A Highly linear CMOS Gm-C Low Pass Filter for Mobile Communication

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Abstract: *A 2nd order low pass Butterworth filter has been designed for direct conversion receiver. Direct conversion receivers require only low pass filtering in the analogue baseband. This architecture has been applied for multi-mode transceiver. Low-pass filters are an essential part of the analogue baseband circuit of modern communication receivers, because of the ability to re-configure the bandwidth, gain or linearity in order to fulfill the requirements of different standards subjecting to wireless standards. Highly linear and wide tunable OTA based Butterworth filter has been simulated in CMOS 90nm Technology. Up to 625 MHz filter has a good linearity at supply voltage 1.2 V. The circuit shows a single OTA with common mode feedback has been prepared in microwind3.1 with 90nm CMOS Technology.*

Keywords: Direct conversion receiver, operational transconductance amplifier (OTA), multi-mode transceivers, transconductance, linearity schemes.

1. Introduction

In many wireless applications, a multimode filter is used in a direct conversion receiver. Various wireless standards [Bluetooth, Global System for Mobile (GSM) Communication, Wide Band Code Division Multiple Access (WCDMA), and Universal Mobile Telecommunication System (UMTS) are used in multimode transceiver. A low pass filter that has a wide tuning range is essential component for multimode filter because it can be adapted to all system specifications and forms essential element in a direct conversion receiver (DCR). It has been observed that in receivers, very demanding high-performance analog filters are typically used to block interferers and providing anti-alias filtering before the subsequent analog to digital conversion stage.

However it is challenging to design analog filters with low power consumption and large dynamic range [1], [2].To meet receiver specifications [3] filter design has been carried out using an operational transconductance amplifier as a basic building block [4]. OTA-C (or Gm-C) based filters has been used due to their ability to operate over a wide tuning range maintaining a high linearity. Various linearity schemes are reported [5]-[20] to realize OTAs. Cascode structures are useful for high speed application but are limited in output swing, thereby decreasing transconductance value. Also these structures are quite complex and increase power consumption. To further optimize OTA-C filter's performance, it is desired to achieve a good balance between signal-to-noise ratio (SNR) and signal-to-distortion ratio (SDR). Therefore, the use of linearity scheme without sacrificing other important parameters such as frequency response, noise level, power efficiency and is a must. In this paper highly linear CMOS operational transconductance amplifier (OTA) using a source degeneration and auxiliary differential pair (ADP) technique is proposed.

2. OTA Design

This section gives description of the techniques which improve linearity.

2.1 OTA with Source Degeneration

In "*Fig. 1(a)*," the double ended differential pair structure shown, is a basic cell, all transistors operate in saturation region then single ended output current (i_d) can be expressed by Taylor series expansion as;

$$i_d = \sum_{n=0}^{\infty} G_{MN\_(2n+1)} \cdot v_{in}^{2n+1} \tag{1}$$

For moderate signal swing, first two terms G_{MN_1} and G_{MN_3} are the most significant; is linear transconductance term and undesired third – order nonlinear term respectively. A rough approximation has been made based on saturation-square model and taking the mobility degradation effect due to lateral and transversal Electric field represented as an equivalent resistor at the source [3] leads to

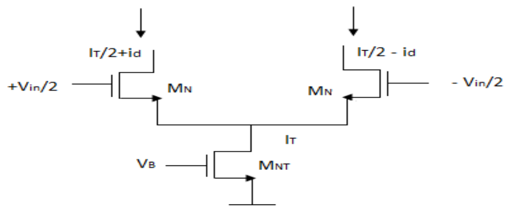
$$G_{MN\_1} = \frac{1}{2}\frac{\sqrt{K_P\left(W_N/L_N\right)I_T}}{1+\frac{2}{\epsilon_{crit}}\sqrt{\frac{I_T}{W_N\,L_N K_P}}} \tag{2}$$

The third-order coefficient of nonlinearity is

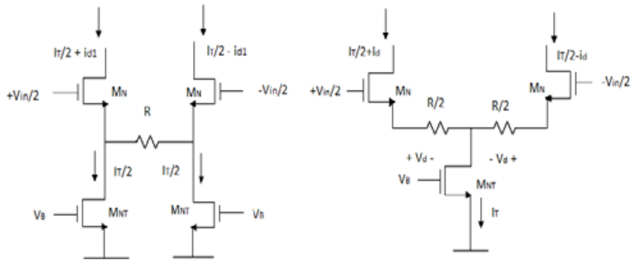
$$G_{MN\_3} = -\frac{G_{MN\_1}}{8\left(\frac{I_T L_N}{K_P.W_N}\right)\left(1+\frac{2}{\epsilon_{crit}}\sqrt{\frac{I_T}{W_N L_N K_P}}\right)^3} \tag{3}$$

Where W_N is width and L_N is length of transistor M_N. I_T is the tail current, K_P technological parameter and ε_{crit} is critical electrical field.

