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A 3rd-order Low-Pass Filter with High Linearity and Tunable In-band Attenuation for WiMAX /LTE Receiver

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ABSTRACT

In this paper, a Gm-C Low-pass filter with high linearity and tunable in-band attenuation for WiMAX/LTE receiver is presented. A linear Gm-cell is proposed and used as a key building block in design of the filter. The linear Gm-cell employs a circuit technique to reduce the nonlinearity of the source follower (SF) stage resulting from unmatched signal swings at the gate and source terminals of the input transistor. The Gm-cell was used to build a biquad bloc, which was utilized to construct a 3rd-order Butterworth low pass filter. The proposed filter consists of a buffer and a biquad stage to form the 3rd-order transfer function. The simulation results of the filter for bandwidth of 10MHz show that the IIP3 of the filter was equal to 29.5 dBm. The minimum in-band noise density was 22.6 nV/$\sqrt{}$Hz and power consumption was 10.8 mW. The supply voltage of the filter was 1 V.

Keywords: Gm-C filter, WiMAX/LTE Receiver, Linear Gm, Source-Follower.

1. INTRODUCTION

WiMAX and LTE standards are the core of 4G wireless communications, which provides higher data rate for the end users. The higher data rate will result in better video communication, higher download and upload speed for internet access and other advantages that will benefit the users with endless possibilities.

Every transceiver that is introduced to the market for a new generation of communication standard, is supposed to be compatible and support the old generations as well. This leads to demand in extra hardware, area and power consumption. One way to reduce the area and power overhead is to utilize the multi-standard and reconfigurable hardware inside the transceiver. This feature is demanded in the whole transceiver chain including the baseband filter of the receiver.

The base-band filter is a vital block in receiver section, which has a very challenging specification for realization. In wireless receiver design for new generation of communication standards such as 4G, implementation of a baseband low pass filter is recommended. The implementation is able to cover various standards based on geolocation and different revisions of a specific standard. This entails the investigation of various structures from linearity, noise and in-bond attenuation viewpoint to choose the optimum one (D'Amico 2006).

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Currently, most of the integrated continuous-time filters are generally implemented using Opamp-RC or G_m-C methods. Opamp-RC method has better linearity, yet more chip area and power consumption. On the other hand, the G_m-C method has low power consumption and small area, but lower linearity compared to Opamp-RC method (D'Amico 2005).

Integrated base band filters, which are used in wireless receivers and portable devices should be realized with low power circuits. Moreover, noise, linearity and in-band attenuation are other crucial factors that should be considered in the realization. Thus, structures with the ability of achieving low power consumption are preffered. The G_m-C method is favorable over the Opamp-RC method from this point of view (De Matteis 2007).

In this paper, a linear G_m-cell based on a modified SF stage is introduced. The linear G_m-cell used to improve the linearity and reduce the power consumption of the final G_m-C filter. By using the proposed linear G_m-cell in the realization of the filter, the linearity of the filter is improved without extra power consumption, which is not possible in traditional design methods (Giannini 2006).

The bandwidth tuning in integrated low-pass filters is performed by switching in or out the capacitors and/or resistors, which can provide a wide frequency tuning range. In order to cover several frequency bandwidths that exist in variant versions of the WiMAX/LTE standards, two distinct coarse and fine schemes for tuning of the filter bandwidth are employed in this paper (Palarti 2013).

The rest of this paper is structured as follows; the Section 2 of the paper is dedicated to explanation of the proposed linear G_m-cell. In Section 3, structure and circuit details of the proposed 3rd-order filter is presented. Section 4 describes the bandwidth tuning procedures. The simulation results are presented in Section 5 and finally Section 6 concludes the paper.

2. LINEAR G_m-CELL

In portable devices, circuit blocks which have low power consumption and high linearity are in high demand. In order to address this fundamental approach for the baseband filter, a linear G_m-cell is introduced that is targeting the low power consumption and high linearity features. This linear Gm-cell is later used inside a biquad that is the key building block of the proposed filter. Using a linear G_m-cell inside the biquad, a G_m-C filter is constructed based on this biquad. This will lead to a linear filter having low power consumption.

The SF stage can achieve high linearity with low effective voltage, which is very critical in advance processes with low supply voltage. In this way, the SF stage is capable to significantly reduce the power consumption, which is more pronounced in high order filters. The SF circuit does not have non-dominant pole, which could emerge at higher frequencies and degrade the excess phase of the filter. The SF stage can drive resistive loads without affecting the filter performance in terms of the linearity and accuracy of the filter transfer function (Razavi 1998).

The above-mentioned points make the SF stage an interesting choice for a G_m-Cell inside a low-pass filter. Although, the SF stage has good linearity for typical applications, but not enough to be used as a G_m-cell inside a low pass filter for baseband section of a wireless receiver. The main source of nonlinearity in SF is due to changes in bias current which caused by input signal variations. The unmatched signal swings at the source and gate terminals of the transistor is the reason for bias current variations. This unmatched swing at the gate and source terminals changes the gate-source voltage of the transistor. As a result, bias current is altered.

The transconductance (g_m) of a MOSFET transistor is directly related to bias current (I_D). Therefore, variations in bias current will alter the g_m value. The transfer function of the SF circuit in Figure 1 is given in (1),

$$\frac{V_o}{V_{in}} = \frac{r_{ds}||R_L}{\frac{1}{g_m}+(r_{ds}||R_L)} \tag{1}$$

which shows that g_m variations caused by bias current will change the voltage gain of the SF circuit which is transformed into nonlinear behavior of the stage. In Equation (1), the r_{ds} is the output resistance of the transistor and R_L is the load resistor. The change of r_{ds} value due to V_{ds} variation is another source of nonlinearity in a source follower stage (De Matteis 2009).

