

# Design and Implementation of a Radiation Hardened High Output Current Transconductance Amplifier

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# Abstract



# Zusammenfassung



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## Acknowledgments



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# 1 Introduction

Over the last ten years, the electronics industry has exploded. The largest portion of the total worldwide sales is dominated by the MOS market. Composed primarily of memory, micro and logic sales, the total combined MOS revenue contributed approximately 75 percent of the worldwide sales, illustrating the strength of CMOS technology. CMOS technology currently continues to mature, with minimum feature sizes now approaching 10nm. This has resulted in development in process technologies and have allowed design of circuits using devices running at lower currents, lower voltages and at higher unity gain frequencies. This has done signal processing world a great deal of good, resulting in creation of amplifiers with high speed and converters operating under low voltage conditions.

The necessity for high voltage integrated circuits, capable of driving high currents, remains greater than ever before, with countless applications related to emerging technologies. There are numerous applications of high current amplifiers which can be used in any signal processing context involving driving loads which require high current. One of the common applications is in the Current Transformer circuits.

Operational Transconductance Amplifier(OTA) is a building block for most of the analog circuits with linear input-output characteristics. Operational Amplifiers (OP AMP) too are used as a basic building block in implementing a variety of analog applications such as amplifiers, summers, differentiators, integrators, etc to a more complicated applications like Oscillators.

For more than 50 years, silicon-on-insulator (SOI) technologies have been developed for radiation-hardened military and space applications. The use of SOI has been motivated by the full dielectric isolation of individual transistors, which prevents latchup. Many efforts have been made to reduce these parasitic structures and very high levels of radiation hardness have been achieved. In addition, major chip manufacturers are now producing SOI technologies for many nonhardened high performance or low-power applications.

This document describes the design and implementation of such a radiation hard transconductance amplifier, which is used in a DC current transformer. The report is structured as follows: First, the concept of DC Current Transformer is introduced. Then, the theory of the building blocks of the design is provided. After that, the design and implementation of these building blocks which was done in Cadence Virtuoso environment is documented. And then the simulation results of the overall design and the conclusion of the thesis work are documented in the subsequent chapters of this document.

## 1.1 Novel DC Current Transformer

DC Current Transformers (DCCTs) are known as non-intercepting standard tools for online beam current measurement in synchrotrons and storage rings. In general, the measurement principle of commonly used DCCTs is to introduce a modulating AC signal for a pair of ferromagnetic toroid. Currently a Novel DCCT (NDCCT) based on modern magnetic field sensors is under development at GSI. The closed-loop NDCCT Design can be considered as an extension to the primary design proposed by GSI. A feedback loop was designed to improve the linearity and sensitivity of the NDCCT. Figure 1.1 shows the closed loop NDCCT structure. The main objective is to provide a feedback current that cancels the magnetic field of the ion-beam inside the air gap. This measurement principle is referred to as "Zero-flux" and it is currently used in commercial DCCT.

The change in magnetic field will be sensed by the TMR (Tunneling Magneto Resistance) sensor and it produces a small voltage. This voltage is amplified by a Voltage amplifier and is converted to a feedback current  $I_{FB}$  by the OTA. This current flows through the  $N$  turns winding wire terminated with a resistive load. And this is where the thesis work fits in. The main task of the OTA is to amplify a small AC voltage and produce a high current that drives a resistive load and cancels the magnetic field of the ion-beam.

As a proof of concept, this Closed Loop NDCCT was constructed with the commercial OTAs OPA860 and OPA861 with multiple of those ICs in parallel to get a good understanding of the requirements from the OTA in terms of specifications and results with a dual bipolar power supply of 5V and -5V.