If a source degeneration resistance is connected between source terminals of two M_N transistors as shown in *"Fig.1 (b)* and *(c)"*



(a)



(b)

Figure 1: differential pairs: a)conventional differential pair; (b) differential pair with degeneration resistance; and (c) differential pair with degeneration resistance and tail current transistor

The single ended ac component of the output current approximately becomes

$$i_{d1} \cong \frac{G_{MN\_1}}{(1+G_{MN\_1}R)} V_{in} + \frac{G_{MN\_3}}{(1+G_{MN\_1}R)^4} V_{in}^3 \qquad (4)$$

Due to the source degeneration resistance, factor of the source degeneration is defined as $N_r = G_{M1}R$, according to this, the improvement in third-order harmonic distortion ($HD3_{imp}$) of the circuits *"Fig. 1(b) and (c),"* over the circuit *"Fig. 1(a),"* is

$$HD3_{imp} = \frac{HD3_{(b,c)}}{HD3_{(a)}} = \frac{1}{(1+N_r)^3} \qquad (5)$$

For the differential pair of *"Fig. 1(b),"* the input referred thermal noise density is given by

$$\frac{\overline{V_{inoise}^2}}{\Delta f} = \frac{4KT}{G_{MN\underline{1}}}\left(Y + N_r + 2YN_r^2 \frac{G_{mT}}{G_{MN\_1}}\right) \qquad (6)$$

Where, T is the temperature (^{0}K), k is the Boltzmann constant, G_{MT} is the small-signal transconductance of the transistor M_{NT}.

OTA's linearity can be improved by increasing N_r according to (5), but (6) shows that this approach will increase the input referred noise level, because the two tail current transistor introduces some differential noise if the value of source degenerated resistors is large.

This parameter can be minimized by placing the tail current transistor in the middle of the source degeneration resistor as shown in *"Fig. 1(c),"* by which the noise contribution of the tail current split equally in both branches and appears as common mode noise which is rejected due to fully differential nature of the topology.

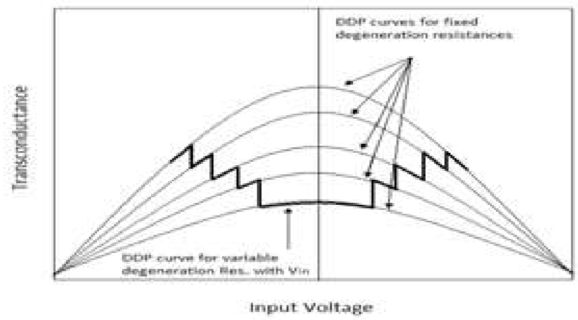
This topology has another advantage that the common node is always unaffected by the differential signal variations.

The drawback of the circuit shown in *"Fig. 1(c),"* is that the additional dc voltage drop through the degeneration resistors, thereby consuming voltage headroom for the input signal which imposes limitation overdriving voltages and source degeneration factor N_r must be used.

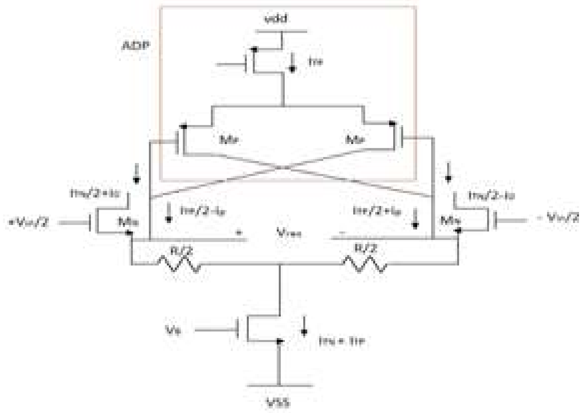
2.2. Auxiliary differential pair technique

To reduce the harmonic distortion components without sacrificing other parameters, auxiliary differential pair technique has been used here. The harmonic distortion can be reduced by other circuit techniques [5]-[20] but most of those techniques have limitations such as significant power consumption, reduced effective transconductance, and limited frequency response.