In this paper, a technique is introduced to match the voltage swings at the gate and source terminals of the SF transistor, which can keep the gate-source voltage and bias current value constant. For this purpose, a linearized (heavily degenerated) common source (CS) stage is used in parallel with SF circuit. This CS stage introduces extra swing to the source terminal, which can compensate for smaller variations in comparison with the gate voltage. The proposed circuit is shown in Figure 2; the M_1 transistor is the main device in the SF stage and target of the linearization. The M_2 transistor degenerated by I_2 current source form the CS stage, which produces the compensating signal at the source of the M_1. Having differential input signal, gate of the CS transistor M_2 is connected to other side of the differential input (V_{in-}) and the correct voltage polarity is created at the source of the M_1. By degenerating the CS stage using I_2 current source, its linearity will not be the limiting factor for the entire stage.

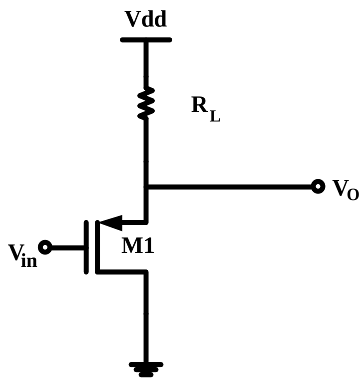

Figure 1. Source follower circuit.

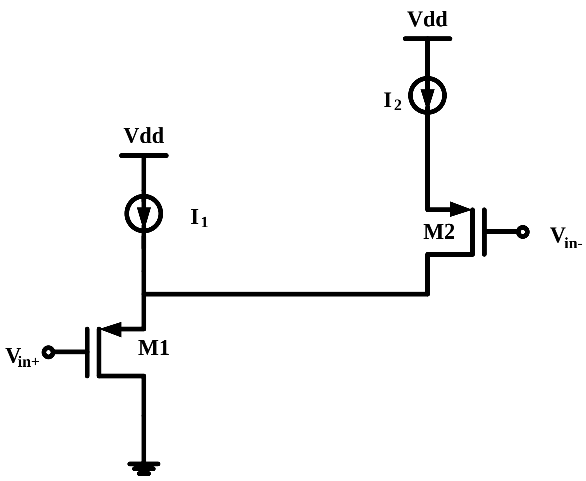

Figure 2. Proposed technique to improve the linearity of the SF circuit.

The complete schematic of proposed linear G_m-cell is shown in Figure 3. In order to reduce distortion and nonlinear effects, the proposed circuit is realized in differential input and output form. The differential circuit is also resistant against common mode signals variations. The coupling capacitors C_2 and C_3 are used to isolate the DC operating point of the M_3 and M_4 transistors, which will otherwise go to triode region.

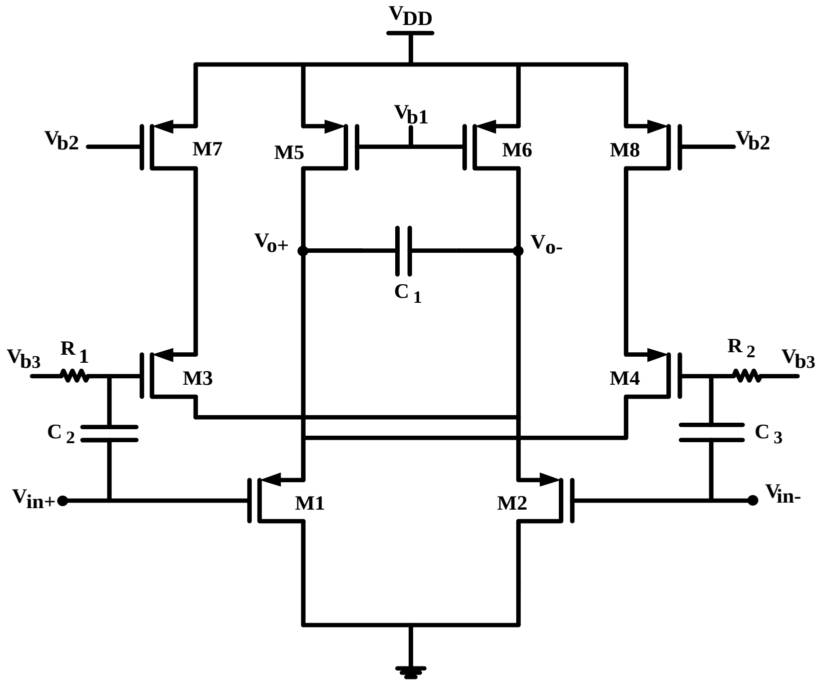

Figure 3. The circuit schematic of the proposed linear G_m-cell.

3. THE Gm-C FILTER

In this section, the design of a 3rd-order low pass Butterworth filter for WiMAX/LTE receiver is presented. The 3rd-order filter is implemented using a buffer (g_m) stage in series with the biquad block. The block diagram of this G_m-C filter is shown in Figure 4. The biquad block is shown inside a rectangular box. The biquad block consists of three G_m-cells, which are designed based on the linear G_m-cell that was introduced in the previous section. The low-pass and high-pass outputs are designated by LP and HP respectively in the figure.

