# Project - Table of Contents

# 1. Project theme

The requirements of the project will be to design a 3$^{rd}$ order Chebyshev low-pass filter with a 1dB ripple and a cutoff frequency of 4 Mhz using a TT-structure. Another requirement was the implementation of a programmable gain amplifier as a cascade of two-stages with different gain steps (coarse + fine). Each gain stage should be realized as a 2Gm PGA using a Kwan-Martin transconductor. Both using a lower supply voltage of 1.8 V and a fully differential approach. This will bring a larger output voltage swing, but with a disadvantage that they will require a common-mode feedback circuit to control the common-mode output voltage.

# 2. LPF

The sizing and denormalization of the filter components remain the same as for the single-ended design.

For the low pass filter the first step is to compute the transfer function of the filter using the required specifications. The goal for the ripple is 1dB for an implementation with real components (finite GBW operational amplifier) so in obtaining the transfer function we go for a lower ripple of 0.2dB.

The normalized transfer function is:

$$H_0(s) = H_0 \cdot \frac{c_1}{s + c_1} \cdot \frac{d_2}{s^2 + c_2 s + d_2} \tag{1}$$

There are tables that contain the factors of the denominator polynomials normalized Chebyshev low-pass filters.

| n | 0.2-dB Ripple (ε = 0.21709) |
|---|---|
| 1 | s + 4.60636 |
| 2 | $s^2$ + 1.92709s + 2.35683 |
| 3 | (s + 0.81463) ($s^2$ + 0.81463s + 1.41363) |
| 4 | ($s^2$ + 0.44962s + 1.19866) ($s^2$ + 1.08548s + 0.49155) |
| 5 | (s + 0.46141) ($s^2$ + 0.28517s + 1.11741) ($s^2$ + 0.74658s + 0.55839) |
| 6 | ($s^2$ + 0.19705s + 1.07792) ($s^2$ + 0.53835s + 0.64491) ($s^2$ + 0.73540s + 0.21190) |
| 7 | (s + 0.32431) ($s^2$ + 0.14433s + 1.05566) ($s^2$ + 0.40441s + 0.71644) ($s^2$ + 0.58439s + 0.29343) |
| 8 | ($s^2$ + 0.11028s + 1.04183) ($s^2$ + 0.31407s + 0.77124) ($s^2$ + 0.47004s + 0.38855) ($s^2$ + 0.55445s + 0.11795) |

*Figure 1: The factors of the denominator polynomials normalized Chebyshev low-pass filter for a 0.2dB ripple*

*Table 1: The coefficients of the normalized transfer function*

| c1 | c2 | d2 |
|---|---|---|
| 0.81463 | 0.81463 | 1.41363 |

The next step is to obtain the denormalized transfer function by replacing $s \rightarrow s/\omega_n$:

$$H_{denormat}(s) = H_0 \cdot \frac{c_{1d}}{s + c_{1d}} \cdot \frac{d_{2d}}{s^2 + c_{2d} s + d_{2d}} \tag{2}$$

Where fc = 4MHz:

$$c_{1d} = c_1 \omega_n \tag{3}$$

$$c_{2d} = c_2 \omega_n \tag{4}$$

$$d_{2d} = d_2 \omega_n^2 \tag{5}$$

$$\omega_n = 2\pi f_c \tag{6}$$

The resulting coefficients are:

*Table 2: The coefficients of the transfer function*

| c1d | c2d | d2d |
|---|---|---|
| 2.04e7 | 2.04e7 | 8.92e14 |

Once the coefficients are obtained, we must compute the values of the passive components but for this we need to consider the implementation of the filter. For this we use a lossy integrator and a Tow-Thomas biquad.

The general transfer function for a lossy integrator:

$$H(s) = H_0 \cdot \frac{\omega_0}{s + \omega_0} \tag{7}$$

In this case for the lossy integrator the transfer function is:

$$H(s) = 1 \cdot \frac{2.04739 \cdot 10^7}{s + 2.04739 \cdot 10^7} \tag{8}$$

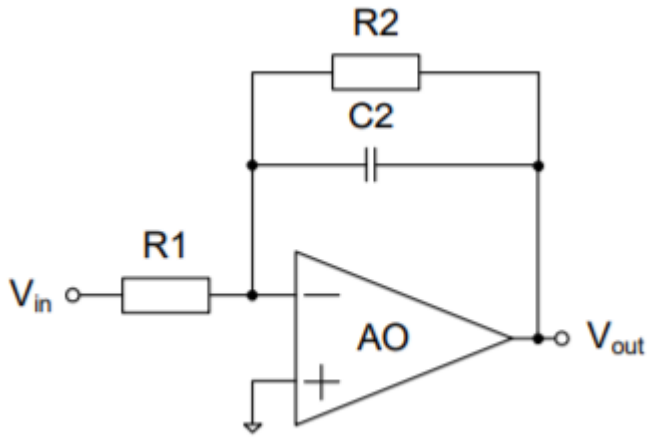


*Figure 2: The schematic of a single-ended lossy integrator*

The transfer function of this lossy integrator is:

$$H(s)=\frac{-R_2}{R_1}\cdot\frac{\frac{1}{R_2C_2}}{s+\frac{1}{R_2C_2}} \tag{9}$$

Since $H_0=1$ then $R_1=R_2$, and if we choose $C_2$=1pF the resulting values for the passive components are:

*Table 3: The values of the passive components of the lossy integrator*

| R1[Ohm] | R2[Ohm] | C2[F] |
|---|---|---|
| 48.8e3 | 48.8e3 | 1e-12 |

The general transfer function for a second order low pass filter:

$$H(s)=H_0\cdot\frac{{\omega_0}^2}{s^2+\frac{\omega_0}{Q}s+{\omega_0}^2} \tag{10}$$

In this case the transfer function for the 2nd order LPF is where $f_0$ = 4.76 MHz:

$$H(s)=1\cdot\frac{\left(2.99\cdot10^7\right)^2}{s^2+\frac{2.99\cdot10^7}{1.46}s+\left(2.99\cdot10^7\right)^2} \tag{11}$$

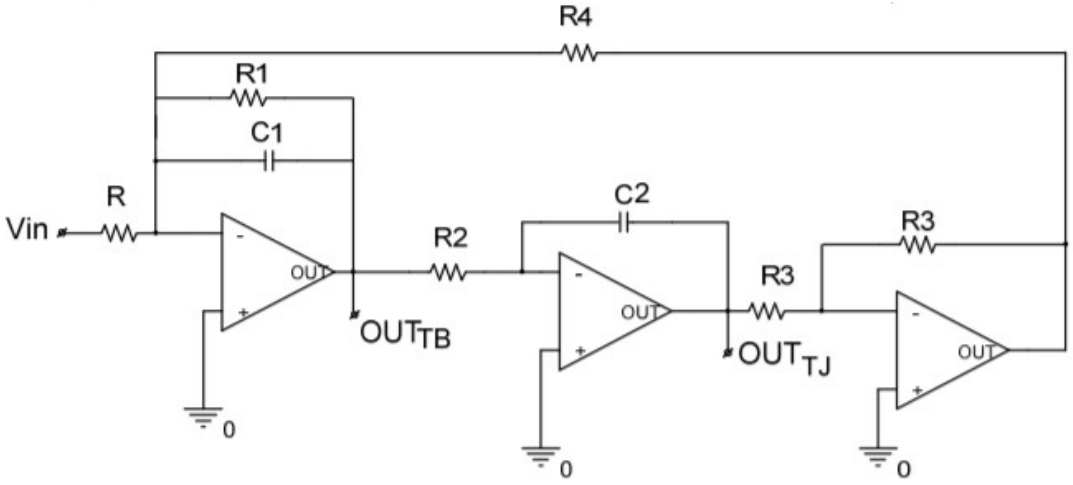


*Figure 3: The schematic of a Tow-Thomas biquad*

The transfer function of the Tow-Thomas biquad is:

$$H(s)=\frac{R_4}{R}\cdot\frac{\left(\frac{1}{\sqrt{R_2R_4C_1C_2}}\right)^2}{s^2+\frac{\frac{1}{\sqrt{R_2R_4C_1C_2}}}{R_1\cdot\sqrt{\frac{C_1}{C_2}\cdot\frac{1}{R_2R_4}}}s+\left(\frac{1}{\sqrt{R_2R_4C_1C_2}}\right)^2} \tag{12}$$

Since $H_0=1$ then $R_4=R$ and if we choose $C_1=C_2=$¿1pF and $R_4=R_2=R_3$ then $R_1=Q\cdot R_2$, where $R_2=\frac{1}{\omega_0\cdot 1\,pF}$. The values for the passive components are:

*Table 4: The values of the passive components of the tow-thomas-biquad*

| R1[Ohm] | R=R2=R3=R4[Ohm] | C1=C2[F] |
|---|---|---|
| 48.8e3 | 33.5e3 | 1e-12 |

## 2.1 LPF design at transistor level

The active component of the LPF is a fully differential AO Miller operational ampifier.

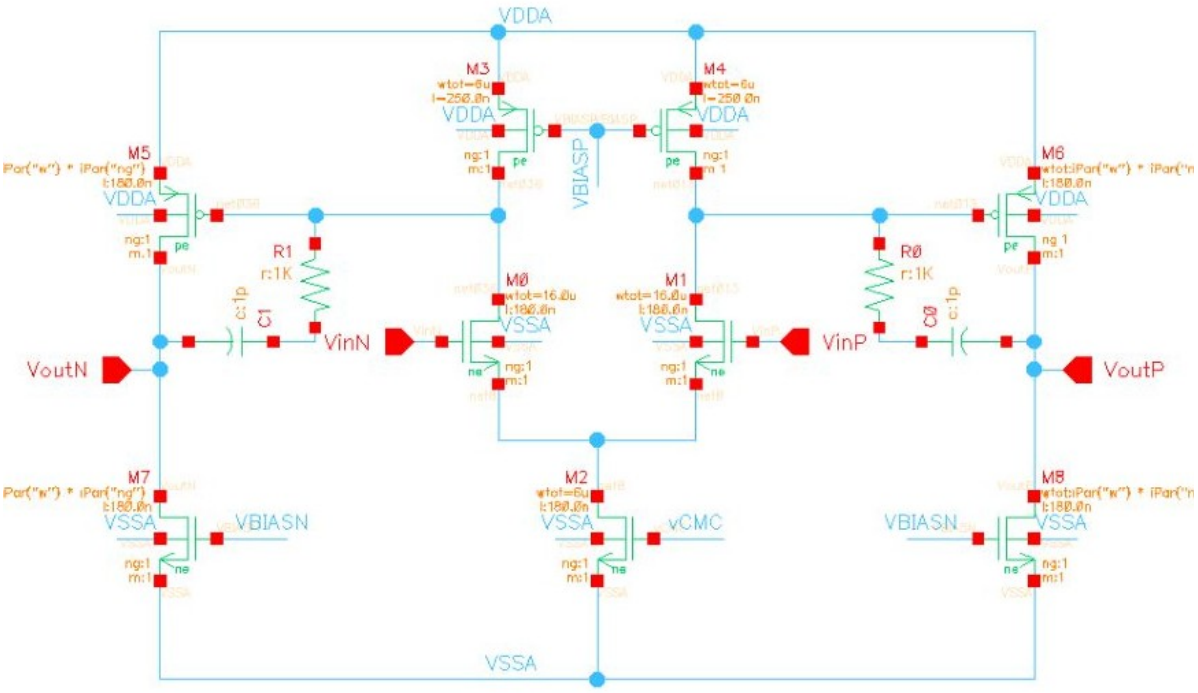


*Figure 4: The schematic of the fully-differential Miller operational amplifier with RC compensation*

The requirements for the OA were a gain bandwidth prooduct of 150 MHz, a Slew-Rate of at least 40 V/µs, a total current consumption of 150 µA, a phase margin greater than 60 degrees and a gain of at least 60 dB for a load of 1 pF.

| GBW[Hz] | $I_{DD}$[A] | SR[ V/µs] | PM[°] | $C_L$[F] | $a_0$[dB] |
|---|---|---|---|---|---|
| 150e6 | 150e-6 | 40 | 60 | 1e-12 | 60 |

Particular eq. for the two stage opamp with CMFB circuit (presented later) are the folowing:

DM path:

$$a_{dm0}=G_{m1}R_{o1}G_{m2}R_{o2}$$

$$|a_{dm0}f_{dm}p_{1d}|=GBW=1\cdot|p_{2d}|\ if\ the\ RHPZ\ has\ been\ eliminated,$$

where $f_{dm}$ is the feedback factor (can be instantiated to 1, if we are not concerned about the loop-gain), and GBW is the gain bandwidth product of the DM path.

$$\frac{G_{m1}R_{o1}G_{m2}R_{o2}}{R_{o1}(G_{m2}R_{o2})C_M}=\frac{G_{m2}}{C_L}\rightarrow\frac{G_{m1}}{C_M}=\frac{G_{m2}}{C_L} \tag{13}$$

CM path:

$$a_{cmc0}\approx g_{m2h}[(r_{o1}g_{m1}r_{o2h})\|r_{o4}]G_{m2}R_{o2}$$

$$\left(\frac{W}{L}\right)_{2h}=\frac{(W/L)_2}{2}\ \text{and}\ I_{d2h}=I_{D2}/2$$

$$|a_{cmc0}a_{cms0}p_{1c}|=1\cdot|p_{2c}|\rightarrow\frac{g_{m2h}}{C_M}|a_{cms0}|\approx\frac{G_{m2}}{C_L} \tag{14}$$

where $a_{cms0}$ is the low-frequency gain through the CM-sense circuit.