To illustrate the concept of auxiliary differential pair technique, let us consider several differential pairs, each one with its own degeneration resistor; the transconductance curves for fixed degeneration resistances and curves for variable degeneration resistance with input voltage are plotted in *"Fig. 2 (a),"*. By using ADP technique results in a flattened transconductance curve and enhanced linearity. Therefore degenerated resistor has been replaced by the parallel of a degenerated resistor and ADP shown in "*Fig. 2(b)"*.



(a)



(b)

Figure: 2.Linearity enhancement: (a) principle of operation; and (b) circuit implementation with auxiliary differential pair (ADP)

Based on the current to voltage relation as depicted in (1) for the differential pair of "*Fig. 1(a),"* the output current of the circuit in *"Fig. 2(b),"* written as

$$i_0 = G_{MN\_1}(v_{in} - v_{res}) + G_{MN\_3}(v_{in} - v_{res})^3 \qquad (7)$$

Where G_{MN_1} is the linear coefficient transconductance and G_{MN_3} is third-order nonlinear coefficient for the differential pair without source degeneration

V_{res} is voltage across the resistance and the ADP. The ADP current is given by

$$i_p = -G_{MP\_1}v_{res} - G_{MP\_3}v_{res}^3 \qquad (8)$$

Where G_{MP_3} is nonlinear term for the auxiliary diff. pair without source degeneration and G_{MP_1} is the small signal transconductance.

Since $i_0 = i_p + V_{res}/R$. It is shown in Appendix A, the output current can be found as

$$i_0 = \frac{G_{MN1}}{1+\frac{RG_{MN1}}{1-RG_{MP1}}} V_{in} + \frac{G_{MN3}-G_{MP3}\frac{G_{MN1}^4}{\left(R^{-1}-G_{MP1}\right)^4}}{\left(1+\frac{RG_{MN1}}{1-RG_{MP1}}\right)^4} V_{in}^3 \qquad (9)$$

The term with v_{in}^3 in (9) can be further reduced if the numerator of the right most term is minimized whereas in (4) can only be attenuated by increasing the source degeneration. The third-order intermodulation distortion (IM3) improvement factor can be described by

$$\frac{HD3_{withADP}}{HD3_{withoutADP}} \cong 1 - \left(\frac{G_{MP\_3}}{G_{MN\_3}}\right)\left(\frac{G_{MN\_1}R}{1-G_{MP\_1}R}\right)^4 \qquad (10)$$

The ADP must be designed such that the right hand most term remains near to 1. As the ADP transconductance is small, $G_{MP\text{-}1}$<<1, and it can be neglected; e.g. - 0.1

3. OTA Circuit

The proposed circuit of operational transconductance amplifier including CMFB is shown in "Fig. 3".

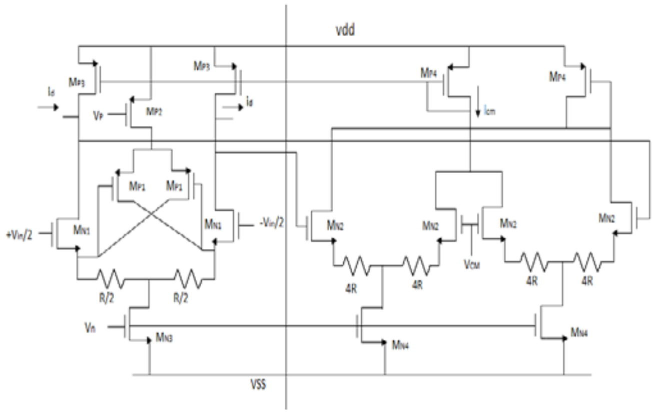


Figure 3: Complete Transconductor with common mode feed-back

The transconductor core consists of a degenerated differential pair with auxiliary differential pair technique. In the proposed circuit the transistor M_{N1} utilizes an overdrive voltage and the tail current flowing through the tail current transistor i.e. M_{N3}. The use of folded cascading or any other additional circuit has been avoided.