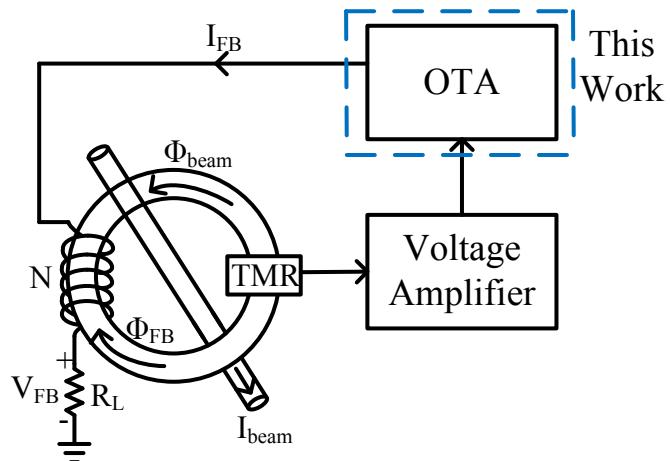


Figure 1.1.: Closed Loop Novel-DC Current Transformer

## 1.2 Specification

As the transconductance amplifier will be a part of the NDCCT, the specifications of the design were obtained from the commercial OTAs OPA860 and OPA861 which were used on a PCB with a power supply of 5V and -5V. Since the OTA is being targeted for multiple applications, it is important to make the design programmable. The specifications are tabulated in detail in Table 1.1.

Parameter	Value/Specification
Circuit Design	Programmable via I/V/external R or C
Transconductance Gain(Gm)	75 .. 140 mA/V
Linear Input Voltage Range	±200 mV
Output Current Range	±15 mA .. ±30 mA
Bandwidth	10 MHz
Slew Rate	±900 V/μs
Rise/Fall time	4.4 ns
Input Referred Noise	3 nV/√Hz @few KHz
Input Impedance	0.5 MΩ
Output Impedance	55 KΩ
HD2	Less than -75 dBc
HD3	Less than -80 dBc
Open Loop Voltage Gain	Not less than +5 V/V
PSRR	±20 μA/V

Table 1.1.: Specifications of the Design

Remarks:

- Supply Voltage: 5V
- Max Load Resistance: 1kΩ
- Technology to be used: XT018 from XFAB

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### 1.3 Technology

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The technology used for designing the building blocks of the system is the XT018 technology. The XT018 series is XFAB's 0.18-micron Module High-voltage SOI CMOS Technology. It combines the benefits of SOI wafers with Deep Trench Isolation and those of a State of the art six metal layers 0.18-micron process. This platform is specifically designed for the next generation automotive, industrial and medical applications operating in the temperature range of -40<sup>0</sup>C to 170<sup>0</sup>C. The technology related component names used as part of the circuit design are as tabulated in Table.1.2.

Component	Name	Description
PMOS	pe5	5V PMOS
NMOS	ne5	5V NMOS
Resistor	rmtp	Top Metal Resistor
Capacitor	cmm4t	Single MIM Capacitor

Table 1.2.: Components used as part of XT018 technology

The reason for choosing XT018 to design the amplifier is because SOI technology makes the design radiation hard. Silicon on insulator (SOI) technology refers to the use of a layered silicon-insulator-silicon substrate in place of conventional silicon substrates in semiconductor manufacturing, especially microelectronics, to reduce parasitic device capacitance, thereby improving performance. SOI-based devices differ from conventional silicon-built devices in that the silicon junction is above an electrical insulator, typically silicon dioxide or sapphire.

On the other hand, radiation hardening is the act of making electronic components and systems resistant to damage or malfunctions caused by ionizing radiation (particle radiation and high-energy electromagnetic radiation), such as those encountered in outer space and high-altitude flight, around nuclear reactors and particle accelerators, or during nuclear accidents or nuclear warfare. Most semiconductor electronic components are susceptible to radiation damage; radiation-hardened components are based on their non-hardened equivalents, with some design and manufacturing variations that reduce the susceptibility to radiation damage. Due to the extensive development and testing required to produce a radiation-tolerant design of a microelectronic chip, radiation-hardened chips tend to lag behind the most recent developments.

## 2 Theory

### 2.1 The Operational Transconductance Amplifier

The operational transconductance amplifier (OTA) is basically an op-amp without an output buffer. An OTA can be defined as an amplifier where all the nodes are low impedance except the input and output nodes. A schematic symbol of an OTA is shown in Figure.2.1. OTAs are generally programmable usually through the bias current that is provided to a differential amplifier. In the symbol, a bias voltage is being used as a programmable input. This bias voltage in turn controls the bias current of the amplifier. More on this later.