Through an iterative process we can start by selecting CM, and verifing the previous equations again:

1. We choose $C_M=0.22\cdot C_L=220\,fF$

2. From the equation for the gain bandwidth product we obtain $G_{m1}$

$$G_{m1}=2\pi C_M GBW \tag{15}$$

$$G_{m1}=g_{m0,1}=207\,\mu S \tag{16}$$

3. We choose $V_{dsat\,0,1}=150\,mV$

4. From the previous two steps we can compute $I_{BIAS}$

$$I_{BIAS}=G_{m1}\cdot V_{dsat\,0,1}=31.1\,\mu A \tag{17}$$

5. We choose $V_{dsat\,3,4}=150\,mV$

6. We can compute the current through the output stage as the difference between the current budget and the curent through the first stage.

$$I_{D6}=I_{D5}=\frac{I_{DD}-I_{BIAS}}{2}=\frac{150\text{e-}6-31.1\text{e-}6}{2}\approx 60\text{e-}6 \tag{18}$$

7. We choose $V_{dsat\,6,5}=V_{dsat\,3,4}=150\,mV$

8. We can now compute $G_{m2}$

$$G_{m2}=\frac{2\,ID_6}{V_{dsat\,6}}=\frac{2\cdot 60\text{e-}6}{0.15}=800\,\mu S \tag{19}$$

9. We have to split the low frequency gain between the two stages. For this we need to consider the linearity of the amplifier. We have to ensure that even for the maximum output voltage the input voltage for the second stage is not larger than the linear range of $M_6$. Also since $R_{O2}$ is the equivalent output of the amplifier we don't want it to be too large, so we prefer to have a higher gain in the first stage. Taking all this into consideration we obtain: $a_{v1}=50\,V/V$ and $a_{v2}=20\,V/V$.

10. Since we know both the gain and the transconductance of each stage we can compute the output resistance:

$$R_{O1}=\frac{a_{v1}}{G_{m1}}=\frac{50}{207\text{e-}6}\approx 240\,k\,\Omega \tag{20}$$

$$R_{O2}=\frac{a_{v2}}{G_{m2}}=\frac{20}{800\text{e-}6}=25\,k\,\Omega \tag{21}$$

11. Next we have to verify that the computed Slew-Rate is greater than the required one:

$$SR\_OUT=\frac{I_{D5,6}}{C_L}=\frac{60\text{e-}6}{1\text{e-}12}=60\,V/\mu s \tag{22}$$

$$SR\_\mathrm{IN} = \frac{I_{BIAS}}{C_1} \approx \frac{31.1\text{e-}6}{C_M} = 141\, V/\mu s \tag{23}$$

12. Taking into consideration the requirement for the phase margin we can compute the value for $R_M$.

$$PM = 90° - \tan^{-1}\frac{GBW}{f_{p2}} - \tan^{-1}\frac{GBW}{f_z} \tag{24}$$

where

$$f_{p2} = \frac{G_{m2}}{2\pi C_L} \tag{25}$$

$$f_z = \frac{1}{2\pi C_M\left(\frac{1}{G_{m2}} - R_M\right)} \tag{26}$$

From these equations it follows that for a PM of 60° we need: $R_M = 100\,\Omega$

We can verify eq. 1:

$$\frac{G_{m1}}{C_M} = \frac{G_{m2}}{C_L} \rightarrow C_M = C_L\frac{G_{m1}}{G_{m2}} = \frac{1\text{e-}12 \cdot 207\text{e-}6}{792\text{e-}6} = 261\, fF$$

And from eq. 2 we can get $a_{cms0}$:

$$\frac{g_{m2h}}{C_M}|a_{cms0}| \approx \frac{G_{m2}}{C_L} \rightarrow |a_{cms0}| = \frac{\frac{G_{m2}}{C_L}\cdot C_M}{g_{m2h}} = 0.15$$

These are the resulting values of the transistors design sheet:

| M | \|ID\| [A] | \|Vdsat\| [V] | gm [A/V] | W [m] | [L] | TYPE |
|---|---|---|---|---|---|---|
| M3,4 | 15.5e-6 | 0.15 | 207e-6 | 5.44e-6 | 250e-9 | PMOS |
| M0,1 | 15.5e-6 | 0.15 | 207e-6 | 2.4e-6 | 250e-9 | NMOS |
| M2 | 31e-6 | 0.11 | 565e-6 | 4e-6 | 250e-9 | NMOS |
| M5,6 | 60e-6 | 0.15 | 792e-6 | 22.8e-6 | 180e-9 | PMOS |
| M7,8 | 60e-6 | 0.11 | 1e-3 | 7.7e-6 | 250e-9 | NMOS |

*Table 5: Resulting values for the transistors dimensions*

### 2.1.1 CMFB

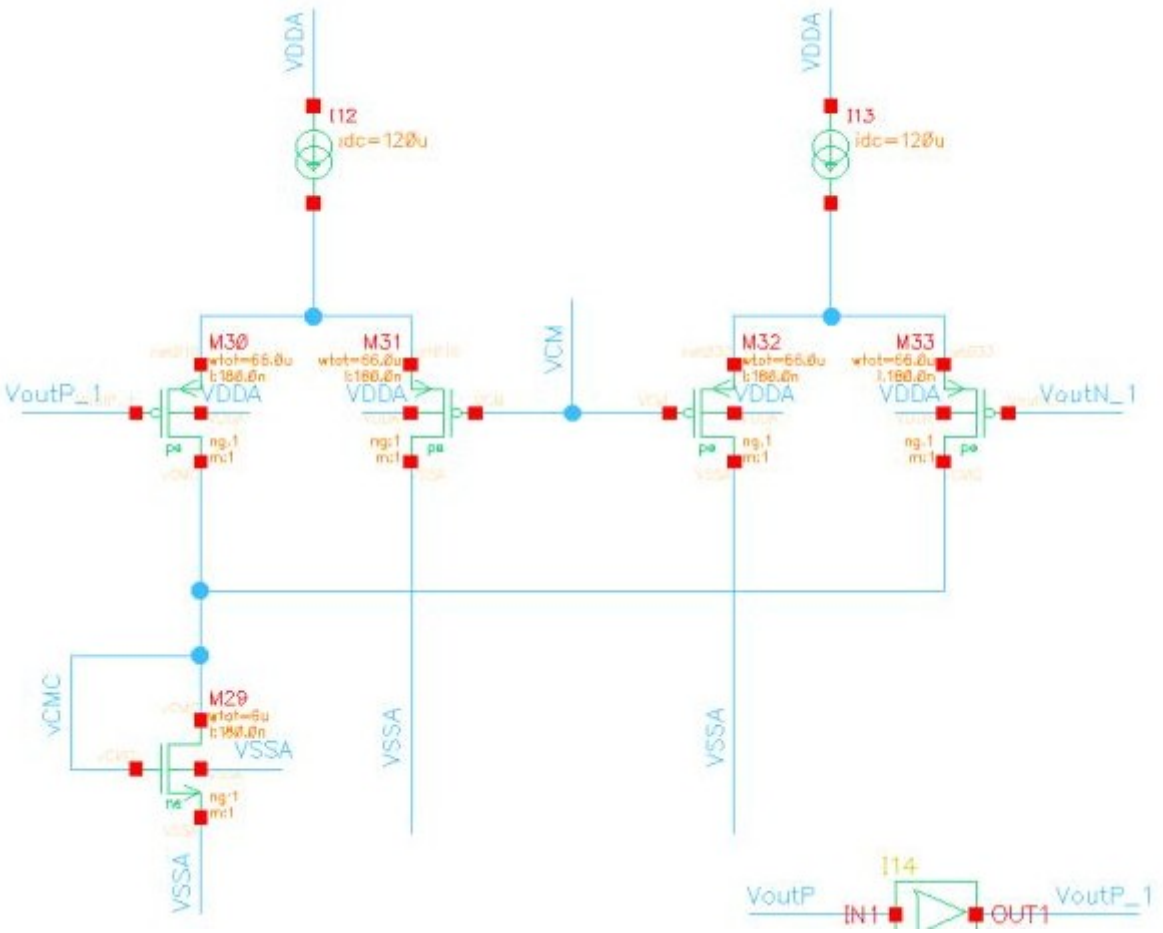


*Figure 5: CMFB Circuit*

→ small-signal analysis:

$$I_{D30}=\frac{I_{12}}{2}+g_{m30}\frac{(V_{CM}-V_{outP\_1})}{2} \tag{27}$$

$$I_{D33}=\frac{I_{13}}{2}+g_{m33}\frac{(V_{CM}-V_{outN\_1})}{2} \tag{28}$$

These currents are summed in diode-connected M29 to give the CM sensor output current:

$$I_{cms}=I+g_{m30}\left(V_{CM}-\frac{V_{outP\_1}+V_{outN\_1}}{2}\right)=I+g_{m30}\left(V_{CM}-V_{oc}\right) \tag{29}$$

*where* $I_{12}=I_{13}=I$, and $g_{m30}=g_{m33}$

We note that one design option is to select $I_{13}=I_{12}=|I_{D3}|+|I_{D4}|$ and $(W/L)_{29}=(W/L)_2$.

## 2.2 Verifying and characterizing the LPF

### 2.2.1 Verifying and characterizing the operational amplifier

Using the previously obtained values we obtain the following results for the amplifier:

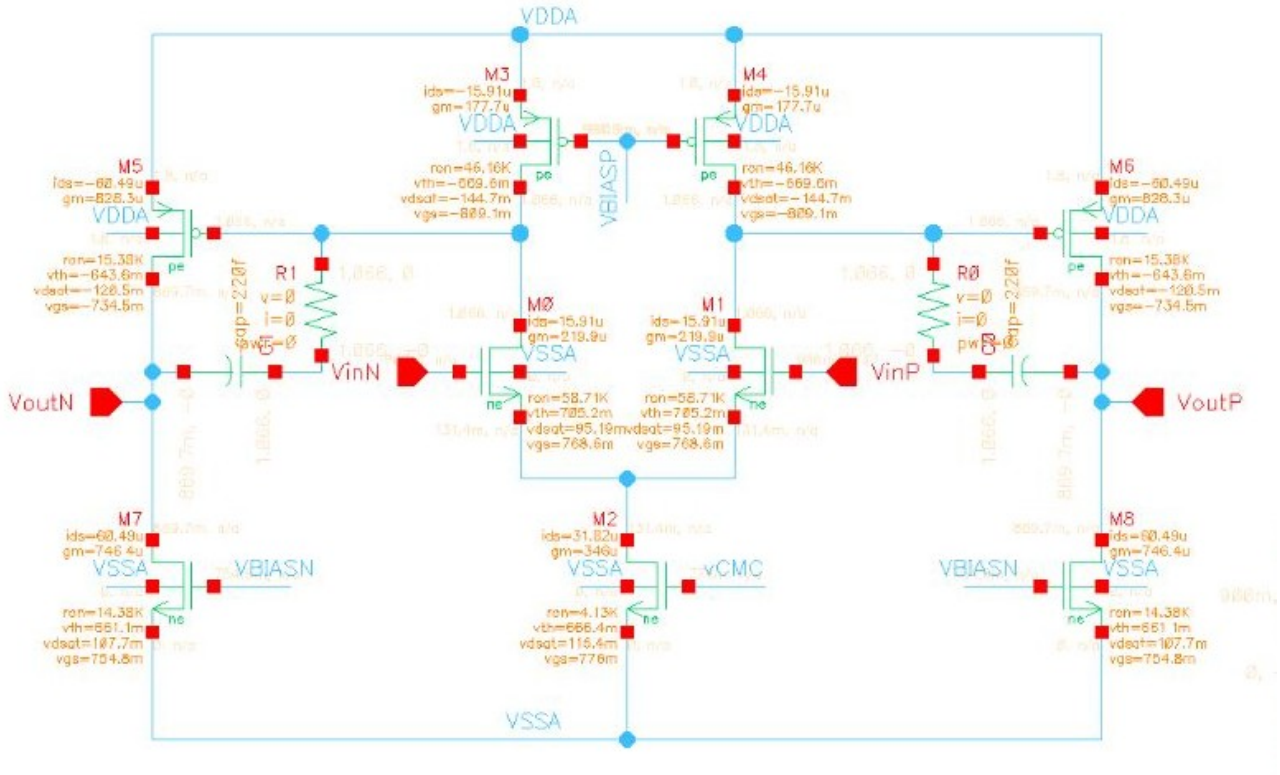


*Figure 6: The DC operating point of the transistors*

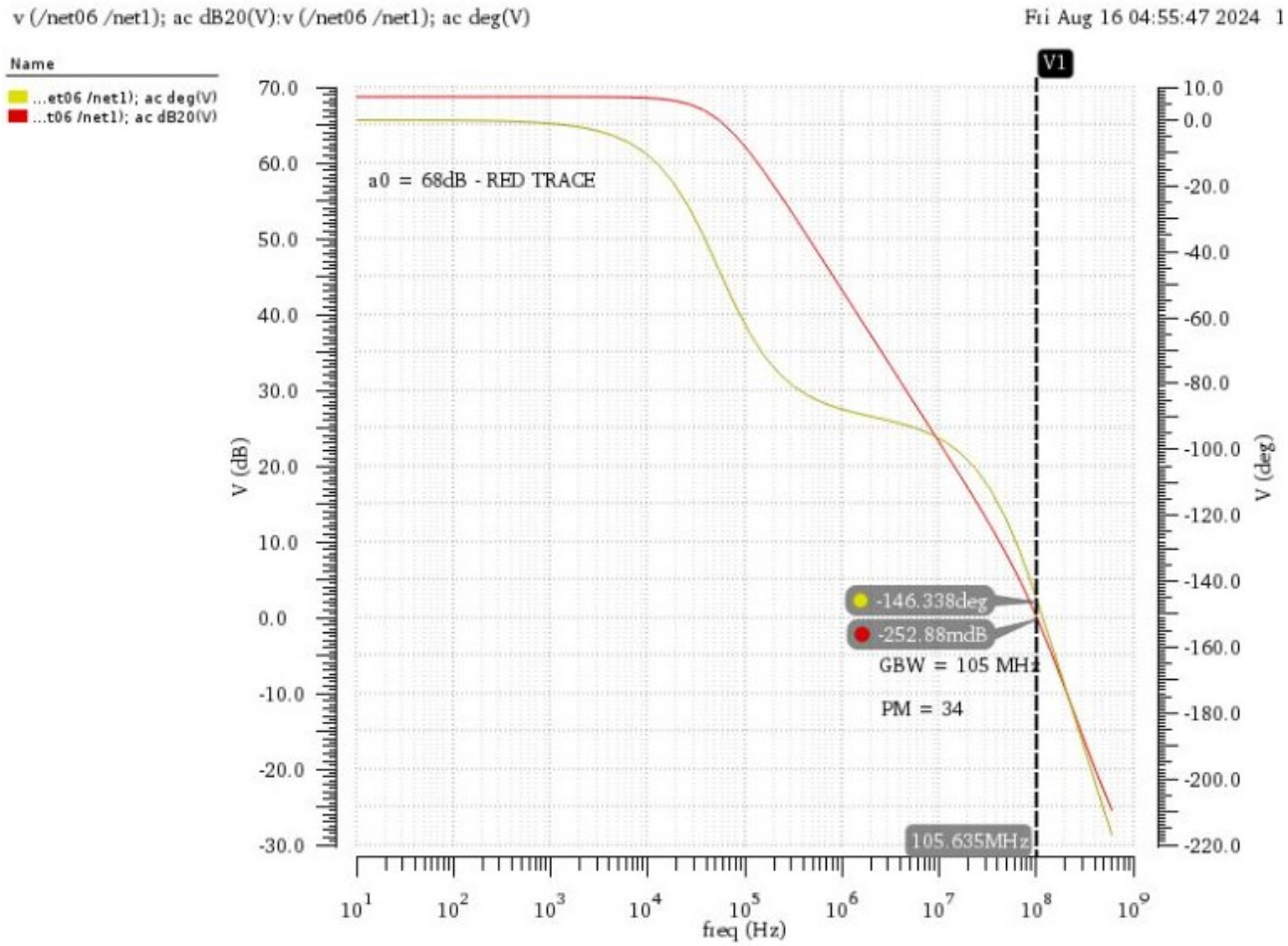


*Figure 7: The transfer function of the amplifier*

| $a_0$ [dB] | GBW [Hz] | PM[°] | $I_{DD}$ [A] w/o CMFB |
|---|---|---|---|
| 68 | 105e6 | 34 | 168e-6 |