To optimize the structure for power and noise performances the output is taken directly from the drains of M_{N1}. For filters implementation the study of saturation voltages and threshold voltages for transistors M_{N3} has been carried out.

4. Filter Architecture

To implement the filter, Gm-C approach has been selected. The reason behind the selection of Gm-C concept is that the easy tuning capability by varying the Gm value of the transconductors. Also the Gm-C filter has a low noise floor but the ability to handle large signals is limited. In order to archive the required value in terms of the linearity the G_M which depends on the width, length and the bias current of the CMOS transistor has to be chosen carefully.

A circuit realization for the second order filter is used because of its advantages in design and layout. The fully differential circuit block is shown in *"Fig. 4,"*

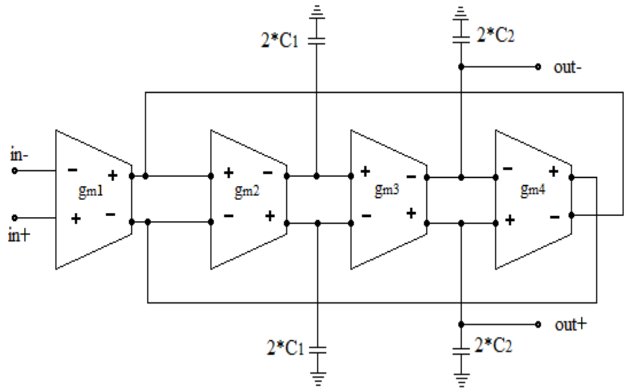


Figure 4: 2^{nd} order Gm-C low-pass filter circuits

The transfer function of *"Fig. 4,"* is given by [21]

$$H(S) = \frac{V_{out}}{V_{in}} = \frac{\frac{g_{m2}g_{m3}g_{m4}}{g_{m4}\ C_1C_2}}{S^2+\frac{g_{m2}}{C_1}S+\frac{g_{m3}g_{m4}}{C_1C_2}} \qquad (11)$$

The common transfer function of 2^{nd} order LPF is

$$H(S) = \frac{\omega_0^2}{S^2+\frac{\omega_0}{Q}S+\omega_0^2} \qquad (12)$$

The corner frequency ω_0 and the quality factor Q of the circuit can obtain as

$$\omega_0^2 = \frac{g_{m3}g_{m4}}{C_1C_2} \quad \frac{\omega_0}{Q} = \frac{g_{m2}}{C_1} \qquad (13)$$

$g_{m1}=g_{m3}=g_{m4}=g_m$ and $C_1=C_2=C$ leads to

$$\omega_0 = \frac{g_m}{C} \ \& \ Q = \frac{g_m}{g_{m2}} \qquad (14)$$

Where

$$g_m = \frac{I_{out+}-I_{out-}}{V_{in+}-V_{in-}} \qquad (15)$$

Now the transconductance g_{m2} of the 2^{nd} order filter [21]

$$g_{m2} = 1.848*g_m \qquad (16)$$

The 2^{nd} order low pass filter i.e. three of the four OTAs is identical; the outer OTA can easily be adapted with only changing the OTA current.

5. Simulation Result

The circuit shown in *"Fig. 5,"* shows a single OTA including common mode feedback and in *"Fig. 6,"* the circuit shows a complete filter structure has been prepared in microwind 3.1 lite version with 90nm CMOS Technology.

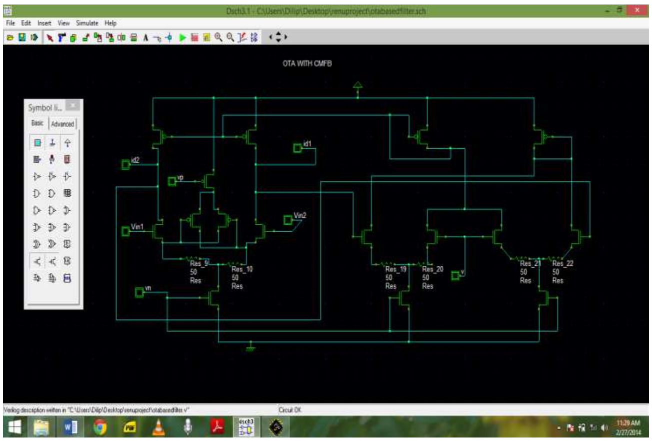


Figure 5: Operational transconductance amplifier (OTA) including common mode feedback (CMFB)