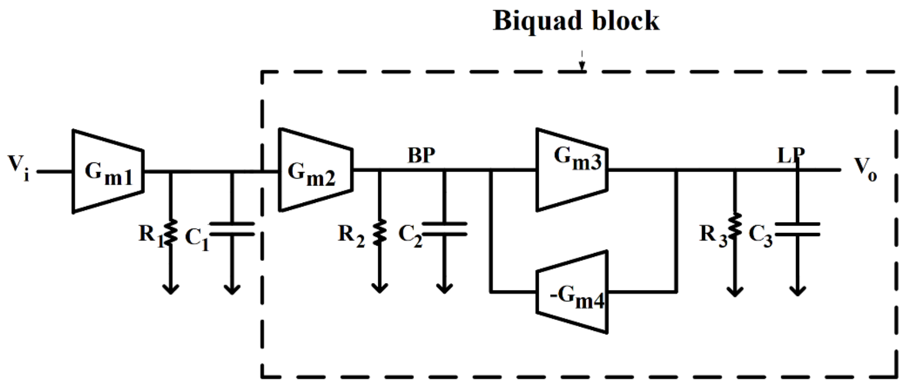

Figure 4. Block diagram of the 3rd-order G_m-C filter.

The circuit schematic of a differential SF that is used as G_{m1} stage in the proposed filter structure, as shown in Figure 5. This circuit introduces necessary third pole to the filter transfer function. The Linear G_m-cell shown in Figure 3. is used to realize the G_{m2} and G_{m4} blocks. However, G_{m3} is designed using the circuit shown in Figure 6. This circuit is in contrary to the original linear G_m-cell, is using the NMOS transistors at the input instead of the PMOS. The NMOS transistors at the input enable the circuit to be directly (without using coupling capacitors) connected to other G_m stages inside the biquad. The coupling capacitors attenuate the signal content close to DC frequencies and change the filter transfer function. In the G_m stage of Figure 6, the coupling capacitors at the input for M_3 and M_4 are no longer necessary. The circuit operating principles are similar to the original circuit with PMOS inputs.

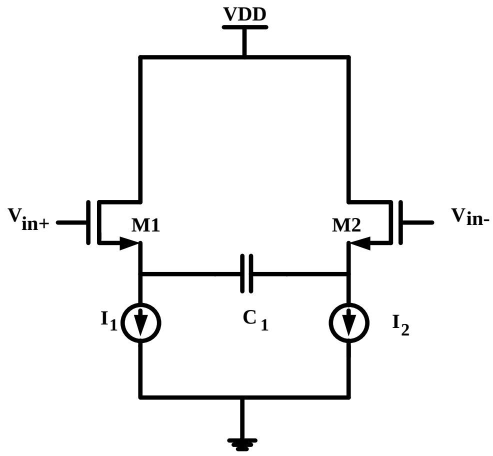

Figure 5. The circuit schematic of the differential source-follower.

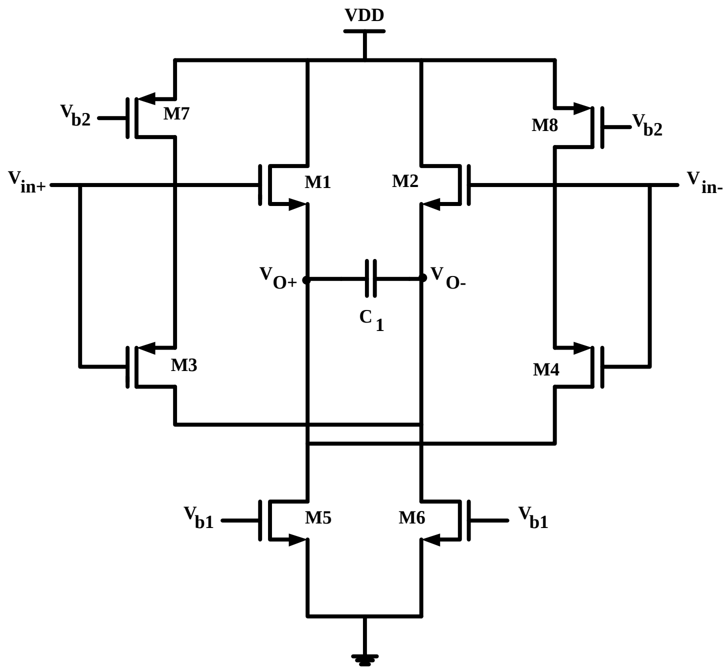

Figure 6. The circuit schematic of the proposed linear G_m with NMOS input.

The small signal circuit model of the biquad block in Figure 7(a) is shown in Figure 7(b). In this model, R_2 resistor is equivalent to the output resistance of G_{m2} in parallel with output resistance of the G_{m4} and R_3 is the output resistance of the G_{m3}. Assuming that transconductance cells do not have non-dominant poles, the bandwidth (ω_0) and the quality factor (Q) of the filter can be written as in Equation 2 and Equation 3:

$$\omega_0 = \sqrt{\frac{G_{m2}R_2 + G_{m3}G_{m4}R_2R_3}{R_2R_3C_2C_3}} \simeq \sqrt{\frac{G_{m3}G_{m4}}{C_2C_3}} \tag{2}$$

$$Q = \frac{\omega_0}{\frac{1}{R_2C_2}+\frac{1}{R_3C_3}+\frac{G_{m2}}{C_2}} \simeq \sqrt{\frac{C_3}{C_2}\frac{G_{m3}G_{m4}}{G_{m2}^2}} \tag{3}$$