CMFB:

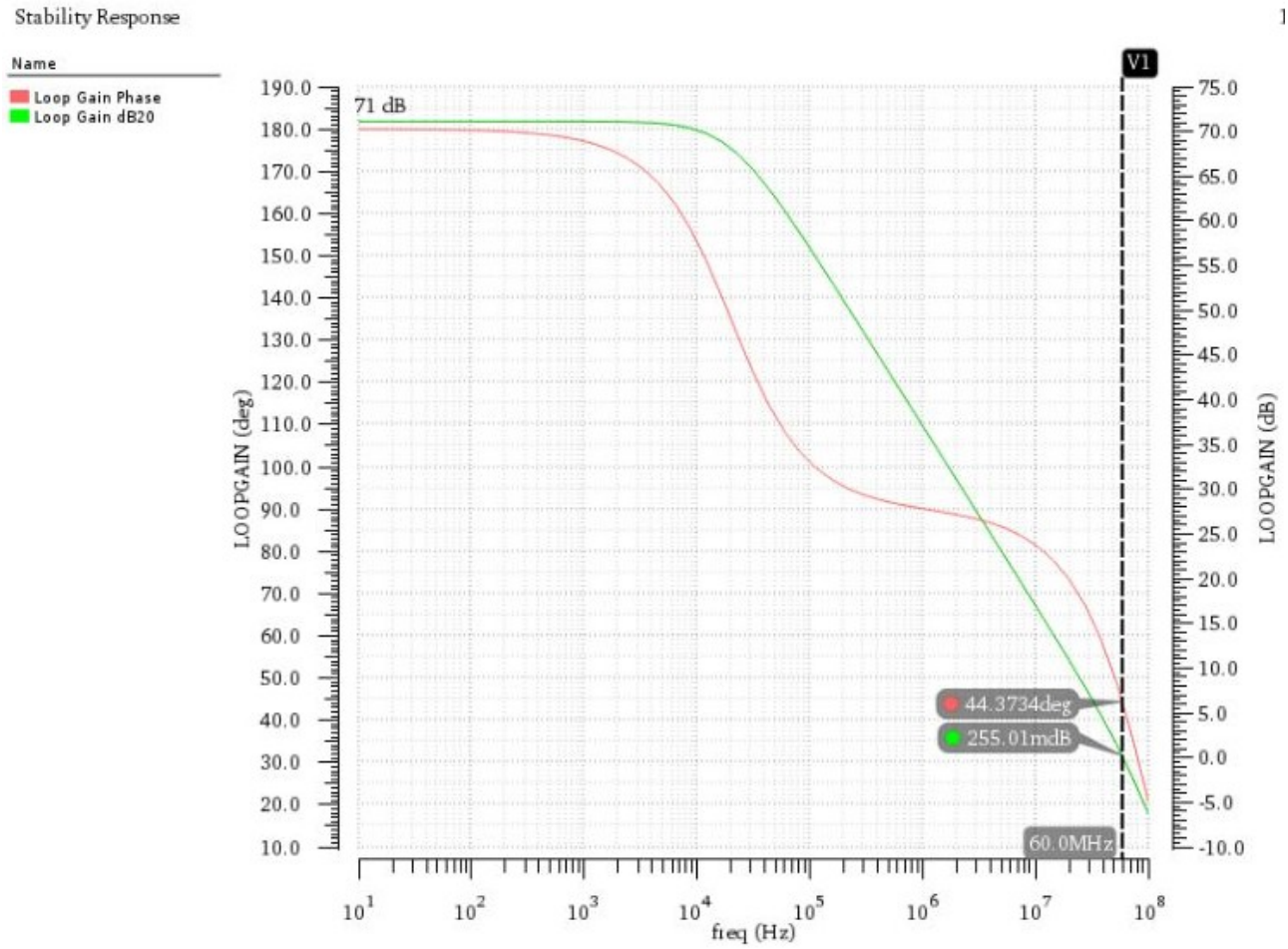


*Figure 8: Loop-gain of CMFB circuit*

**As can be seen above after the first iteration both the phase margin and the gain bandwidth product are too small.**

To make for the GBW and PM which are too small for the previous design we can restart the iteration with a higher GBW specification and a smaller $C_M$. The folowing table reveals the new transistors data for a GBW of 200 Mhz and a CM of 120 fF.

| M | \|ID\| [A] | \|Vdsat\| [V] | gm [A/V] | W [m] | [L] | TYPE |
|---|---|---|---|---|---|---|
| M3,4 | 11.3e-6 | 0.15 | 150e-6 | 4e-6 | 250e-9 | PMOS |
| M0,1 | 11.3e-6 | 0.15 | 150e-6 | 1.72e-6 | 250e-9 | NMOS |
| M2 | 22.6e-6 | 0.11 | 411e-6 | 2.92e-6 | 250e-9 | NMOS |
| M5,6 | 63.69e-6 | 0.15 | 792e-6 | 25e-6 | 180e-9 | PMOS |
| M7,8 | 63.69e-6 | 0.11 | 1.11e-3 | 8.25e-6 | 250e-9 | NMOS |

*Table 6: Resulting values of the transistors dimensions for the second iteration*

After the second iteration I obtained the following results:

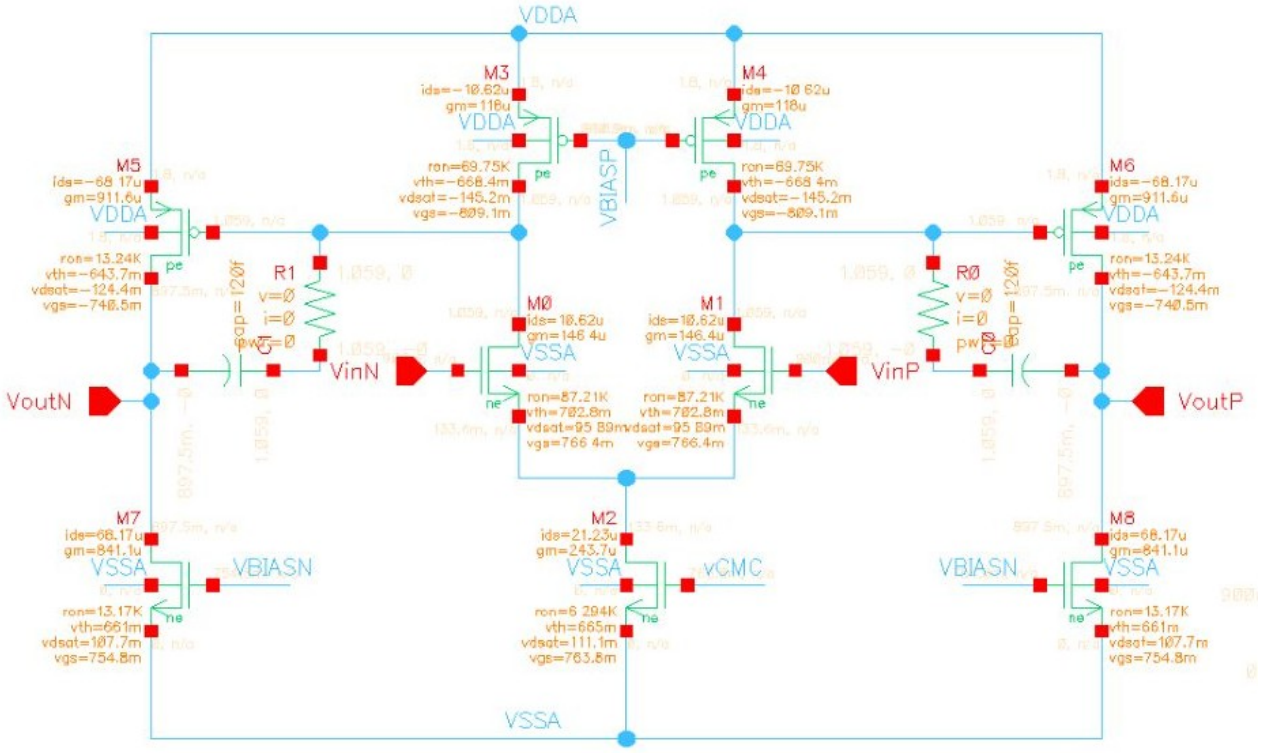


*Figure 9: The DC operating point of the transistors*

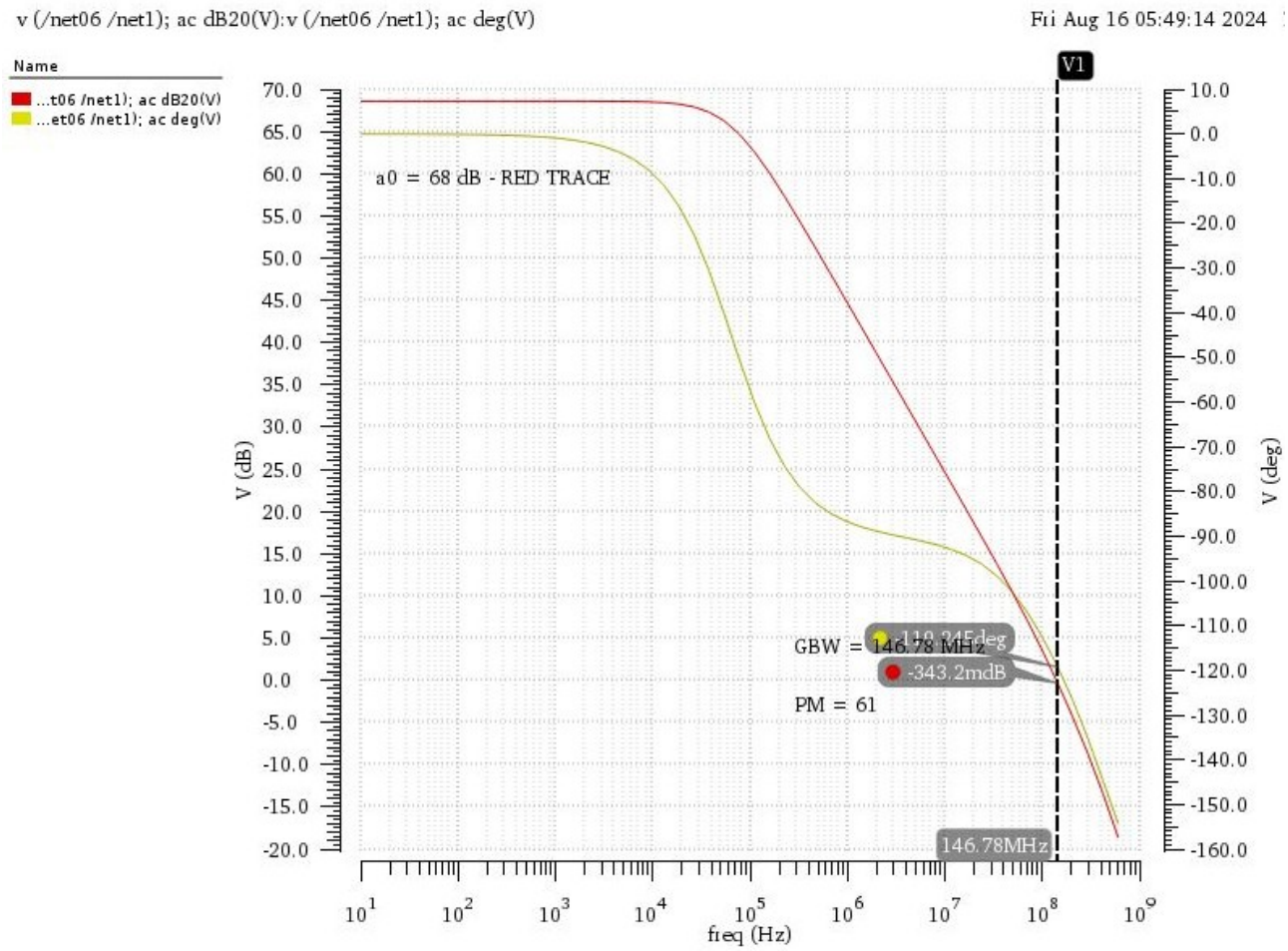


*Figure 10: The DM open-loop transfer function of the amplifier*

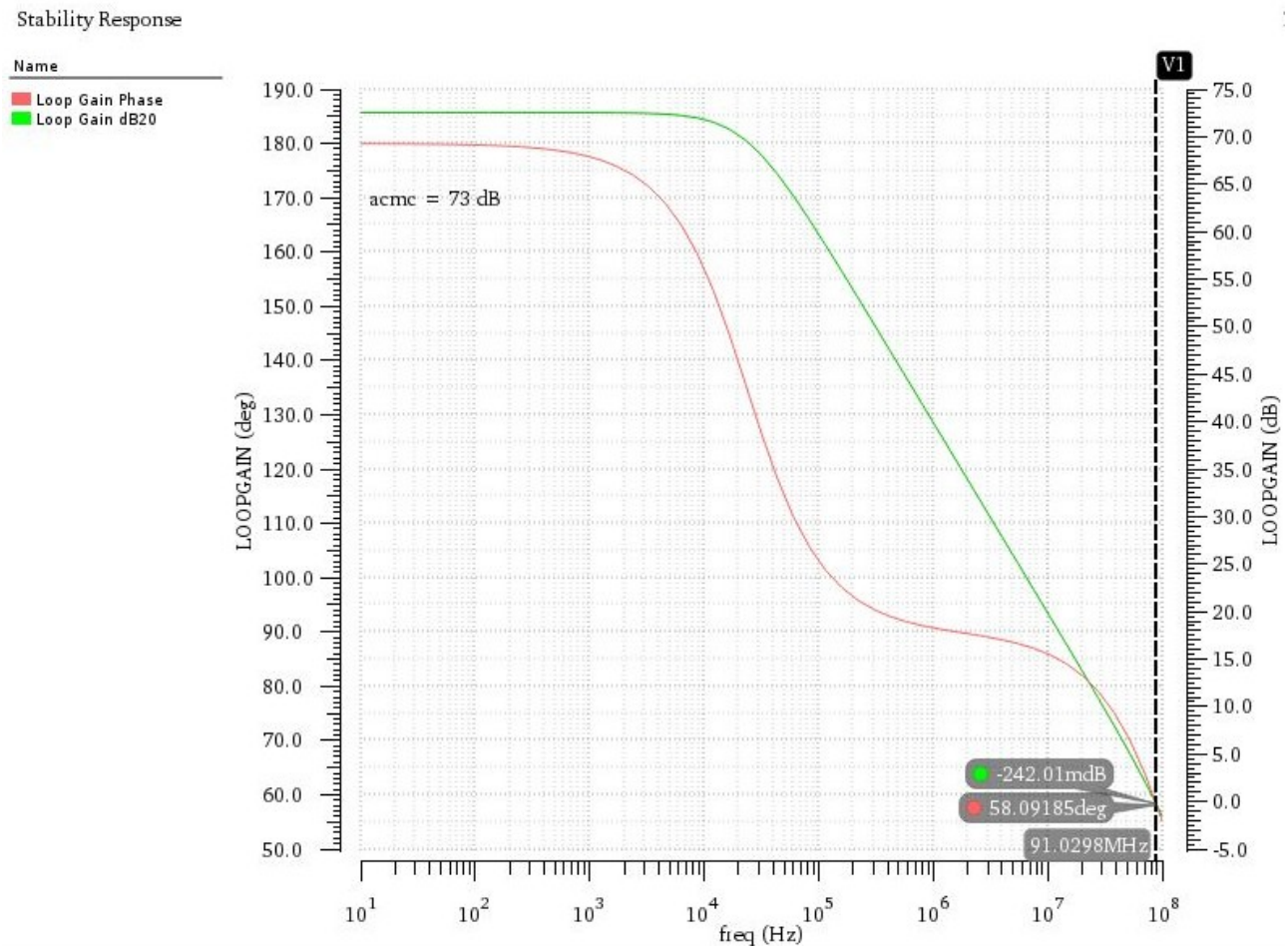


*Figure 11: The CMFB loop-gain of the amplifier*

Note that the CMFB loop is overcompensated due to the higher impedances at the common nodes between CM and DM paths.

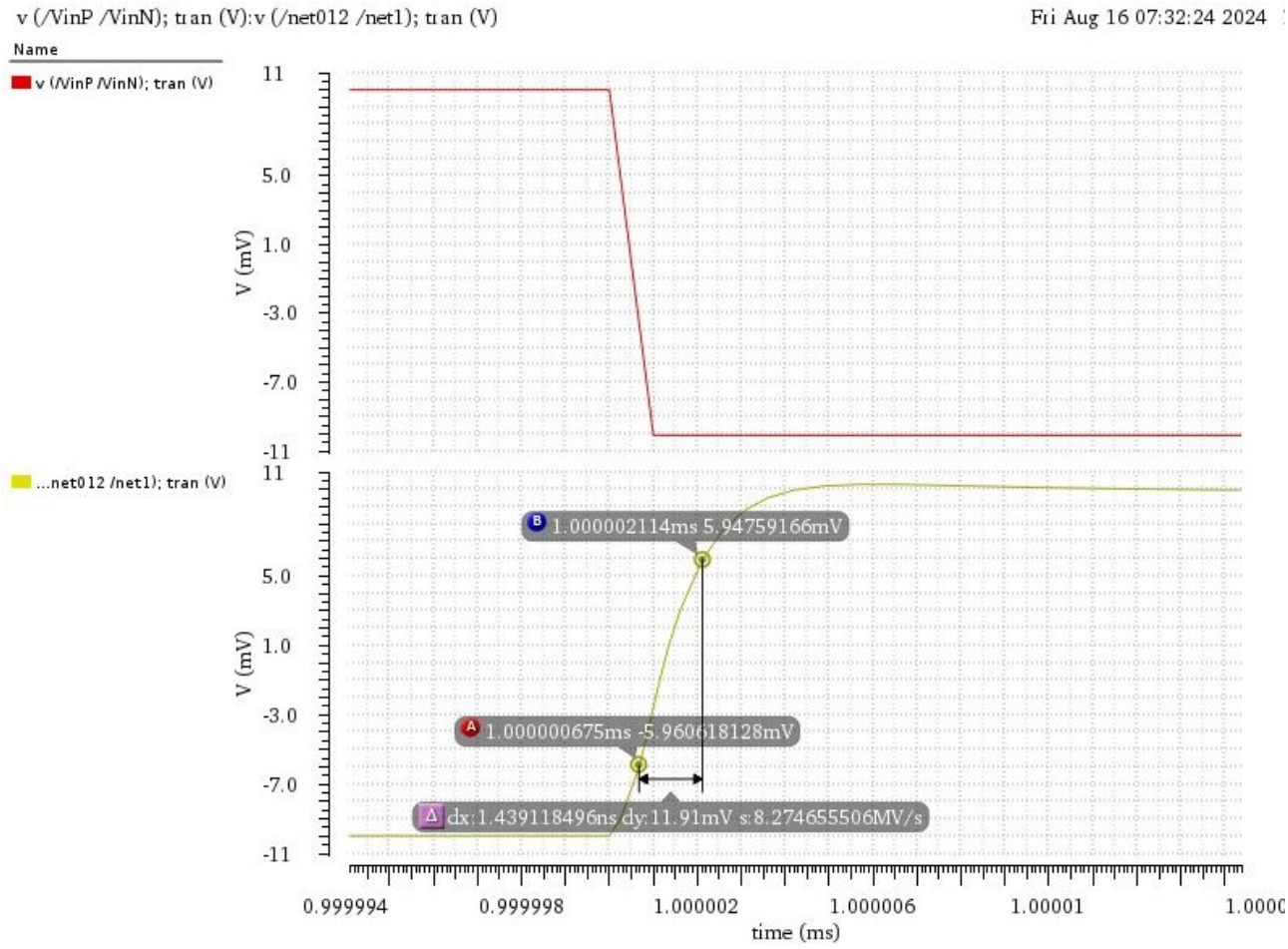


*Figure 12: The SR for a negative variation of th einput voltage*

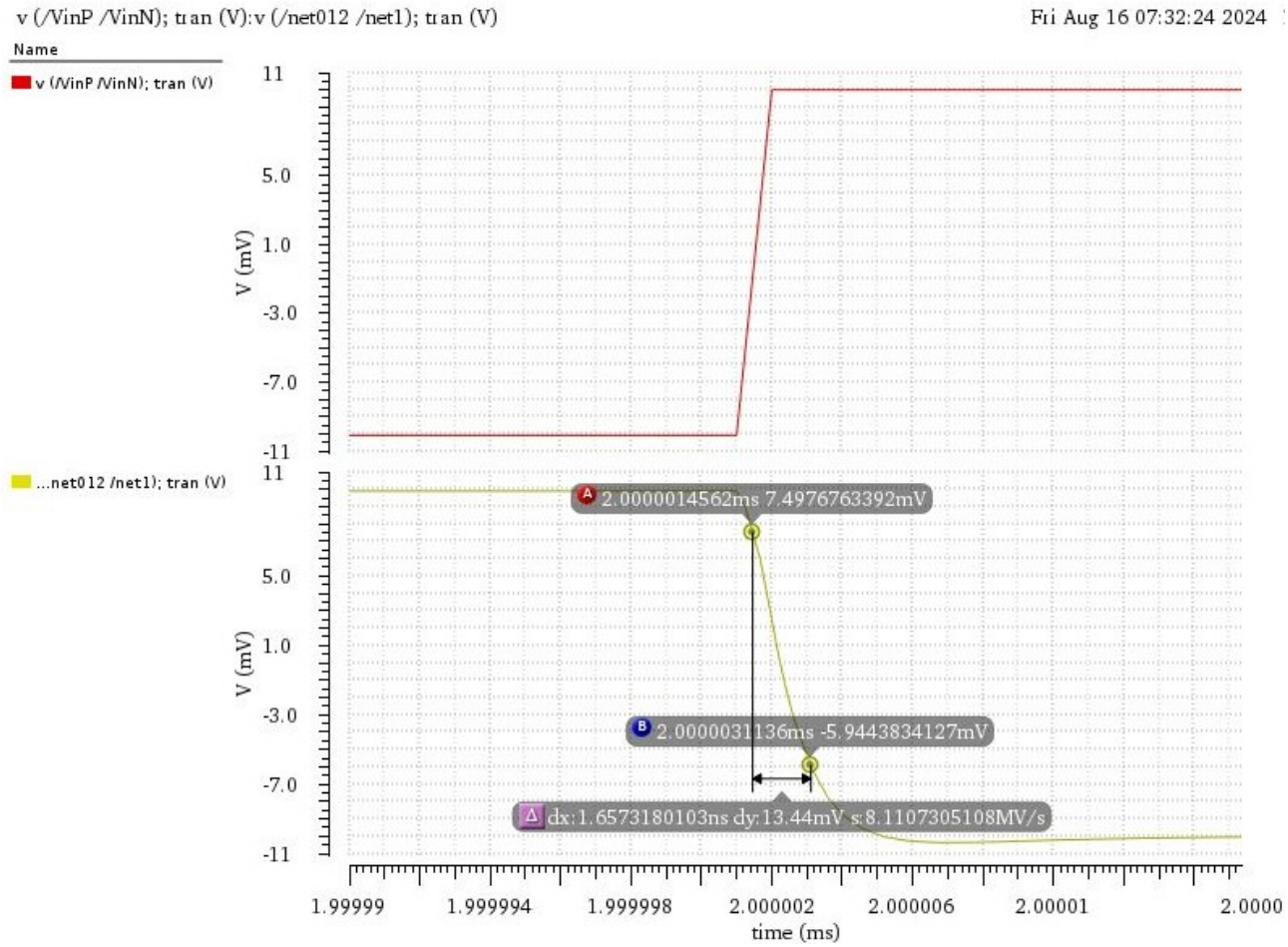


*Figure 13: The SR for a positive variation of the input voltage*

→ ~~this AO cannot drive resistive loads~~

DM path:

| A0 [dB] | GBW [Hz] | PM[°] | IDD [A] | IDD_CMFB [A] | SR_RISE[V/us] | SR_FALL |
|---|---|---|---|---|---|---|
| 68 | 143e6 | 62 | 173e-6 | 44e-6 | 8.27 | 8.11 |

*Table 7: The main parameters of the AO*

| $V_{DD}$[V] | $V_{ICMmin}$ [V] | $V_{ICMmax}$ [V] | $V_{OCMmin}$ [V] | $V_{OCMmax}$ [V] | $V_{ICM\_MEDIU}$ [V] | $V_{OCM\_MEDIU}$ [V] |
|---|---|---|---|---|---|---|
| 1.8 | 0.8775 | 1.76 | 0.107 | 1.67 | 1.31 | 0.885 |