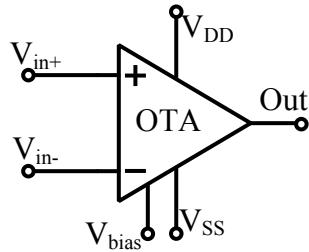


Figure 2.1.: Symbol of an OTA

#### 2.1.1 Conventional Current Mirror OTA

The schematic of a conventional current mirror based OTA is as shown in Figure.2.2. The OTA employs a differential input pair in conjunction with three current mirrors. The differential input pair comprises of transistors  $M_1$  and  $M_2$ . The differential pair is biased by the current mirror formed by  $M_{nB1}$  and  $M_{nB2}$ . PMOS current mirrors formed by  $M_3$ ,  $M_5$  and  $M_4$ ,  $M_6$  reflect currents generated in the differential pair onto the output shell. The current generated by the mirror formed by  $M_3$  and  $M_5$  is reflected to the output via the NMOS current mirror formed by  $M_7$  and  $M_8$ . The current mirror gain factor is  $K$ . Increase in  $K$  will increase the slew rate and gain bandwidth along with the output current of the conventional OTA at the cost of increased area, static power dissipation and a decrease in phase margin.

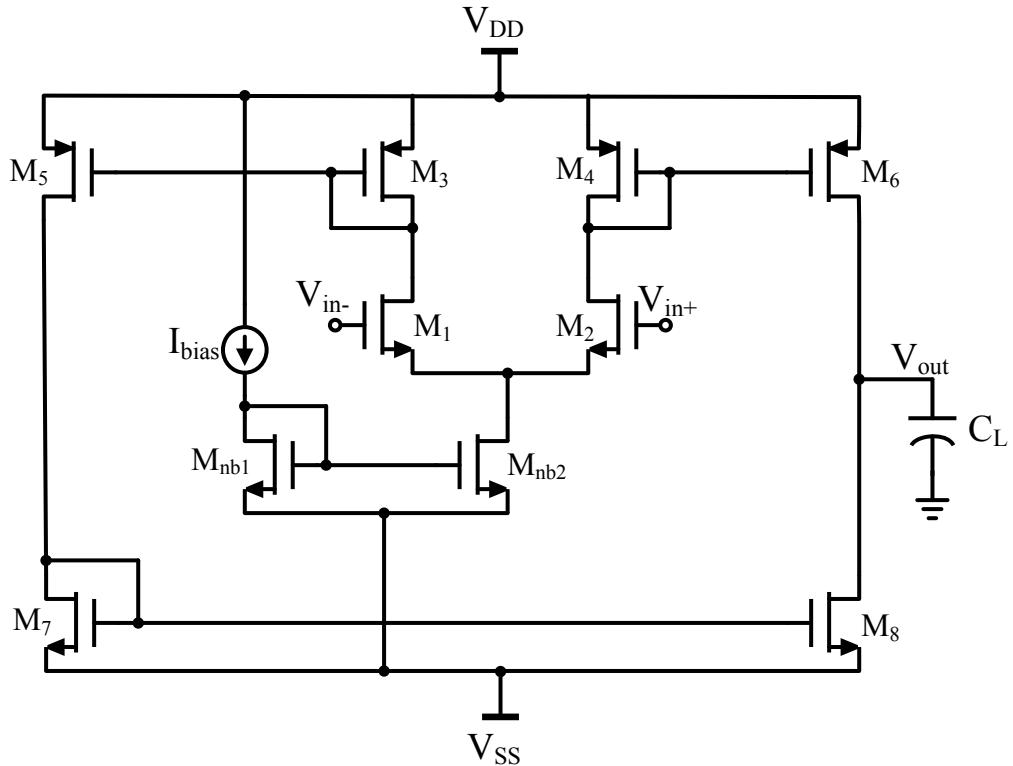


Figure 2.2.: Schematic of a Conventional Current Mirror OTA

Common mode signals are ideally rejected. i.e., when  $V_{in+} = V_{in-}$ ,  $I_{out} = 0$ . A differential input signal will generate an output current proportional to the applied differential voltage based on the transconductance of the differential pair. And that is why OTA is known as Voltage controlled Current Source. Although the output stage is push-pull structure, the conventional OTA is only capable for producing an output current with a maximum amplitude equal to the bias current in the output shell, provided the current mirror gain is unity, i.e.,  $K=1$ . For this reason, conventional OTA is referred to as a Class A amplifier.

The conventional OTA does not employ an output buffer and hence is only capable of driving capacitive loads and the capacitive load consumes all the current generated in the circuit. The gain of the OTA ( $G_m R_o$ ) is dependent on the large output resistance of the output shell of the OTA formed by the transistors  $M_3$  to  $M_8$ . The output resistance is reduced to  $G_m R_o // R_L \approx G_m R_L$  if a parallel resistive load  $R_L$  is applied.

Let us consider a hypothetical scenario that we do not need a resistive load to be driven as part of the system. In other words, we have a capacitive load as part of the specification. An advantage of this would be the fact that the need of an output buffer is eliminated and thereby the static power dissipation and the area occupied is reduced. But on the downside, the amplifier behaviour for the output current would be like that of a band pass filter. So this scenario will not be useful as the specification demands a bandwidth of 10MHz as the overall system would be targetting low frequency applications. So a capacitive load will limit the output current only to a certain band of frequencies and hence it will not make sense to just use a one stage design. This gives rise to design an output buffer for the OTA.