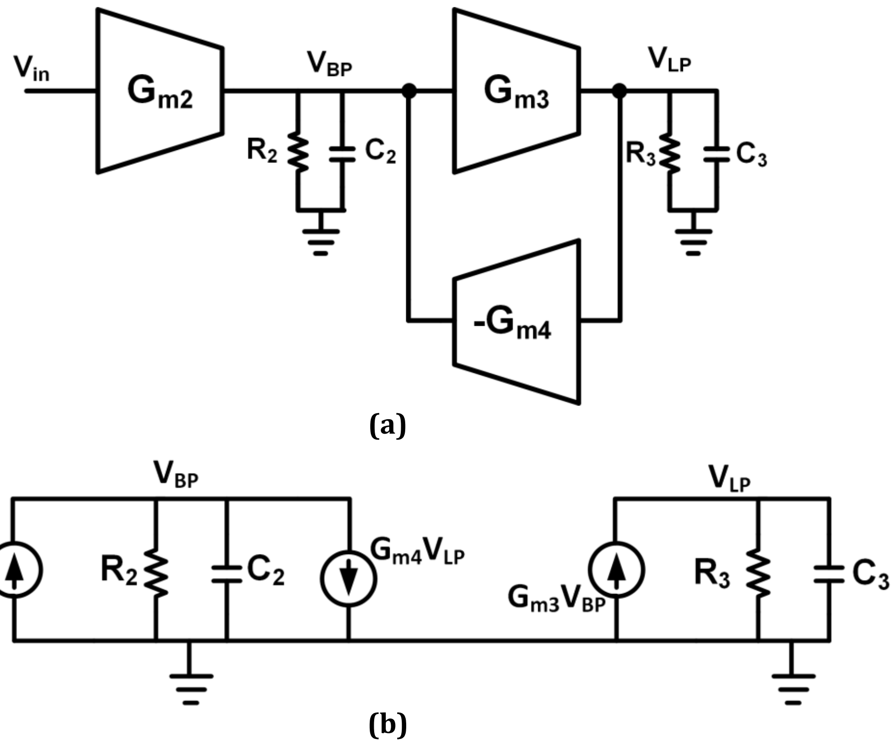

Figure 7. The small signal circuit model of the biquad block.

According to Equation 2 and Equation 3, filter bandwidth can be tuned by changing the C_2 and C_3 capacitors. In order to avoid the changes in Q value, the ratio of these capacitors should be kept constant. If the G_m values are equal, it is possible to change the bandwidth by adjusting the capacitor values while keeping the Q value constant.

Using transconductors which have non-dominant poles close to the dominant one can affect the stability, frequency response, quality factor and excess phase of the filter. To alleviate the problems associated by non-dominant poles in G_m-C or Opamp-RC filters, the transconductor or Opamp stages inside the filter must be designed as a single-pole simple stage (Oskooei, 2011).

4. THE BANDWIDTH TUNING CIRCUIT

The proposed filter is designed for a WiMAX/LTE receiver. Therefore, bandwidth of the filter must be tunable for different operation bandwidths depending on the targeted standard. The frequency tuning function was accomplished through a combination of coarse and fine-tuning techniques. The coarse tuning was performed by switching in/out a capacitor array which set the poles of the filter transfer function and the bandwidth for the targeted standard. On the other hand, fine-tuning was performed by changing the transconductance of the G_m-cells inside the biquad circuit. Both tuning processes were performed without affecting the Q of the filter transfer function (Oskooei 2011).

4.1 Coarse Tuning

The coarse tuning process was performed by changing the value of C2 and C3 capacitor in Equation 2 which are connected to LP and BP nodes in Figure 4. The values of capacitors were changed, for example k times to k.C, the cutoff frequency moves from f_0 to f_0/k, while the shape of the filter frequency response is retained. The coarse tuning scheme was able to set the filter frequency bandwidth to 5 MHz, 10 MHz, 15 MHz and 20 MHz that were aimed at different definitions for the baseband part of the WiMAX and LTE standards.

4.2 Fine Tuning

The fine-tuning process was performed by changing the transconductance of the G_m-cells inside the biquad circuit. This was achieved by altering the reference current sources (I_{ref}) which are shown along with linear G_m circuit in Figure 8 and the differential SF in Figure 9. These reference current sources can tune the value of the transconductances, depending on the ratio of the I_{ref} to main bias current (Oskooei 2011). For instance, with the filter bandwidth set to 10 MHz and the maximum I_{ref} value of 100μA, the proposed fine tuning technique can adjust the bandwidth in the range of the 9.6 MHz to 10.4 MHz.