## 2.2.2 Verifying and characterizing the LPF as a functinal block

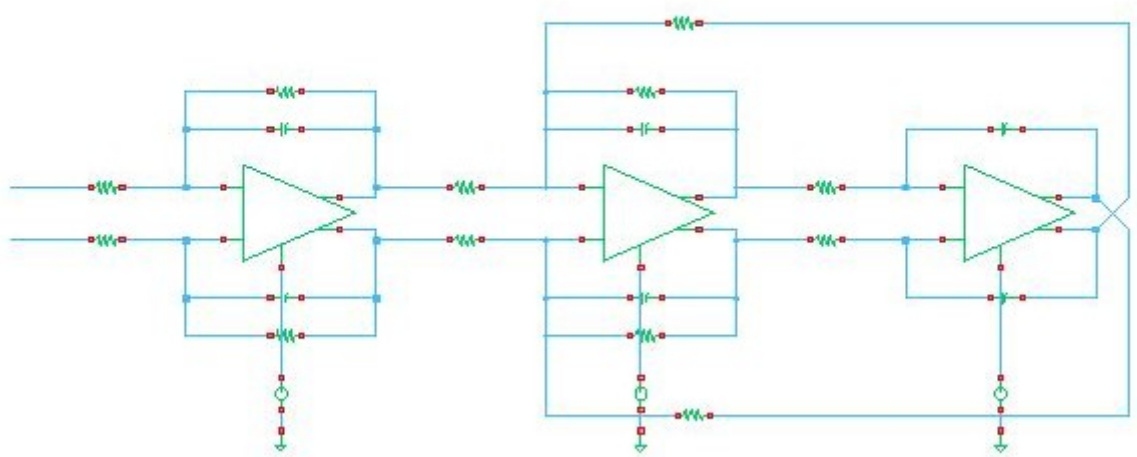

*Figure 14: The implementation of the 3rd order chebyshev LPF*

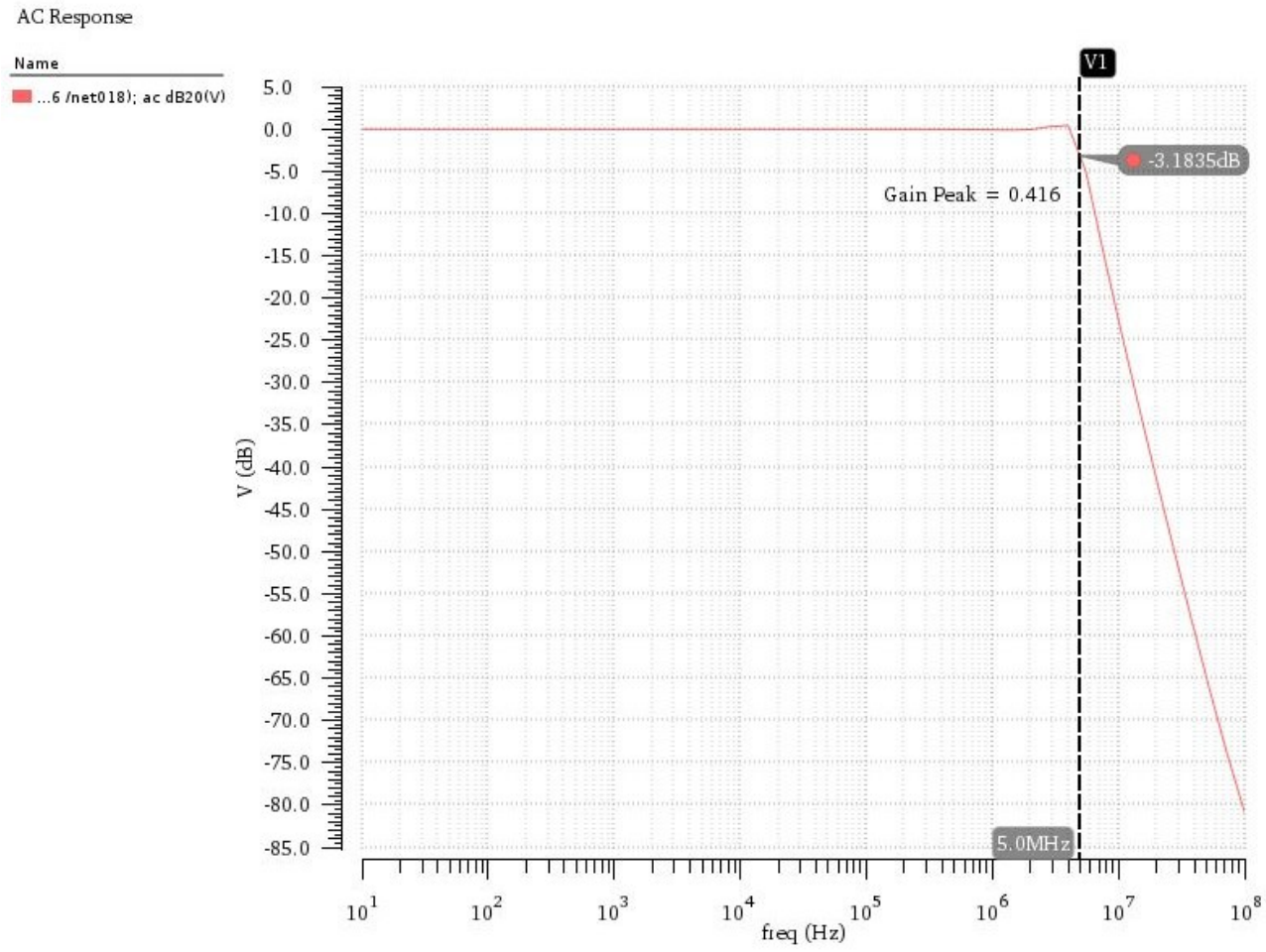


*Figure 15: The frequency characteristic of the filter*

# 3. PGA

## 3.1 Sizing the PGA

### 3.1.1 Sizing the passive components of the PGA

Requirements: different gain steps (coarse + fine), 0 dB – 20 dB, $BW_{\text{PGA}} = 20\,MHz$, CL = 5 pF, linear input of at least 50 mV

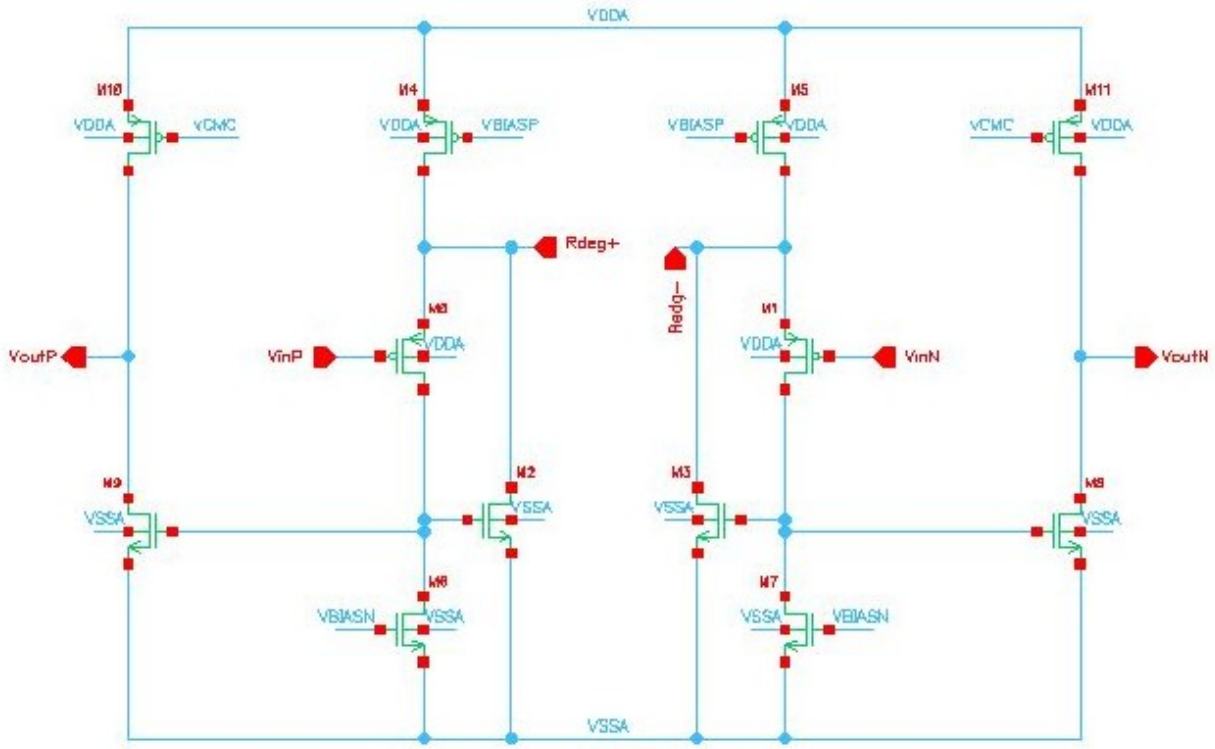


*Figure 16: Kwan-Martin transconductor*

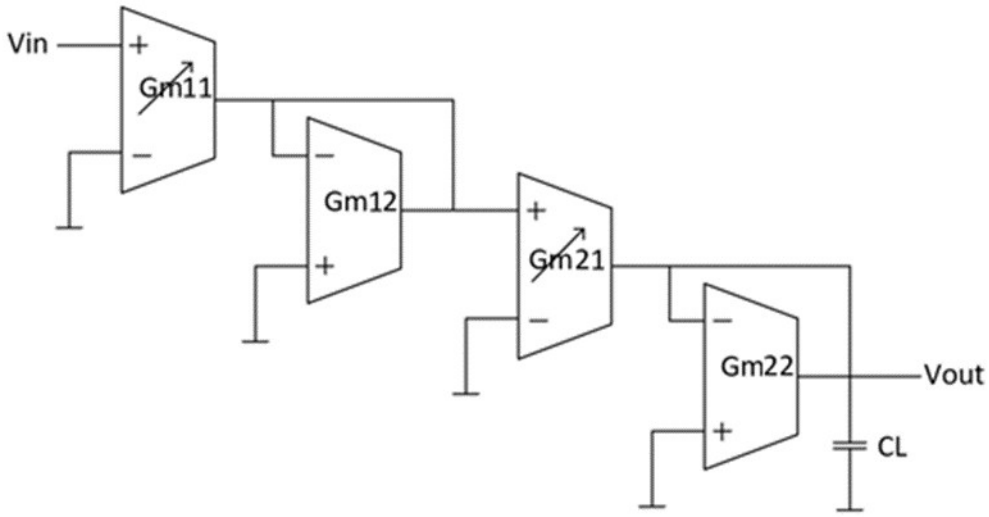


*Figure 17: The schematic of the PGA*

Where G_m11 and G_m12 realize the first stage and G_m21 and G_m22 realize the second stage.

The bandwidth of the PGA is given by:

$$BW_{PGA} = \frac{G_{m22}}{2\pi C_L} \tag{30}$$

From the previous equation we can obtain $G_{m22}$ = 628 uS.

If we consider $G_{m12} = G_{m22}$, then we can compute the values for $G_{m11} = A_{v1} G_{m12}$, and $G_{m21} = A_{v2} G_{m22}$.

| Av1 [dB] | Av1 [V/V] | Gm12 [S] | Gm11[S] |
|---|---|---|---|
| 0 | 1 | 6.28E-04 | 6.28E-04 |
| 6 | 1.995262 | 6.28E-04 | 1.25E-03 |
| 12 | 3.981072 | 6.28E-04 | 2.50E-03 |
| 18 | 7.943282 | 6.28E-04 | 4.99E-03 |

| Av2 [dB] | Av2 [V/V] | Gm22 [S] | Gm21[S] |
|---|---|---|---|
| 0 | 1 | 6.28E-04 | 6.28E-04 |
| 1 | 1.122018 | 6.28E-04 | 7.05E-04 |
| 2 | 1.258925 | 6.28E-04 | 7.91E-04 |
| 3 | 1.412538 | 6.28E-04 | 8.87E-04 |
| 4 | 1.584893 | 6.28E-04 | 9.95E-04 |
| 5 | 1.778279 | 6.28E-04 | 1.12E-03 |

Each transconductor is implemented with a Kwan-Martin structure with a variable degeneration resistance. The transconductance is computed as:

$$G_m = \frac{A_{ii}}{Rdeg} \tag{31}$$

For a current gain A_ii=4 we obtain the following values for the degeneration resistances.

| Av1 [dB] | Rdeg12 | Rdeg11 |
|---|---|---|
| 0 | 6.37E+03 | 6.37E+03 |
| 6 | 6.37E+03 | 3.19E+03 |
| 12 | 6.37E+03 | 1.60E+03 |
| 18 | 6.37E+03 | 8.02E+02 |

| Av2 [dB] | Rdeg22 | Rdeg21 |
|---|---|---|
| 0 | 6.37E+03 | 6.37E+03 |
| 1 | 6.37E+03 | 5.68E+03 |
| 2 | 6.37E+03 | 5.06E+03 |
| 3 | 6.37E+03 | 4.51E+03 |
| 4 | 6.37E+03 | 4.02E+03 |
| 5 | 6.37E+03 | 3.58E+03 |

### 3.1.2 Designing the PGA at transistor level

For sizing the transistor of each Gm cell I used the following design strategy.

1. Having already obtained the values of the degeneration resistance and knowing the required input range of the transconductor I computed the maximum degeneration current as

$$i_{\text{deg\_max}} = \frac{50\text{e-}3}{R_{\text{deg\_min}}} = 62\text{e-}6 \tag{32}$$

2. To ensure that all transistors remain in conduction we have to ensure that M_(2,3) have a drain current large enough.

$$I_{\text{DM2,3}} > 1.2\, i_{\text{deg\_max}} = 74\text{e-}6\, A \tag{33}$$

3. We can choose V_(dsatM2,3)=150mV

4. Note that the other transconductors could have been designed for lower bias currents due to higher R_(deg_min). If M_(2,3) in the actual implementation were realized using 2 cascodes, the same parameters can be used for the cascode transistors.

5. The current through M_(0,1) I chose as 100μA and V_(dsatM0,1)=150mV

6. Considering that previously I chose A_ii=4 it results that the current through M_(8,9) and M_(10,11) is 2i_(dM2,3)=296μA. V_(dsatM8.9) was chosen 0.150 V and V_(dsatM10,11) was chosen as 182mV.

7. If M_(8,9) and M_(10,11) in the actual implementation were realized using 2 cascodes, the same parameters can be used for the cascode transistors.

8. For M6,7 current sources their bias currents are the same as for M0,1 of 100 uA. While Vdsat = 150 mV. For M4,5 current sources its current value is equal to $I_{DM0,1}+I_{DM2,3}=100\,uA+74\,uA=174\,uA$, while Vdsat = 150 mV.

After applying the sizing strategy I obtained the following results for the transistor dimensions.

| | ID | VDSAT | gm | W[m] | L[m] | |
|---|---|---|---|---|---|---|
| M2,3 | 7.40E-05 | 0.15 | 9.87E-04 | 2.49E-06 | 2.00E-07 | NMOS |
| M0,1 | 1.00E-04 | 0.15 | 1.33E-03 | 3.07E-05 | 2.50E-07 | PMOS |
| M6,7 | 1.00E-04 | 0.165 | 1.21E-03 | 2.78E-06 | 2.00E-07 | NMOS |
| M8,9 | 2.96E-04 | 0.15 | 3.95E-03 | 9.95E-06 | 2.00E-07 | NMOS |
| M10,11 | 2.96E-04 | 0.182 | 3.25E-03 | 4.93E-05 | 2.00E-07 | PMOS |
| M4,5 | 1.74E-04 | 0.182 | 1.91E-03 | 2.90E-05 | 2.00E-07 | PMOS |

## 3.2 Verifying and characterising the PGA

### 3.2.1 Verifying and characterizing the transconductor

I used the previously obtained dimensions to implemement the Kwan-Martin Gm cell. Since the phase margin was not large enough I realised a dominant pole compensation to reduce the corresponding gain peaking by placing a capacitor in the drain of the input transistors.