An output buffer can either be a voltage buffer or a current buffer depending on how the OTA is designed. A current buffer, for example, is used to hold the output current of the OTA at the output of the buffer. So any amplification needed for this current would in turn give rise to the need of another stage, making the design complicated. In this case usually a current feedback amplifiers are designed which theoretically has a low input impedance non-inverting terminal input and a high input impedance inverting terminal input. So the OTA output would directly be connected to this non-inverting terminal and with the use of internal buffers, the output current follows the input current.

In this work, however, a voltage buffer amplifier is designed. Therefore, the OTA is used as a programmable block to control the output voltage rather than the output current.

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### 2.1.2 Other Topologies

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The Conventional OTA considered in the previous section is a Class A type of amplifier. These amplifiers can produce output currents that are equal to the bias currents for a unity current mirror gain. Need for high speed and high current with low power dissipation gave rise to research in Class AB Amplifiers. There are many topologies of the Class AB type OTAs that are classified based on their structure as:

- Folded Cascode OTA
- Telescopic OTA
- Local Common Mode Feedback OTA
- Cascode Voltage Flipped Follower based OTA

One of the figures of merit to compare these OTAs is the Current Enhancement factor (CE).  $CE = I_{out\_max}/2I_{bias}$ . Class A amplifiers typically have a CE of 1. Class AB OTAs have a high CE value. A family of OTA denoted as *SuperClass – ABOTA* have remarkably large CE values. Typical value of CE ranges from 100 to 500 depending on the implementation of the Class-AB differential pair and the output branches. Cascode Voltage Flipped Follower based OTA is one such example of a Super Class-AB OTA. These OTAs have the largest reported CE in literature.

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### Super Class AB OTA

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The schematic of a Super Class AB OTA based on a Cascode Voltage Flipped Follower is as shown in Figure.2.3. It uses a Class-AB differential pair to boost the bias current by a factor of  $k_1$ . Typical values of  $k_1$  range from 30 to 60 depending on the implementation of the Class-AB differential pair and supply voltages. An additional current boosting by a factor of  $k_2$  is obtained in the output branches. The typical values of  $k_2$  range from 3-8. Using cascode transistors maximum output current and subsequently the CE values are reduced. But lack of cascode transistors results in very low open loop gain. Therefore, a simple power efficient technique is used that allows the use of cascode transistors in order to achieve even higher open loop gain and gain bandwidth by maintaining a large CE value. This is done with the help of dynamically biasing the cascode transistors using RC networks (indicated in purple in the figure). The capacitor  $C_{bat}$  acts as a floating battery for dynamic operations that transfers variations in the gate voltages of mirror transistors to the gate of the cascode transistors. This increases the  $V_{DS}$  of the mirror transistors allowing high output currents. Cascode Voltage Flipped Follower (indicated in red in the figure) is characterized by high input swing and can operate over a wide range of supply voltages. It uses local shunt feedback to provide nodes A and B with very low impedance.

