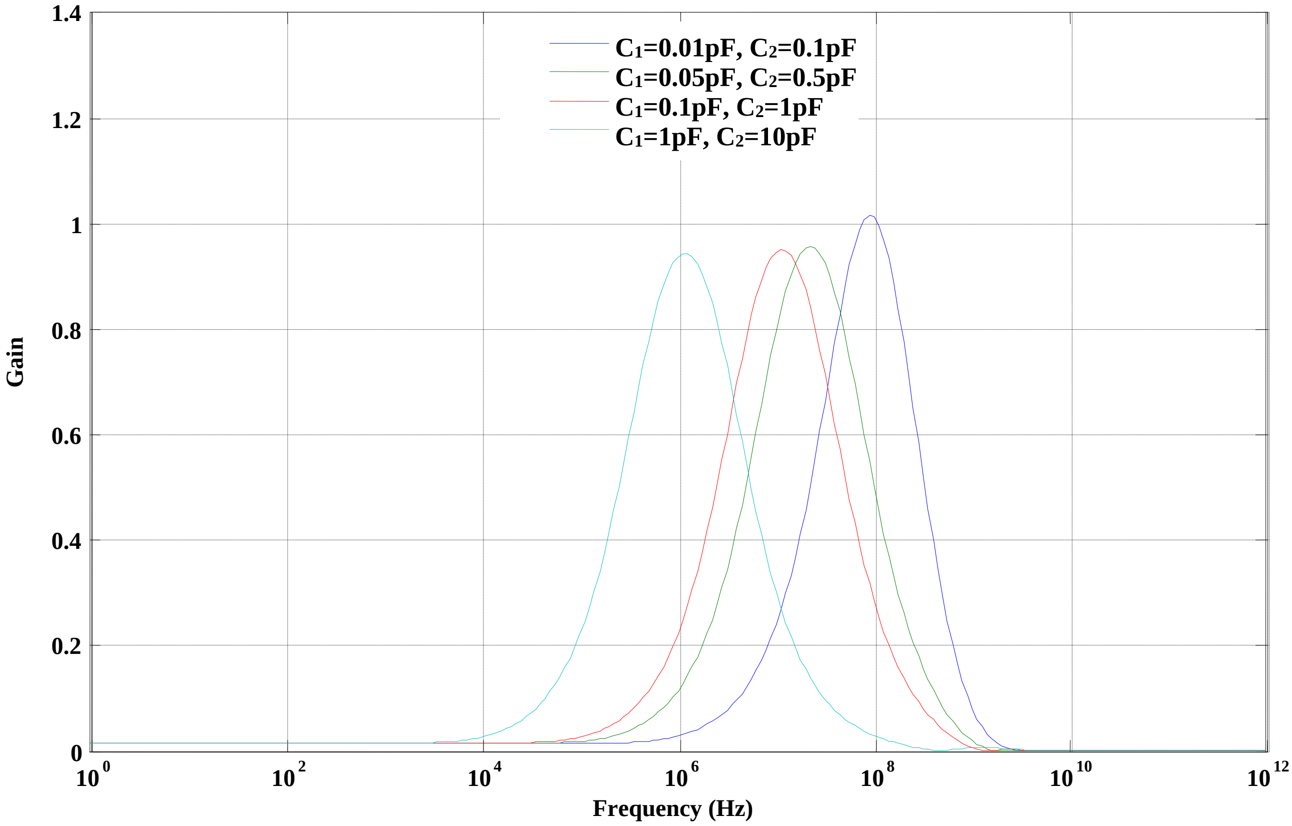

Figure4: Variations in cut-off frequency of band-pass response with capacitance value for Current-Mode KHN-equivalent biquad

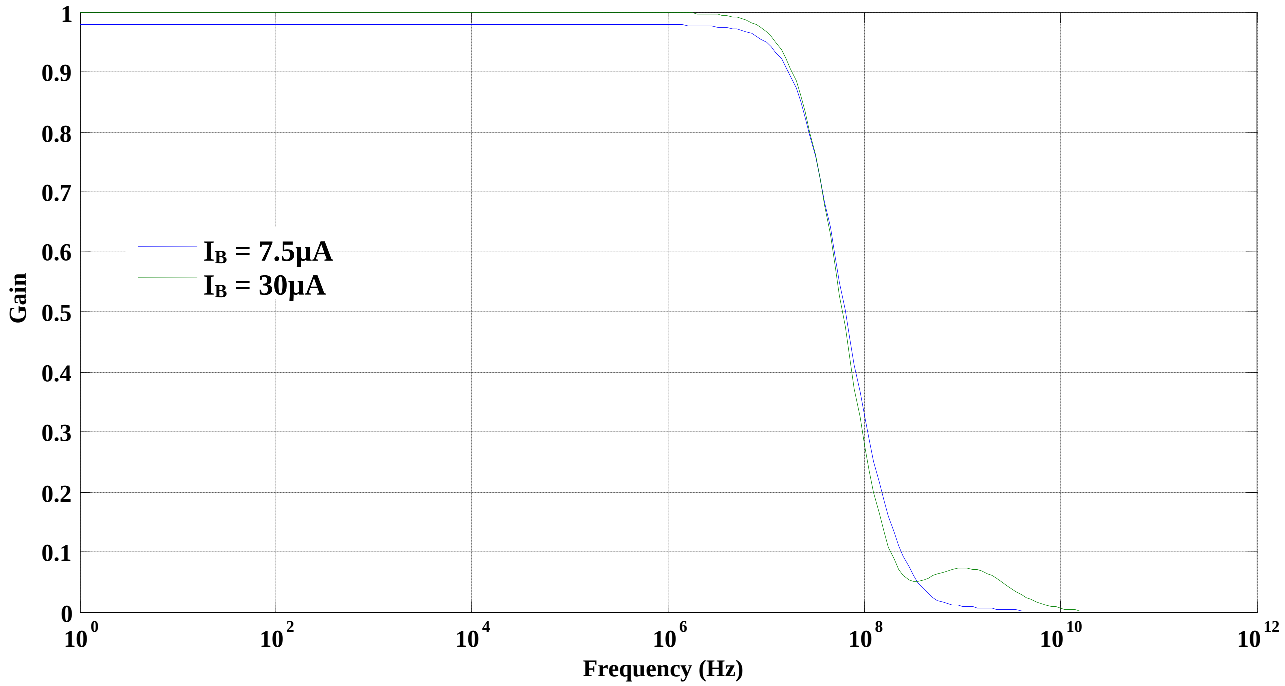

Figure 5: Variation in gain frequency of low-pass response with bias current for Current-Mode KHN-equivalent biquad

Conclusions

In the given paper high frequency current-mode KHN-equivalent biquad filter has been presented which is quite suitable for operation at high frequencies till 100 MHz and these filters can operate at a voltage as low as 5V. Also, it has been verified that variations in the value of either the capacitor or the bias current improves the gain as well as the operating frequency of the filter. The frequency of this filter can be easily and widely controlled by a single DC bias current and thus provides good tunability. All the circuits were tested using SPICE and the verified results confirms the theoretical values.

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