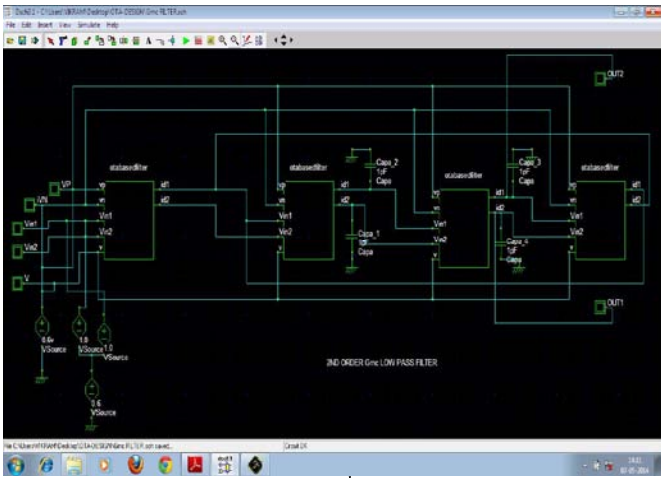


Figure 6: Circuit of 2nd order Gm-C filter.

The graph is drawn between current (mA) v/s time (ns) is shown in "*Fig. 7,*" As time is in nano second which shows the circuit operation in megahertz range.

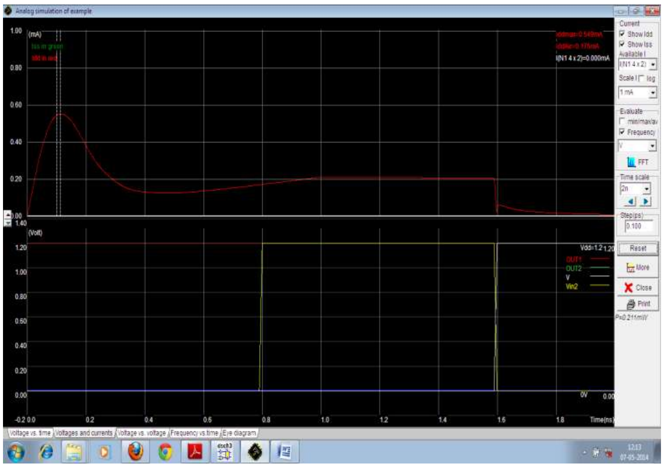


Figure 7: Result of current vs time

The graph of current vs time shows that at voltage 1.2V we find the maximum current 0.549 mA and minimum current 0.224 mA in time period of 1.6ns which shows low power consumption drawn by the circuit. The output voltage is constant at 1.2V showing the stability of the system.

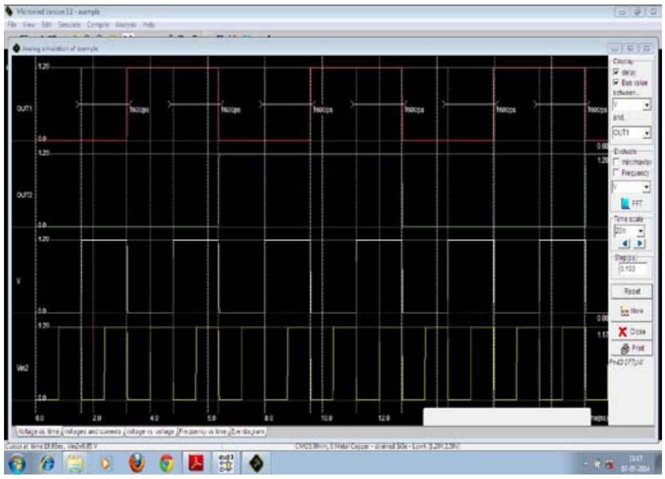


Figure 8: Result of current vs time

In "*Fig. 8,*" the graph is between voltage (V) v/s time (ns). When V_n applied as 0.6V after 1.6 ns the output graph in transient analysis shows the output graph in transient analysis shows the working of the device close to 625 MHz as shown in "*Fig. 9,*".