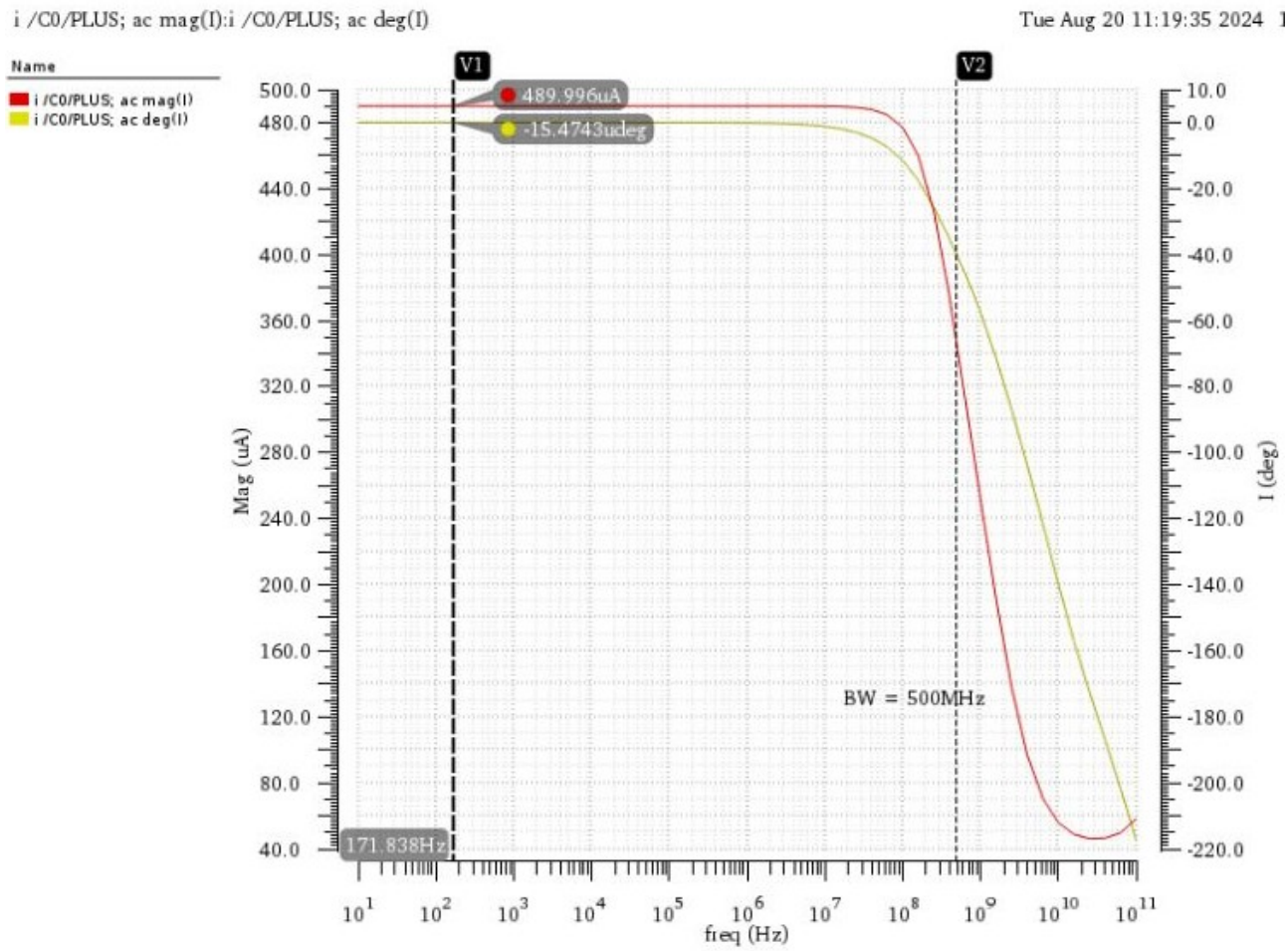


*Figure 18: The minimum transconductance necessary*

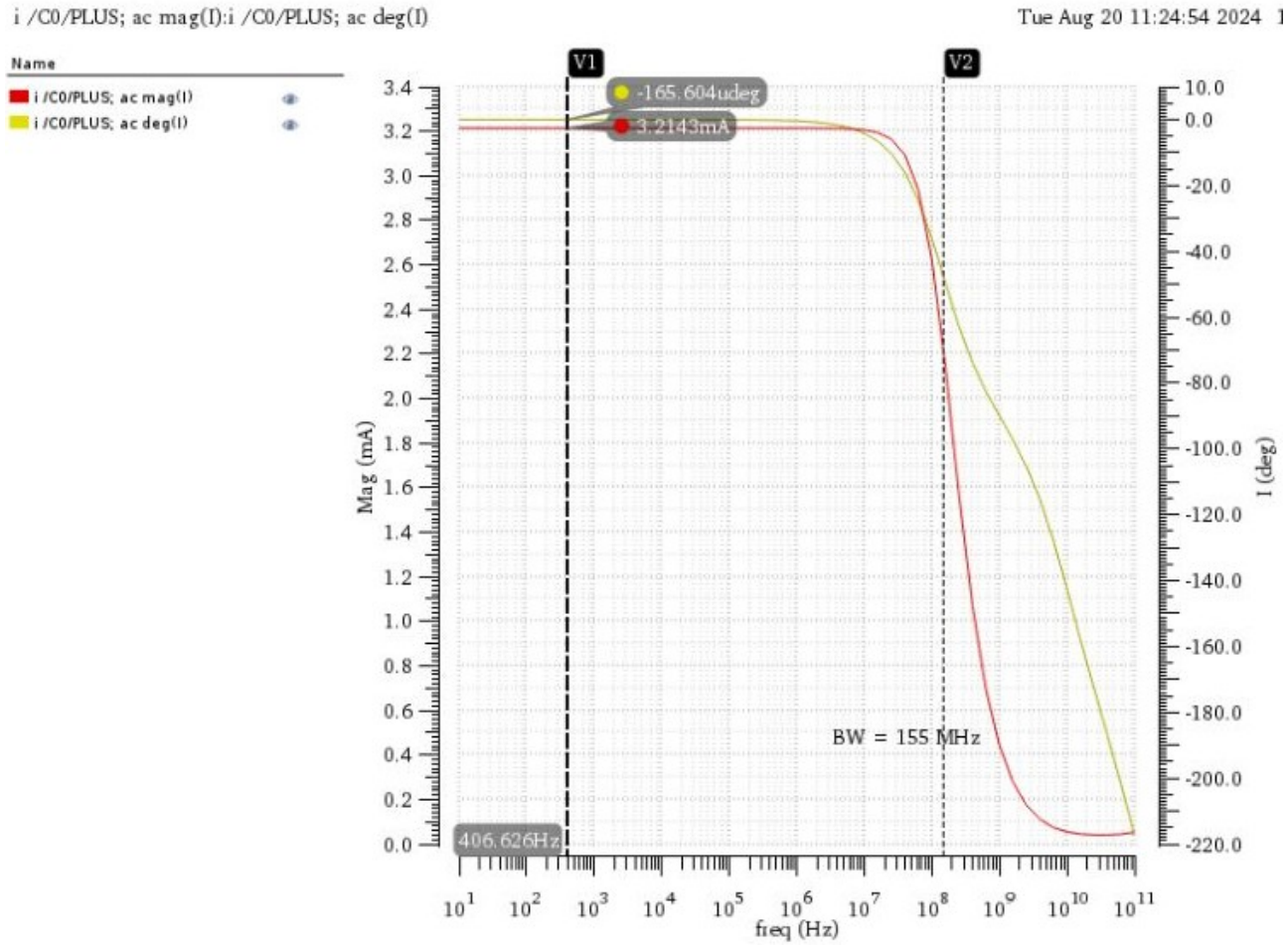


*Figure 19: The maximum transconductance necessary*

Note that while the original project [1] used cascode's, mine hasn't. Because my output resistance which is around 10k Ohms is much lower than hers the Gm also have errors from the nominal values. The output CM voltage also influence Gm.

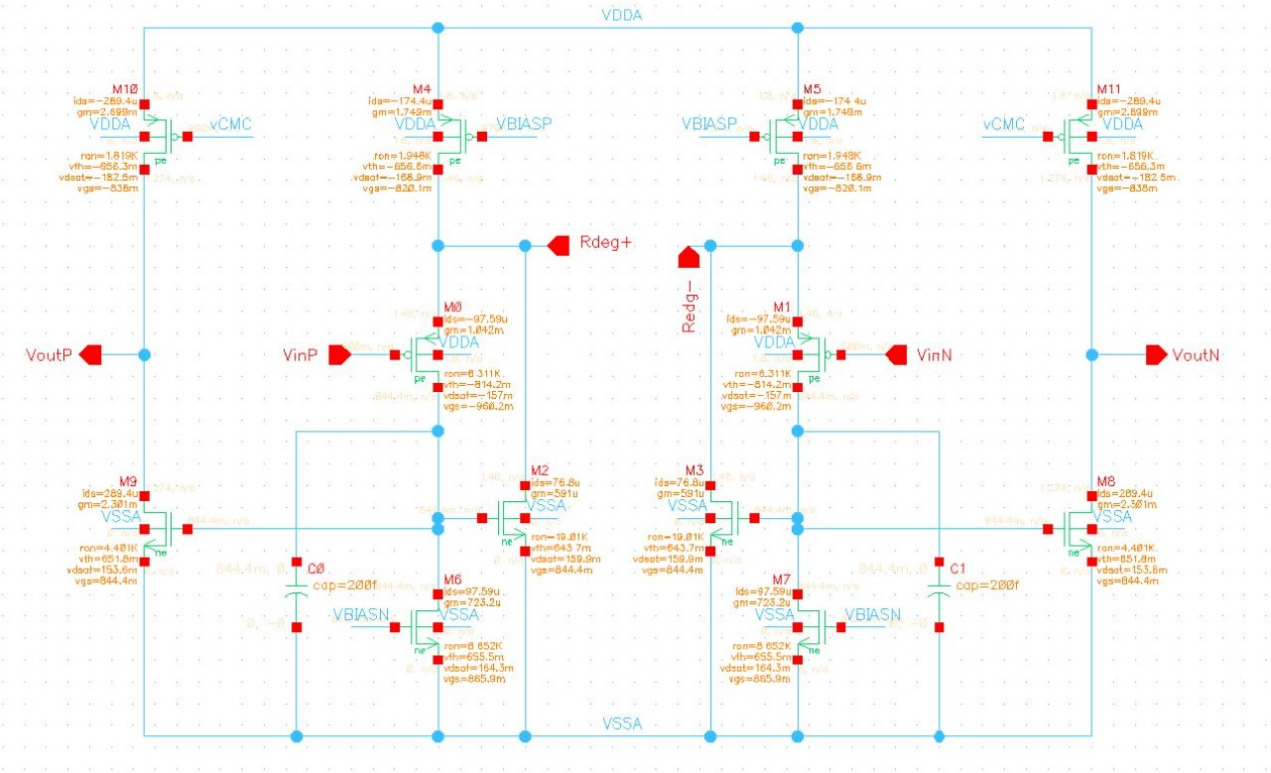


*Figure 20: The schematic of the Gm cell and the DC operating points of the transistors*

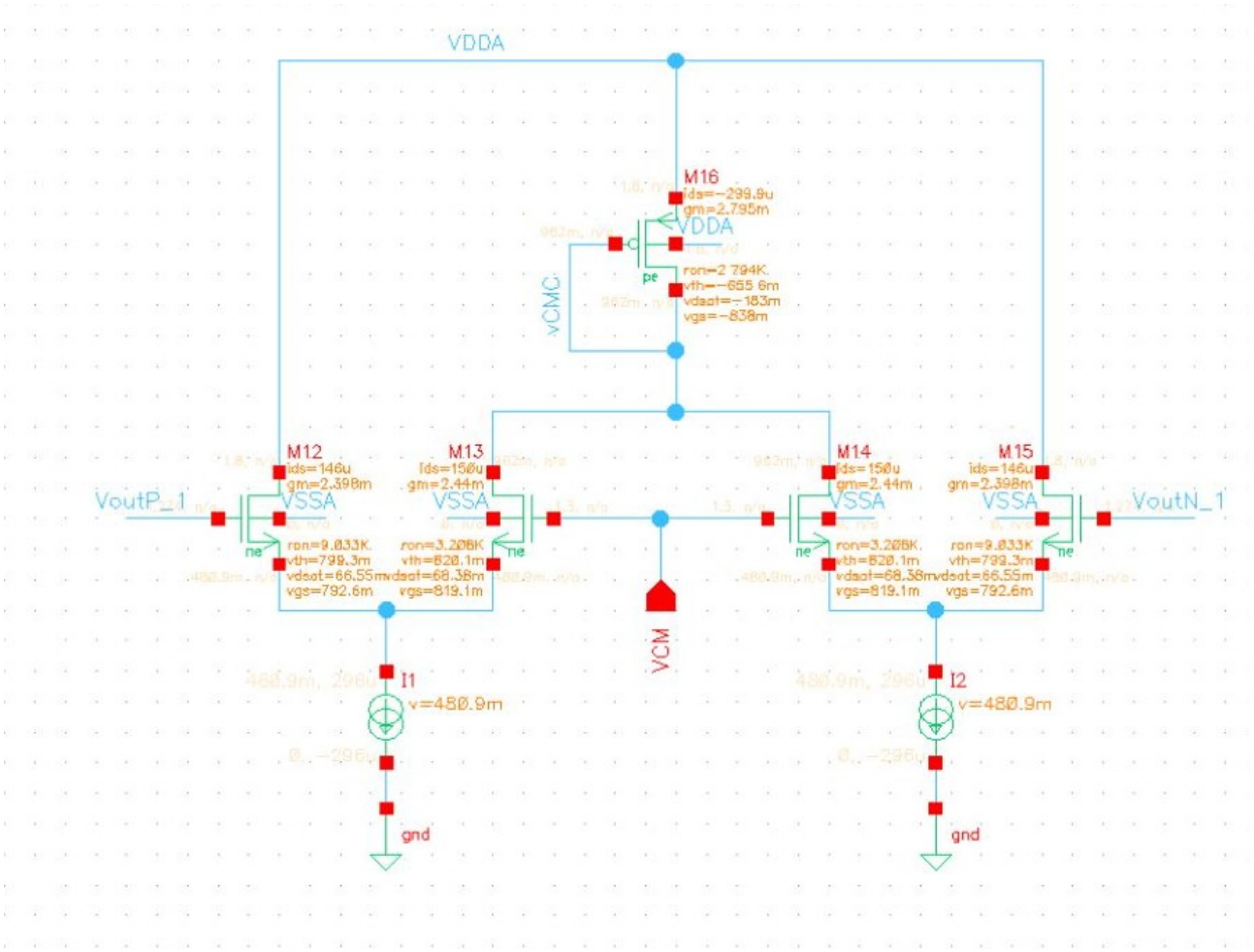


*Figure 21: CMFB circuit*

For Rdeg 6.37k, and Gm of 3.21 mS, an input signal of 1 mV with a frequency of 500kHz the THD is 0.397 %, and for an input signal of 50 mV and frequency of 500 kHz for the same Gm, the THD is 0.720 %.