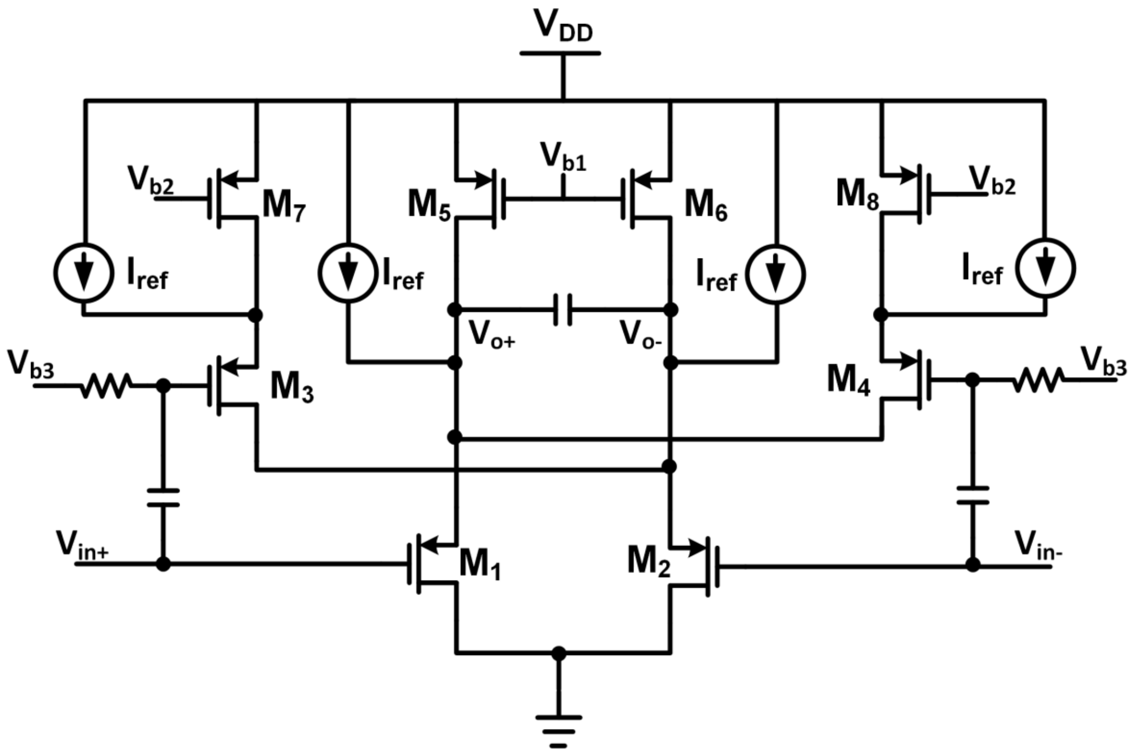

Figure 8. The circuit schematic of the G_m-cell along with fine tuning current sources.

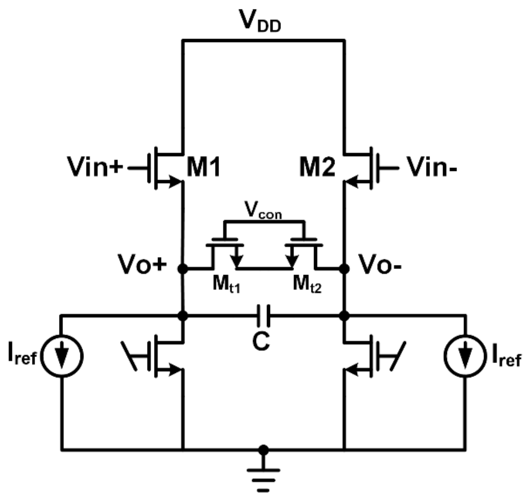

Figure 9. The circuit schematic of the differential SF circuit along with in-band tunable attenuation feature and fine tuning current sources (I_{ref}).

4.3 In-Band Attenuation

In a radio receiver chain, having a base band filter with in-band tunable gain or attenuation is vital. This is significant in setting the receiver overall gain that has to be tunable according to the received signal power (Razavi 1998). In a receiver in addition to the front-end section gain tunability can be performed in the baseband part. This action can be carried out by a combination of a variable gain baseband amplifier and a filter with tunable in-band gain/attenuation. The in-band attenuation of the proposed filter can be tuned by using the NMOS transistors M_{t1} and M_{t2}. The transistors were connected between differential outputs of the G_{m1} cell that is realized as a differential SF circuit. This is shown in Figure 9. These transistors operate in triode region and function as variable resistors; they steer part of the main signal current between the two differential outputs and set the attenuation inside the filter bandwidth. The resistance of these NMOS transistors in Figure 9 was adjusted by using the control voltage (V_{con}).

5. SIMULATION RESULTS

The proposed 3rd-order low pass filter was designed and simulated in CMOS 45nm process. The bandwidth of the filter was set by choosing value for the transconductances, capacitors inside the biquad cell and the input differential SF circuit. The frequency response of the filter for targeted bandwidth of 10MHz is shown in Figure 10. As shown in this figure, the filter attenuation inside the bandwidth is about 6 dB.

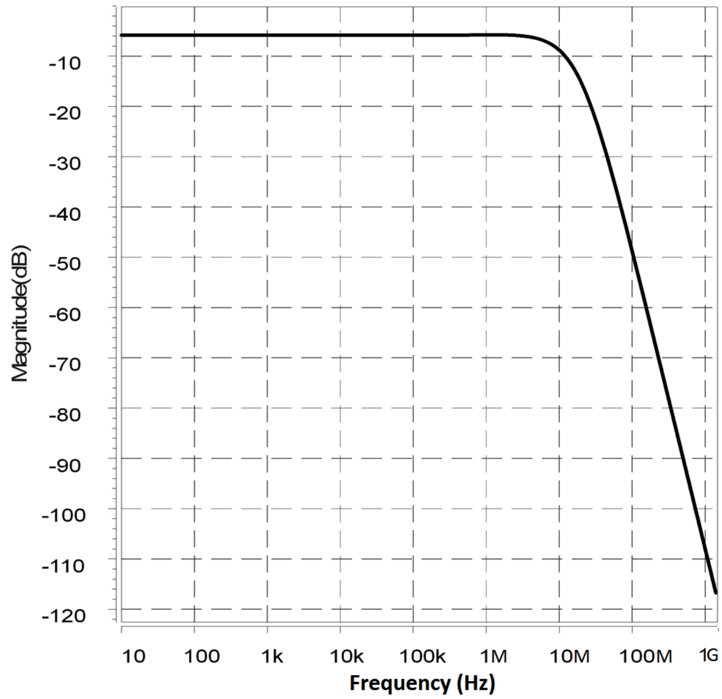

Figure 10. Frequency response of the proposed filter for bandwidth of 10 MHz.

Input referred noise density of the filter is shown in Figure 11. The minimum noise density of the filter is equal to 22.6 nV/√Hz inside the selected bandwidth.