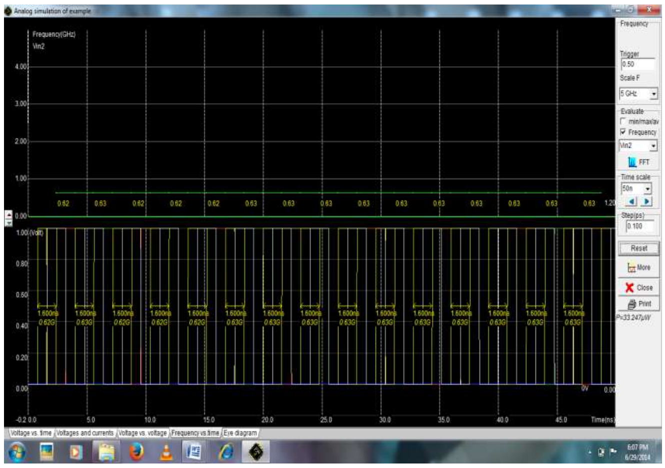


Figure 9: Frequency Vs time

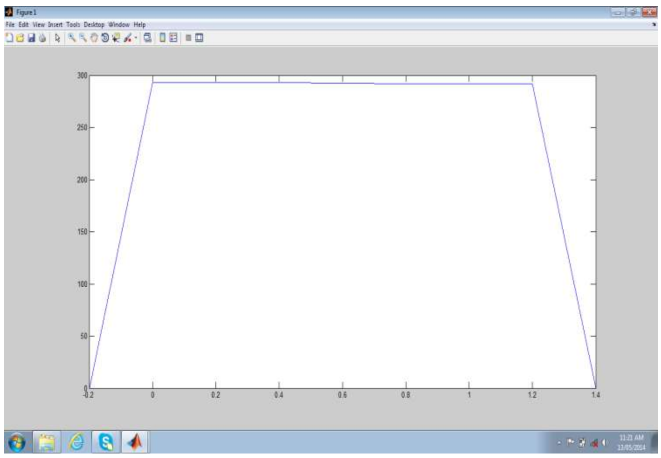


Figure 10: Transconductane plot against input voltage

In "*Fig 10,*" shows when we varying the input voltage value from 0 to 1.2V the then we get the flat transconductance curve and the transconductance value Tuned (varies) from 186µS to 457 µS.

6. Conclusion and Future Scope of Work

A 2nd order Butterworth low-pass Gm-C filter for Direct Conversion Receiver has been designed and simulated. The parameters which affecting the performance of a filter has been discussed. The approach is based on source degeneration and auxiliary differential pair technique allowing excellent performances. The simulated result shows that the filter has slew rate of 750 V/µS with 1.2 V power supply. The result also defines that the filter has a very good linearity over a wide range of frequency i.e. up to 625 MHz having transconductance tunable from 186 µs to 457 µs with the power consumption of 0.65 mw. The 2^{nd} order filter simulated is the best possible solution for direct conversion receiver used in mobile application. Layout of OTA with common mode feedback is as shown in *"Fig. 11"* and Layout of 2^{nd} order Gm-C filter shown in *"Fig. 12"*.