Note that I use a smaller linear input range as the original requirement of 400 mV would have required too large bias currents, and transistor dimensions. (see eq. 20)

| VDD[V] | $V_{ICM\_MIN}$[V] | $V_{ICM\_MAX}$[V] | $V_{OCM\_MIN}$[V] | $V_{OCM\_MAX}$[V] | $V_{ICM\_MED}$[V] | $V_{OCM\_MED}$[V] |
|---|---|---|---|---|---|---|
| 1.8 | 0.041 | 0.671 | 0.153 | 1.618 | 0.356 | 0.885 |

### 3.2.2 Verifying and characterizing the PGA as a functional block

// I had to modify $M8,9\,had\,\frac{W}{L}=\frac{10u}{200n}\,with\,r_{ds}=11k \rightarrow \frac{W}{L}=\frac{44u}{900n}\,with\,r_{ds}=98k$ to increase the output resistance, for a better Gm reproducability. $M10,11\,have\,r_{ds}=50k$.

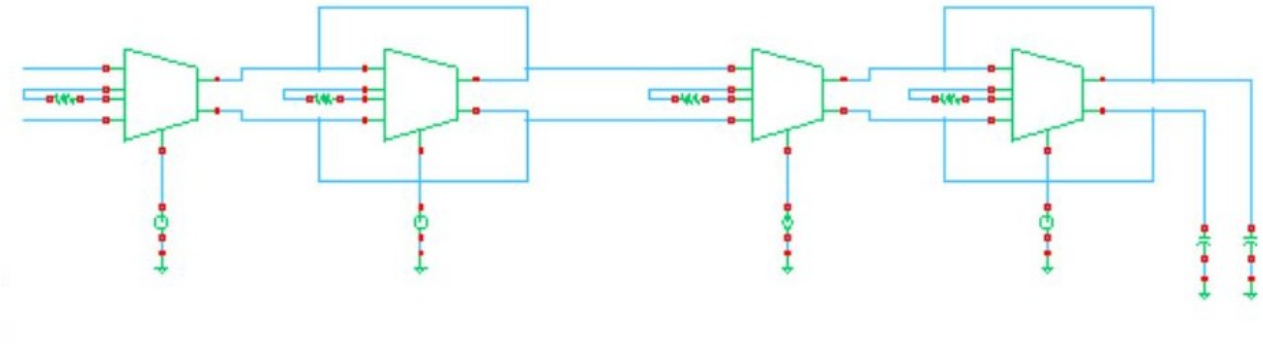

*Figure 22: The implementation of the PGA*

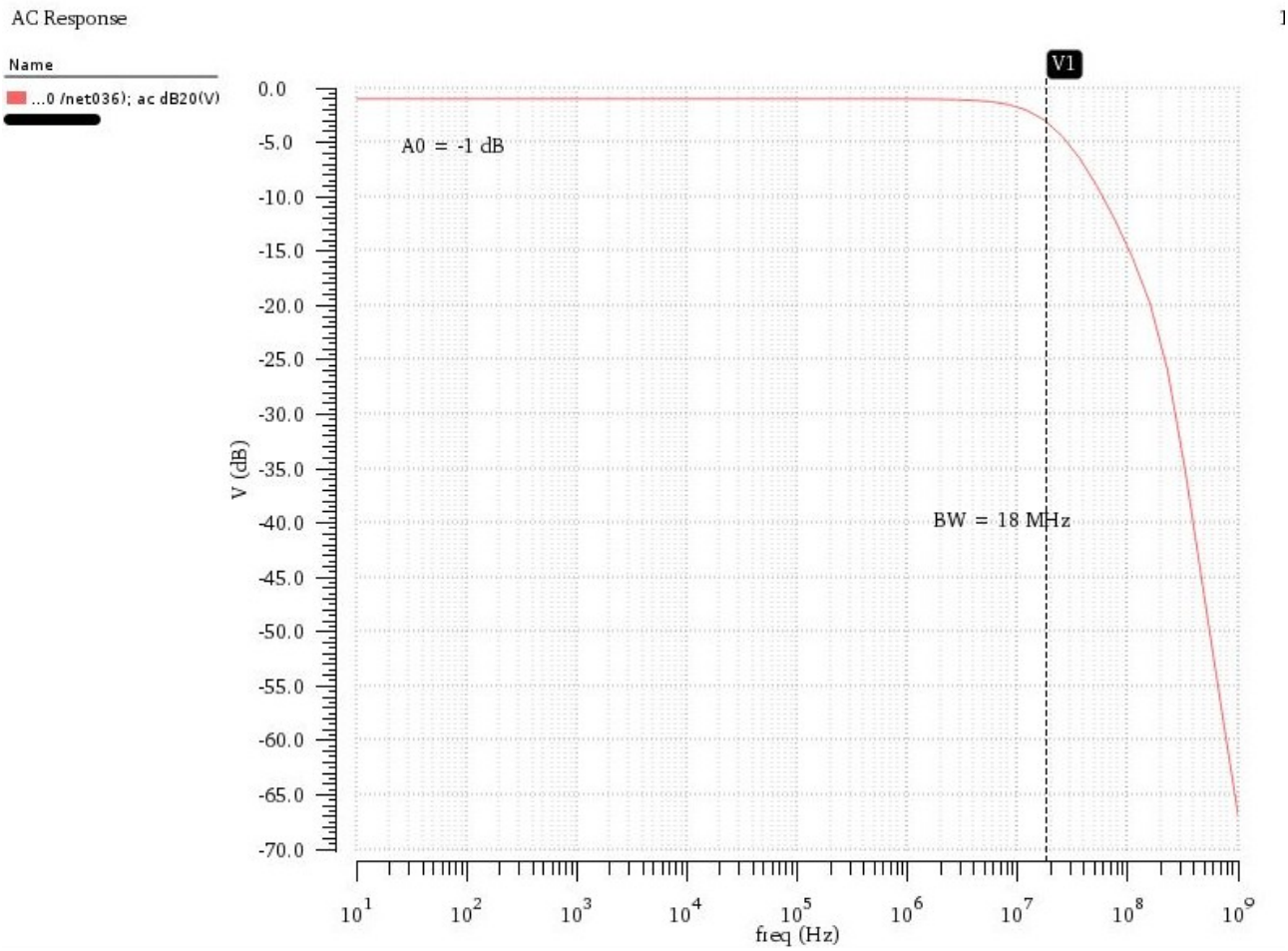


*Figure 23: The frequency characteristic for the minimum gain*

Here we can see that instead of 0 dB I am getting -1 dB. This could be an isue with the CM output voltage being too low for all transconductors. IF VCM is 0.4 V then the error is -1.15 dB.