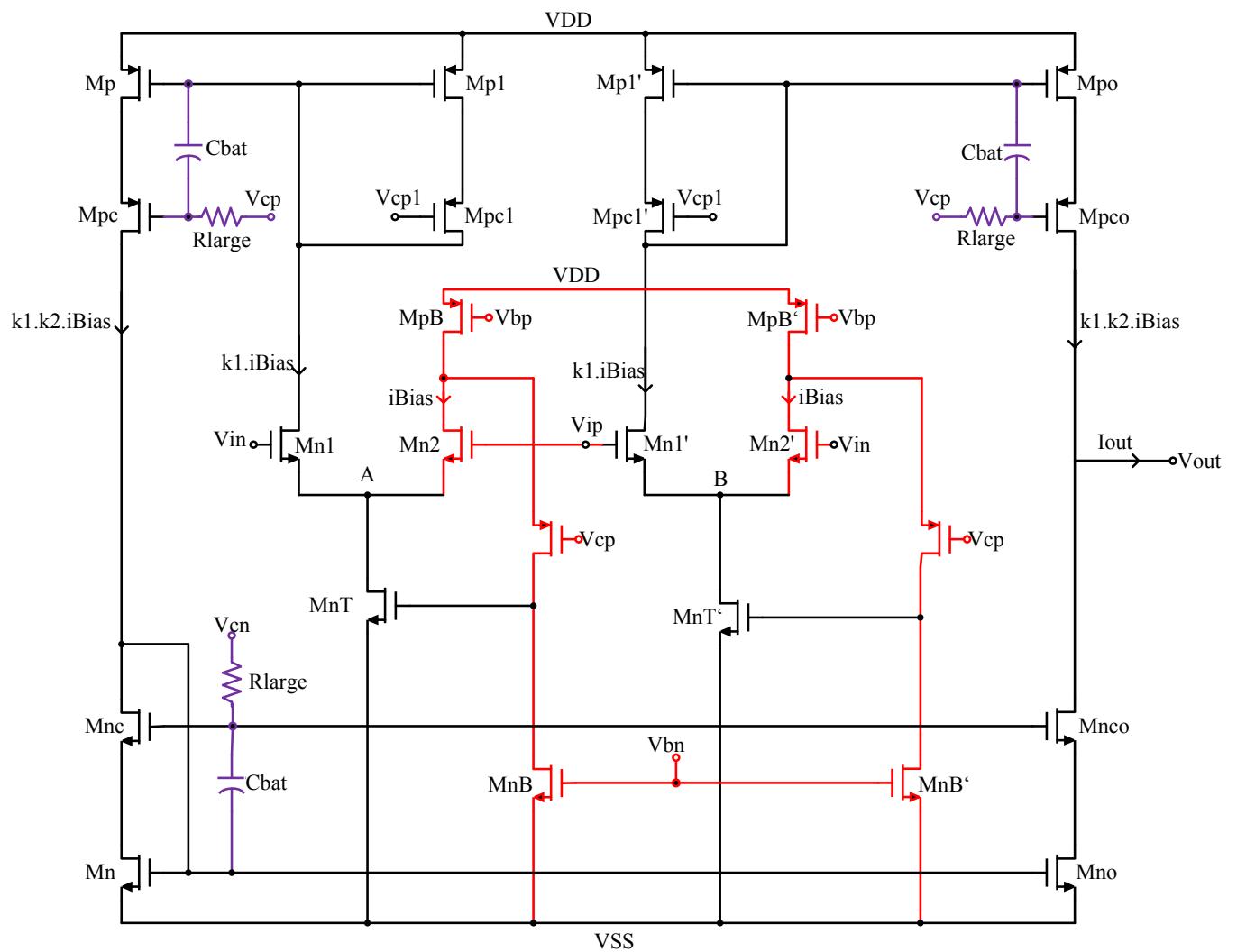


Figure 2.3.: Super Class-AB OTA with a Cascode Voltage Flipped Follower

Taking into account the complexity, output current requirements and load requirements, this design will not be considered further as part of this research work. The reason being - a high current output of this Class AB OTA would give rise to another output buffer. And the capacitors needed to consume such a high current would be in the order of several pico Farads. Having such a big capacitor inside an IC is not recommended. Also, considering how complex CMOS Current Feedback Amplifiers are, it is wise to stick to a simpler design that does the same work but with some minor trade-offs.

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## 2.2 The Opeartional Amplifier

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The operational amplifier (OP AMP) is a fundamental building block in analog integrated circuit design. A typical symbol of an OP AMP is as shown in Figure.2.4. It consists of differential input terminals - inverting ( $V_{in-}$ ) and non-inverting ( $V_{in+}$ ), a positive supply pin ( $V_{DD}$ ) and a negative/ground supply pin ( $V_{SS}$ ) and the output terminal ( $V_{out}$ ). Simple OP AMPS are generally used as Summing amplifier, differentiator, integrator, transimpedance amplfier, and so on. OP AMPS with vastly different levels of complexity are used to realize functions ranging from dc bias generation to high-speed amplification or filtering.