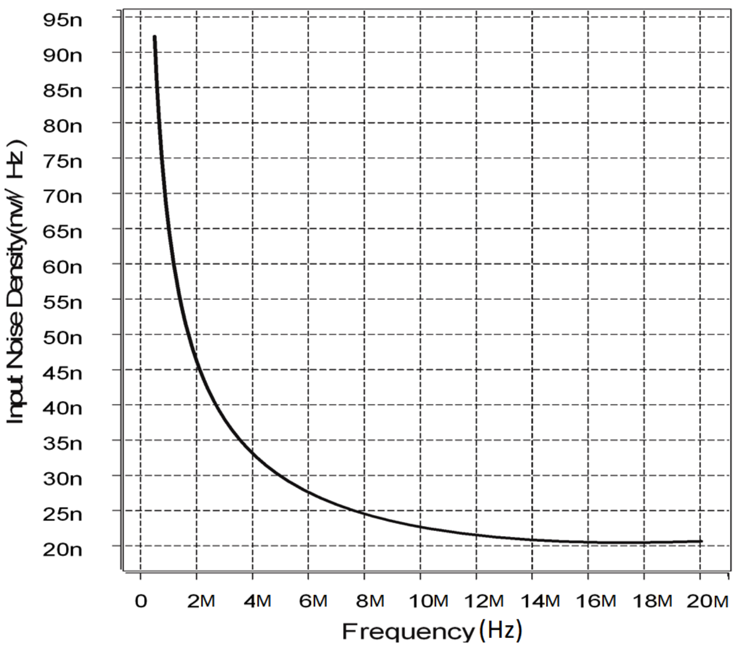

Figure 11. The simulated input referred noise density of the filter for bandwidth of 10 MHz.

In order to evaluate the circuit linearity and total harmonic distortion, a sinusoidal signal with amplitude of 200 mV and the frequency of 797 kHz was applied to the filter. The result of the Fast Fourier Transform (FFT) for output signal of the filter is shown in Figure 12. This result indicates a Total Harmonic Distortion (THD) of about -45 dB for the filter. In Figure 13, the output of the filter in time domain for the same sinusoidal input signal is shown.

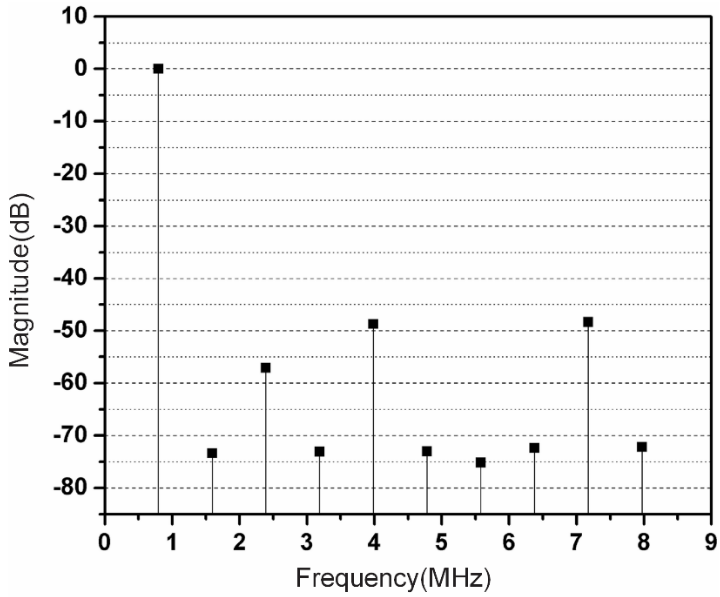

Figure 12. The FFT of the filter output for a sinusoidal signal with amplitude of 200 mV and frequency of 797 kHz.

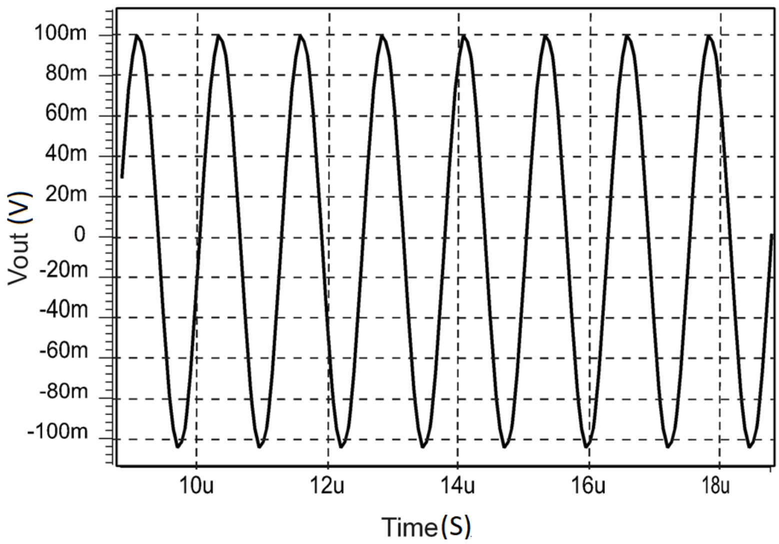

Figure 13. Time domain output of the filter for a sinusoidal input signal.

In Figure 14 simulation result of the filter IIP3 with two input tones at 3 MHz and 4 MHz is shown. As shown in this figure, the IIP3 of the filter is equal to 29.59 dBm.