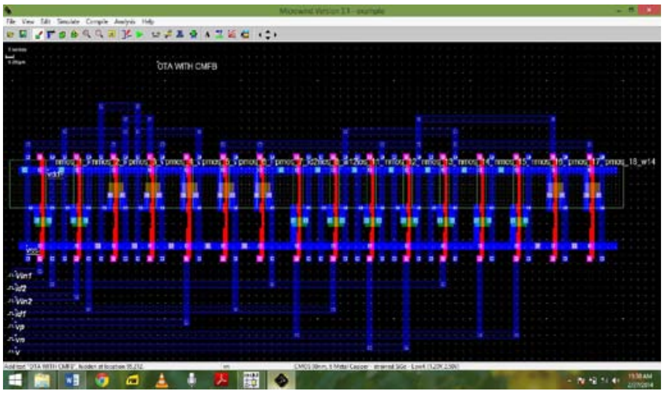


Figure 11: Layout of single OTA

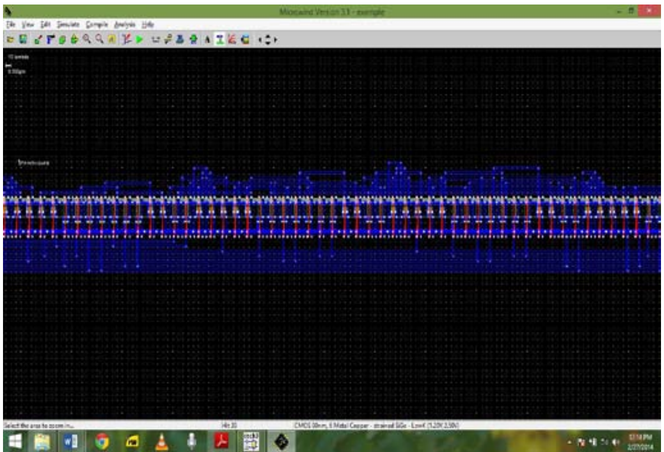


Figure 11: Layout of 2^{nd} order Gm-C filter

The demand of high data rate, higher speed and good linearity has grown steadily over the past decade. These demands will push the envelope of OTA design beyond today's limitations and to find an optimal balance between power consumption, noise and linearity for the requirements of direct conversion receiver circuit for various high speed and wireless application.

The research works achieved in this thesis present the following recommendations for future research as:

1. This is one method to design an Active-RC Butterworth filter using OTA. In future for better linearity and wide tuning range differential pair and cross-couple circuits can be optimized.
2. The design can be developed keeping in view low power and low voltages constraints as the filters are now being used in many portable applications that have the requirement of low power and low voltage.
3. Methods can be developed to reduce steep increment in power dissipation and component count with increases in the filter order.
4. Active devices can be designed in such a way that they may not restrict the passband frequency from expected passband frequency.
5. The transconductance range and linearity can be enhanced by using other suitable methodology.

7. Appendix

The single ended output current of the structure shown in *"Fig. 2(b),"* is

$$i_0 = i_p + v_{res}/R. \qquad \text{(A1)}$$

Using (8), this current results in

$$i_0 = \left(R^{-1} - G_{MP\_1}\right)v_{res} - G_{MP\_3}v_{res}^3. \qquad \text{(A2)}$$

Combining this equation and (7) becomes

$$G_{MN\_1}(v_{in} - v_{res}) + G_{MN\_3}(v_{in} - v_{res})^3 = (R^{-1} - G_{MP\_1})v_{res} - G_{MP\_3}v_{res.}^3 \qquad \text{(A3)}$$

The solution of this is complex and it is difficult to get any insight, let us assume that the v_{res} is an odd function of $v_{in;}$ if the high order terms are ignored then it could be approximated by the following expression:

$$v_{res} = C_1 v_{in} + C_3 v_{in.}^3 \qquad \text{(A4)}$$

In order to find the coefficients C_1 and C_3, (A4) is substituted in (A3), and if only the terms associated with v_{in} and v_{in}^3 are considered, the resulting equation yields

$$\left[G_{MN_1} - C_1\left(G_{MN_1} + R^{-1} - G_{MP_1}\right)\right]v_{in} + \left[G_{MN_3}(1 - C_1)^3 + G_{MP_3}C_1^3 - C_3\left(G_{MN_1} + R^{-1} - G_{MP_1}\right)\right]v_{in}^3 = 0 \qquad \text{(A5)}$$

This equation holds if the coefficients of v_{in} and v_{in}^3 are zero. Then, it is straightforward to find C_1 andC_3; therefore (A4) becomes

$$v_{res} = \frac{G_{MN\_1}}{G_{MN\_1} + R^{-1} - G_{MP\_1}}v_{in} + \frac{(R^{-1} - G_{MP\_1})^3 G_{MN\_3} + G_{MP\_3}G_{MN\_1}^3}{(G_{MN_1} + R^{-1} - G_{MP\_1})^4}v_{in}^3. \qquad \text{(A6)}$$

Finally, (A6) can be used in (A2) to obtain

$$i_0 = \frac{G_{MN\_1}}{1 + \frac{RG_{MN\_1}}{1 - RG_{MP\_1}}}v_{in} + \frac{G_{MN\_3} - G_{MP\_3}\frac{G_{MN\_1}^4}{(R^{-1} - G_{MP\_1})^4}}{\left(1 + \frac{RG_{MN\_1}}{1 - RG_{MP\_1}}\right)^4}v_{in}^3. \qquad \text{(A7)}$$

The meanings of this effect are examined in Auxiliary differential pair technique.

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