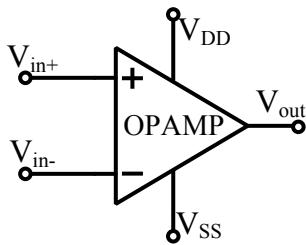


Figure 2.4.: Symbol of an OP AMP

OP AMPs are voltage controlled voltage sources, much to the contrary of OTAs, which are voltage controlled current sources. Design of an OP AMP consists of determining the specifications, selecting the device sizes and biasing conditions, compensating the op-amp for stability, simulating and characterizing the open loop gain, the input common mode range, common mode rejection ratio, power supply rejection ration, output votlage range, current sourcing/sinking capability and power dissipation. OP AMPs are used as a voltage buffer whenever there is a need to drive resistive loads or a combination of capacitive load and a resistive load.

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### 2.2.1 Miller Compensation OP AMP

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The schematic of the a two-stage Miller Compensation OP AMP is as shown in Figure.2.5. First stage of the OP AMP is the differential amplfier formed by the transistors  $M_1 - M_5$  and  $M_8$ . The second stage of the OP AMP is the Common Source amplifier formed by the transistors  $M_6$  and  $M_7$ . Compensation Capacitor or Miller Capacitor denoted by  $C_C$  is used for the splitting of the poles of the two stages thereby providing stability and also controlling the bandwidth and the slew rate.