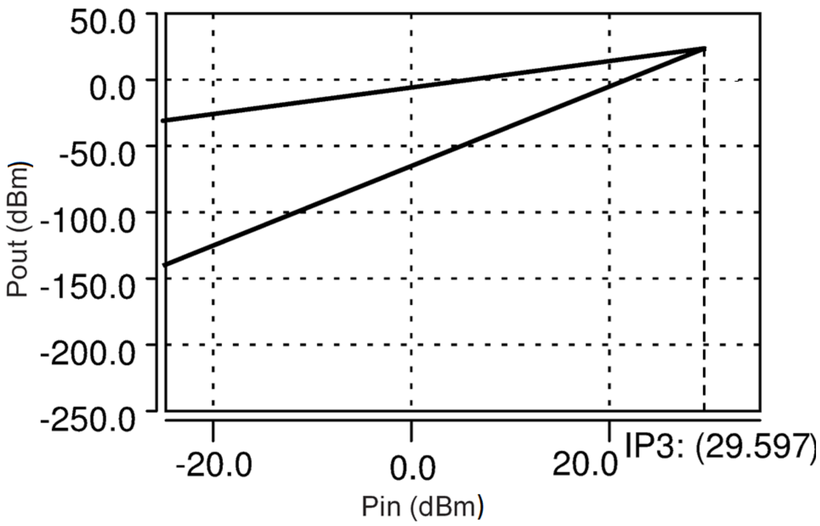

Figure 14. The IIP3 of the filter with two input tones at 3 and 4 MHz.

The result of the fine-tuning mechanism of the filter is shown in Figure 15. For this simulation, the filter bandwidth was set to 10 MHz and then was adjusted by changing the I_{ref} (bias current) from -100 to 100 µA. As a result, the bandwidth of the filter was fine-tuned from 9.6 to 10.4 MHz for this range of I_{ref} sweep.

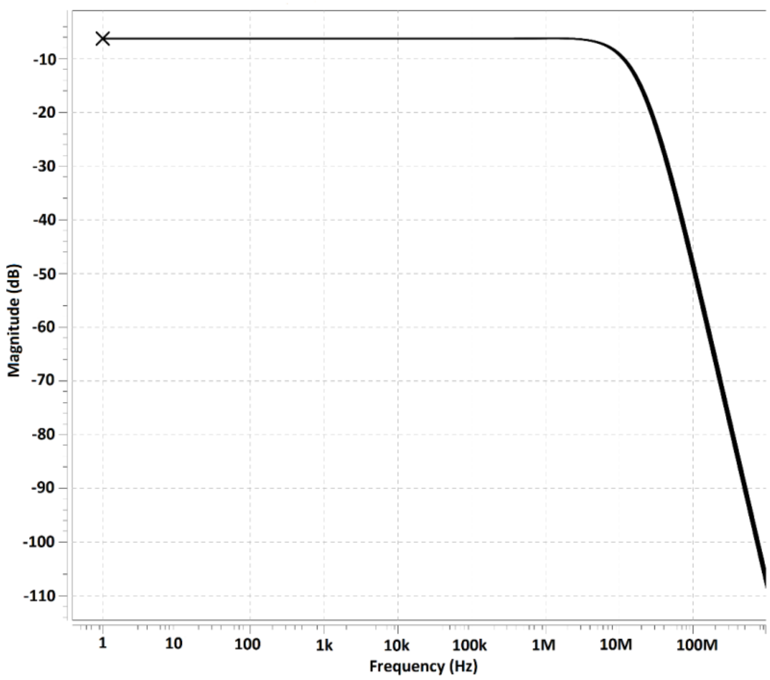

Figure 15. The results of fine tuning process of the filter for bandwidth of 10 MHz.

As shown in Figure 16. the filter bandwidth was set to frequencies of 5 MHz, 10 MHz, 15 MHz and 20 MHz by using coarse tuning mechanism. These frequencies were targeted for different revisions of the WiMAX/LTE standard.

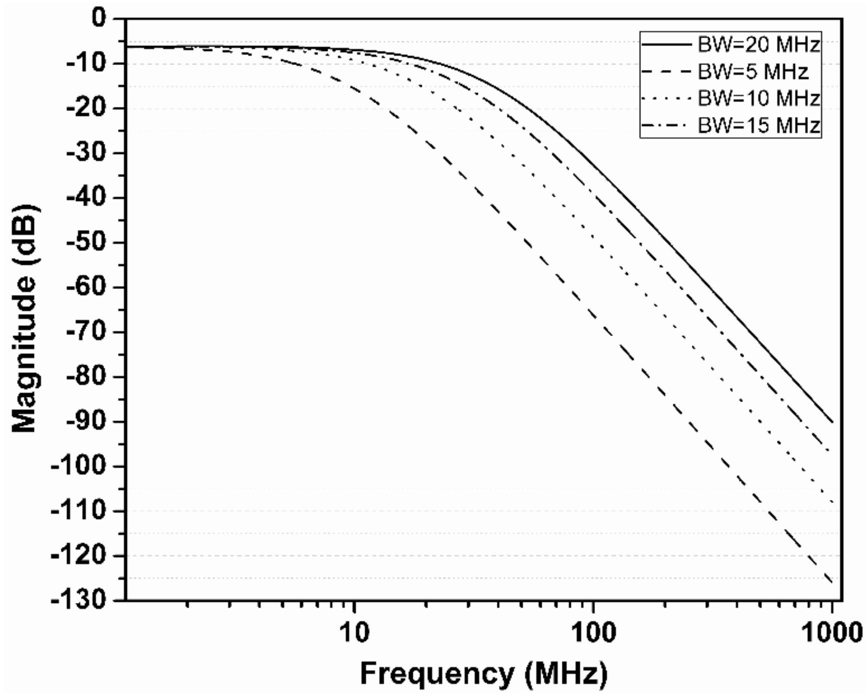

Figure 16. The coarse tuning of the filter bandwidth for frequencies of 5, 10, 15 and 20 MHz.