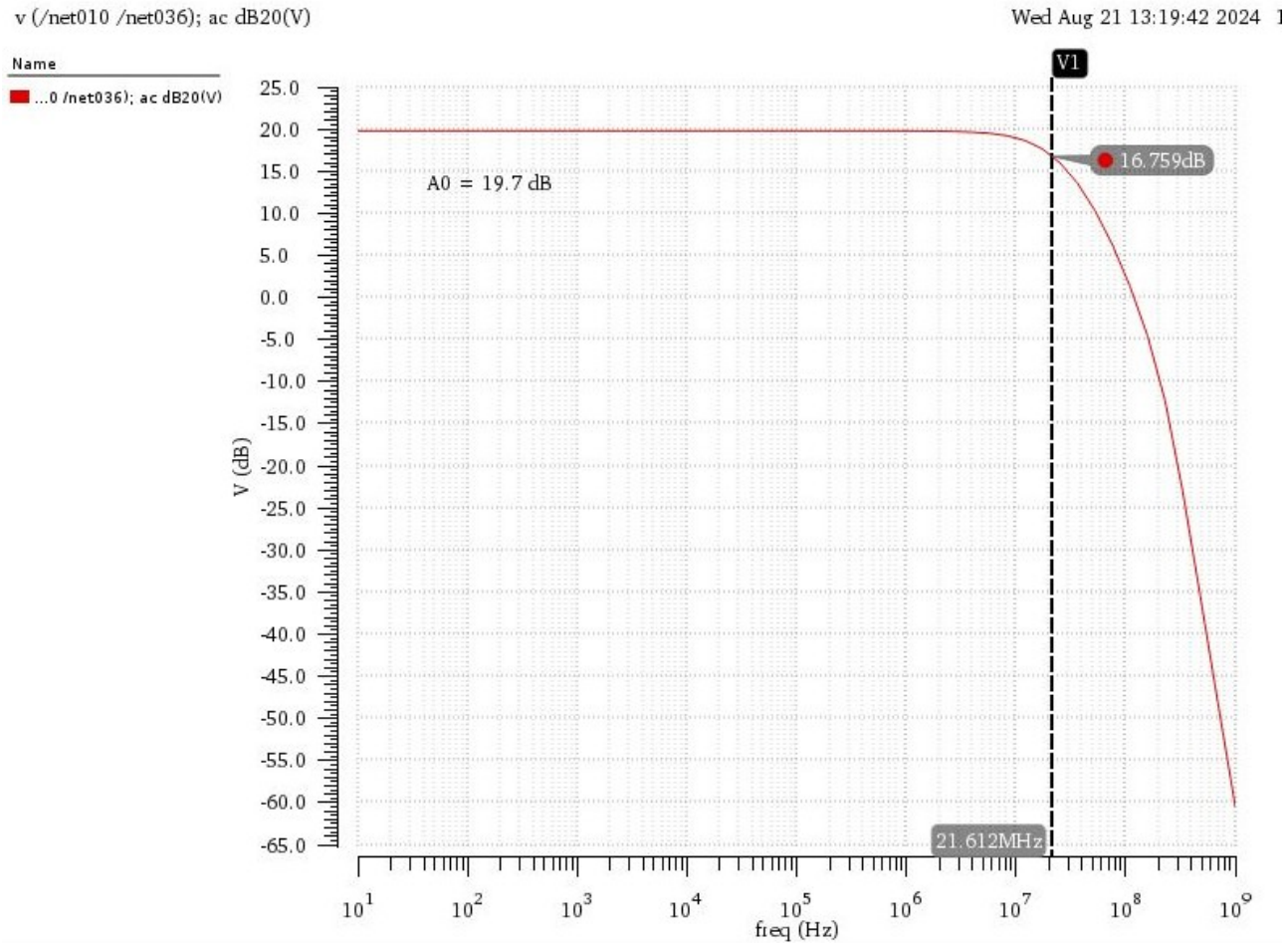


*Figure 24: The frequency characteristic for the maximum gain*

# 4. Verifying and characterizing LPF + PGA

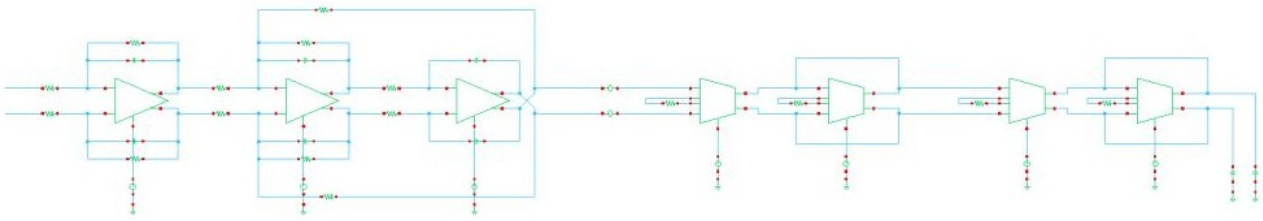

*Figure 25: The schematic implementation*

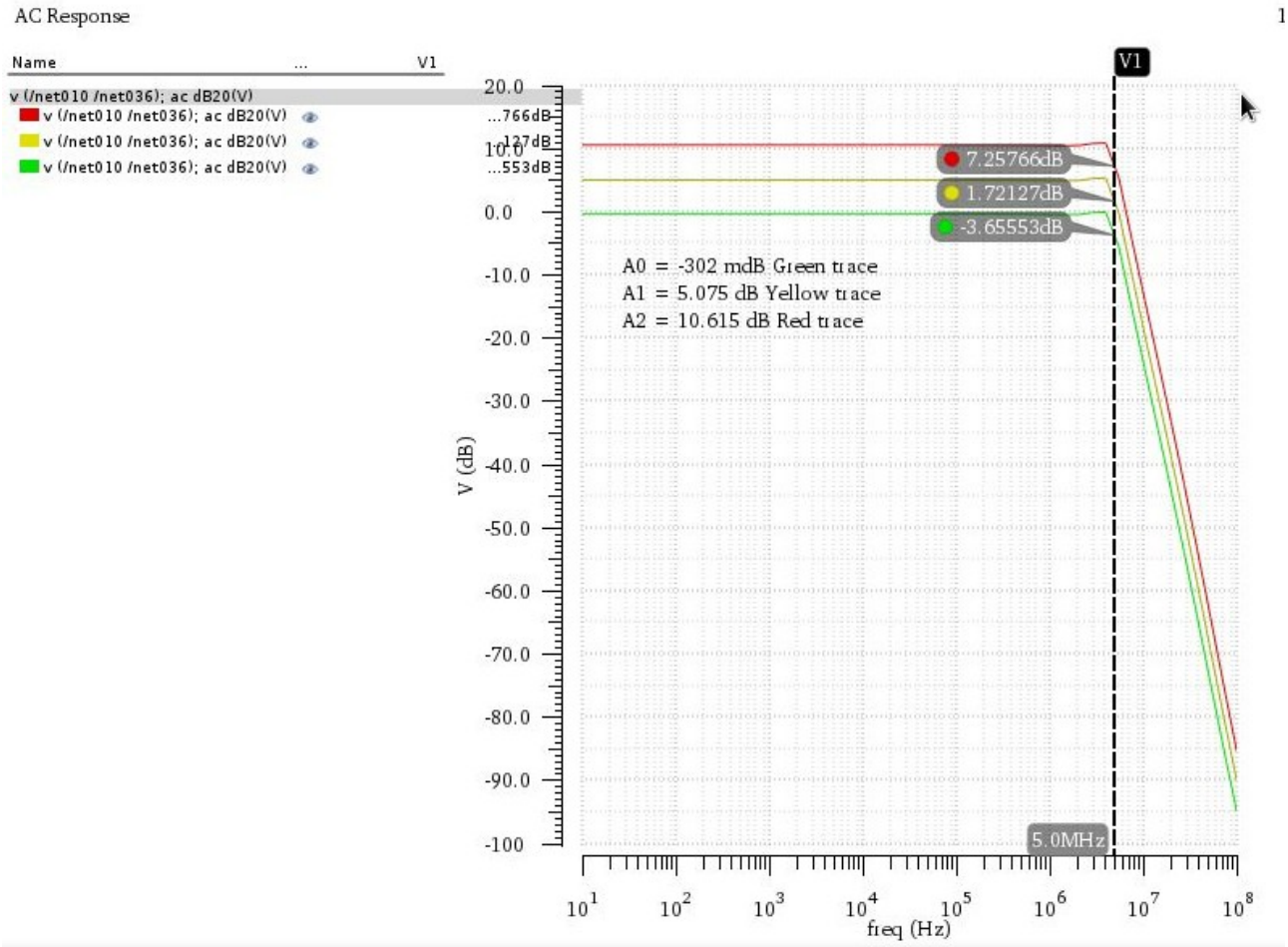


*Figure 26: Frequency characteristic*

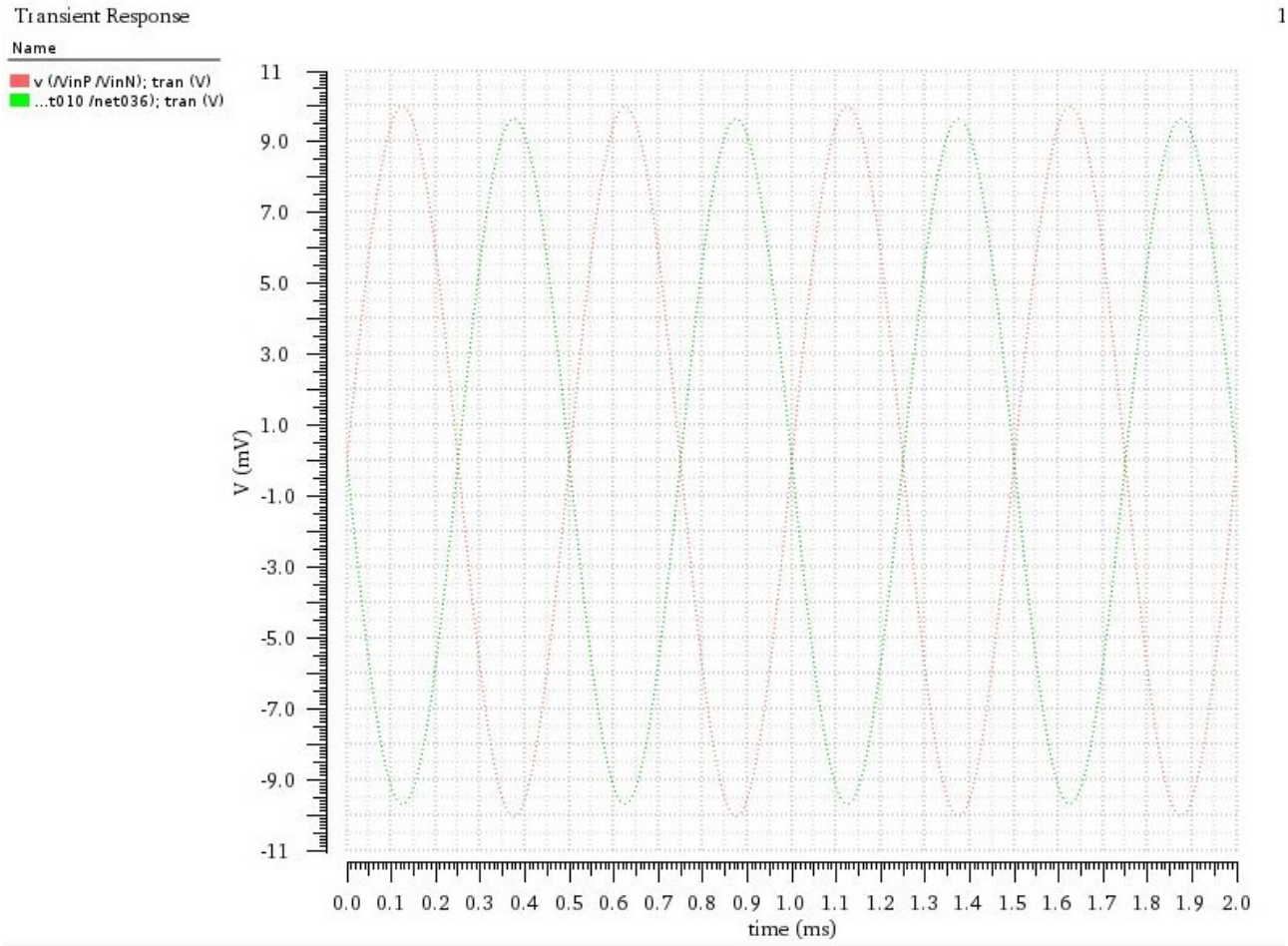


*Figure 27: Transient response for minimum gain*

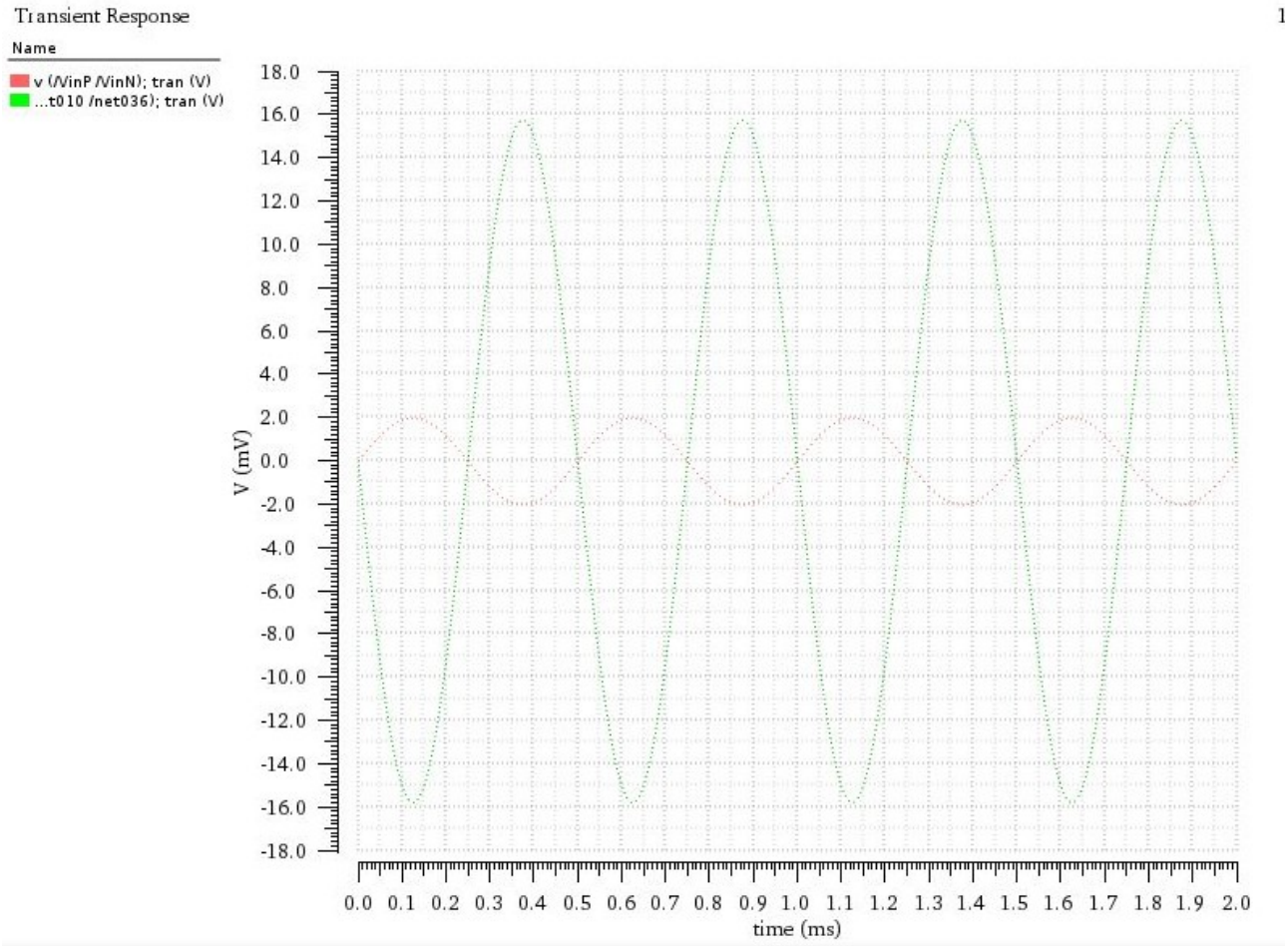


*Figure 28: Transient response for maximum gain*

# 5. LPF-AO Miller Modifications

Because of the initial CMFB circuit used in AO Miller to control the tail current source, seemingly the circuit would struggle to find the DC operating point with a feedback circuit. In this regard the CMF has been changed to drive the active loads of the differential pair.

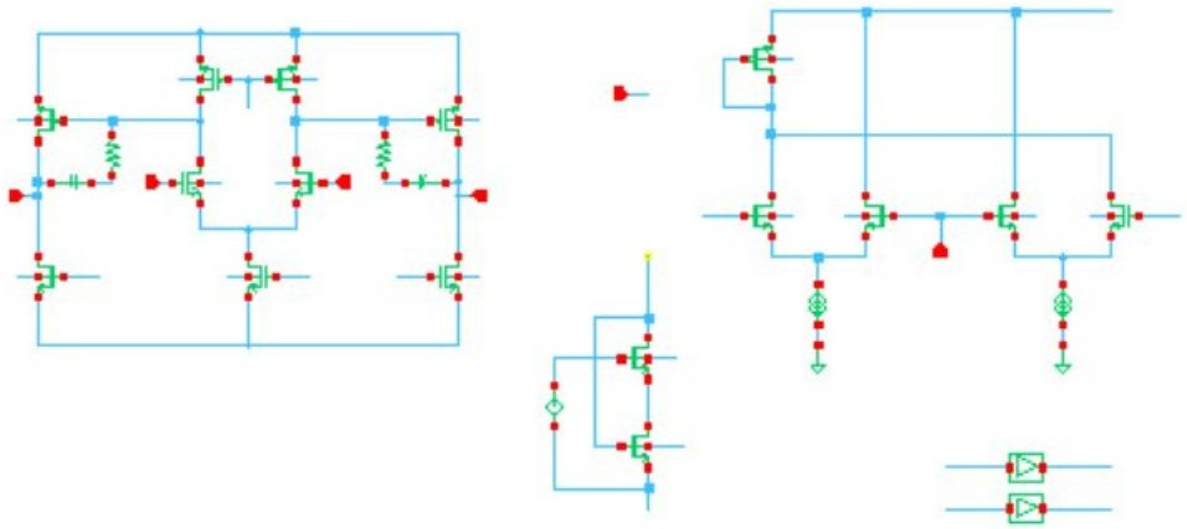

*Figure 29: AO Miller and CMFB*

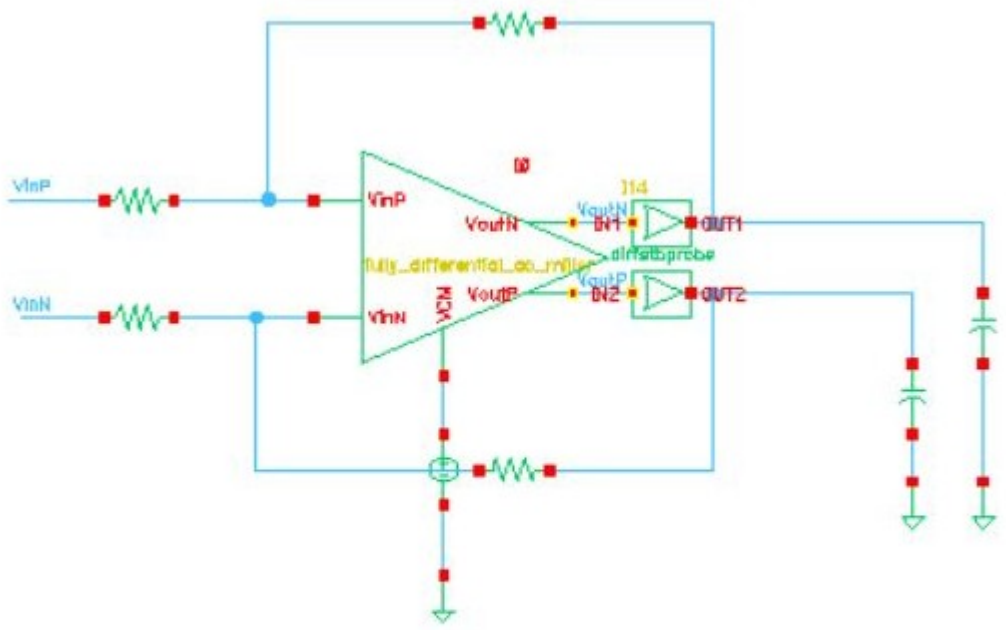


*Figure 30: AO Miller TB*

# Bibliography

[1]