The tuning of the filter in-band attenuation is shown in Figure 17. This tuning was carried out by adjusting the control voltage V_{con}. As shown in this figure, the in-band attenuation of the filter was changed from -38 to -6 dB by sweeping the control voltage from 0 to1 V.

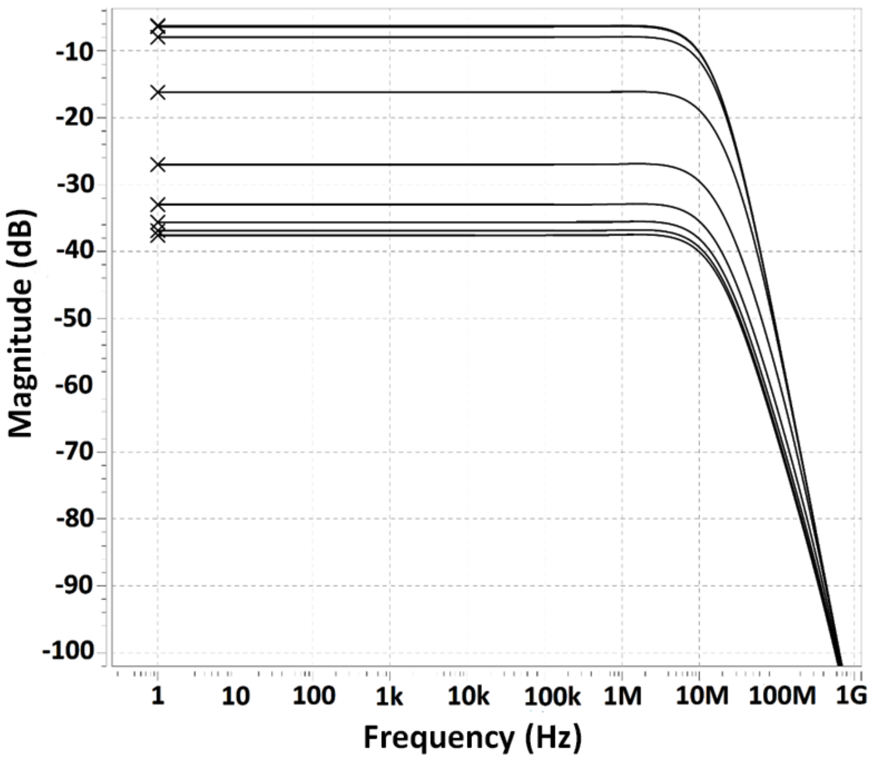

Figure 17. The adjustment of the filter in-band attenuation.

The figure of merit (FoM) (Oskooei 2011) in Equation 4 was used to summarize the filter performance and compare with some other recent designs:

$$FOM = \frac{P_C/_N}{f_C \cdot (SFDR^{4}/_{3})} \tag{4}$$

In Equation 4, P_C is power consumption of the filter, N is number of the poles and zeros, f_c is the cutoff frequency and $SFDR.N^{4/3}$ is the normalized spurious free dynamic range:

$$SFDR = (\frac{IIP3}{P_N})^{\frac{2}{3}} \tag{5}$$

where in this relation P_N is the input referred noise power (Oskooei 2011). According to Equation 4, smaller values for the FOM demonstrate higher performance for the designed filter. In Table 1. specifications of the proposed 3rd-order G_m-C filter are summarized. Table 2 compares the specification of the proposed filter with some recently reported works; the FOM in Equation 4 was used to compare the performance of these reported designs with the proposed filter in this paper.

Table 1 Summary of the filter specification

Technology	45 nm
Power supply	1V
Power consumption	10.08 mW
In-band attenuation	-6 dB
Tunable bandwidth	5- 20 MHz
Minimum In-band noise	22.6 nV/√Hz
In-band IIP3	29.59 dBm
THD for 200 mV input	-45 dB
FOM	0.019 fJ

Table 2 Performance Comparison of the proposed filter with some recent works

	(Oskooei 2011)	**(De Matteis 2009)**	**(Lo 2009)**	**This work**
CMOS Technology (nm)	90	130	180	45
Power Supply (V)	1	0.55	1.2	1
Power Consumption (mW)	4.35	3.5	4.1-11.1	10.08
Filter Order	6	4	3	3
Bandwidth (MHz)	8.1-13.5	11.3	0.5-20	5-20
In-band IIP3 (dBm)	21.7-22.1	10	19-22.3	29.59
In-band noise density (nV/√Hz)	75	33	12-425	22.6
FOM (fJ)	0.0246	0.103	0.0345-3.055	0.019

6. CONCLUSIONS

In this paper, a 3rd-order Butterworth low pass filter for a WiMAX/LTE receiver was presented. A linear G_m-cell was proposed which was later used to construct a biquad cell that was utilized in the filter to improve its linearity. The filter was constructed using a single buffer (G_m-cell) in series with a biquad block. The proposed filter bandwidth is tunable to different versions of the WiMAX/LTE standards. The tuning function comprises the coarse and fine-tuning mechanisms to cover a wider frequency span and compensate the process variations. The simulation results of the filter for bandwidth of 10 MHz show that IIP3 of the filter was about 29.59 dBm, the minimum in-band noise density was 22.6 nV/√Hz and power consumption of the filter was equal to 10.08 mW. The supply voltage of the filter was set to 1 V.

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