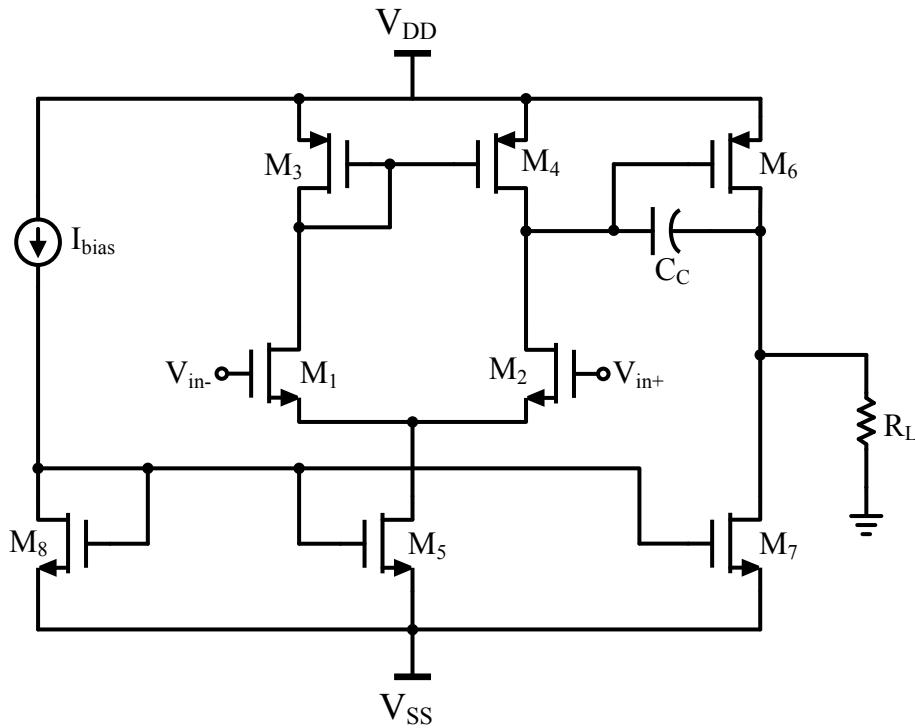


Figure 2.5.: Schematic of a Two-stage Miller Compensation OP AMP

The biasing to the differential input pair is provided by the current mirror formed by  $M_5$  and  $M_8$ . The selection of the bias current,  $I_{bias}$  is determined by gain, ICMR, CMRR, power dissipation, noise, slew rate and matching considerations. If slew-rate is of concern, a cross-coupled differential amplifier should be considered for the first stage. The second stage common source amplifier is used to provide an additional gain to the overall OP AMP since the gain provided by the differential pair is usually not large enough. It is known that OP AMPs have infinitely large open loop gain. This second stage also provided a current boosting since  $M_7$  forms a current mirror with  $M_8$  and the dimensions of  $M_7$  are usually larger than  $M_8$ , thereby setting the current at the output of the OP AMP. The best use of these OP AMPs are made as part of voltage buffers. i.e., the OP AMP is connected in unity gain configuration by providing a direct feedback from the output terminal to the non-inverting terminal.

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### 2.3 The Gm/Id Methodology

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# 3 Design and Implementation

## 3.1 OTA Design

The first stage of the design is the Operational Transconductance Amplifier. This stage uses bipolar power supply of 2.5V and -2.5V. Two separate designs - OTA with PMOS Differential pair and OTA with NMOS Differential pair were designed and simulated in Cadence to get a good understanding of the circuit parameters and how the differential pair impact them.

Generally, amplifiers with PMOS transistors used as differential input offer high linearity, low flicker ( $1/f$ ) noise. A possible reason to choose PMOS transistors as a differential pair comes from the necessity to reduce the influence of the substrate noise. The two transistors are in an N-well and the well is connected to the supply voltage that any substrate interference coupled via the parasitic capacitance to substrate is decoupled to the VDD line. P-channels typically have less flicker noise ( $1/f$  noise) caused by the carriers randomly entering and leaving traps introduced by defects near the semiconductor surface. Since the majority charge carriers are holes in PMOS, there are less potential to be trapped in surface states.

Having stated all these facts, NMOS input transistors would be better in terms of transconductance gain, and hence thermal noise and the bandwidth of the amplifier. An important fact considered in the selection of the type of differential pair is that the OTA in this system is being used in an open loop. This means that the gain of the OTA has to be limited in order to avoid saturation at the crests and troughs at the output of the OTA.

PMOS transistors exhibit their low noise behaviour due to the fact that PMOS transistors are usually bigger (higher W/L ratio) than an NMOS pair. Since we need a small open loop gain, we cannot afford to have big transistors at the input that cause the output to saturate. Along with this, it was also seen that, the range of bias currents needed to obtain a similar range of voltage swing was much wider in case of PMOS than NMOS. Since this indirectly results in a high power dissipation in the 1st stage itself, it was concluded to use the NMOS based design as part of this Thesis work.

### 3.1.1 Schematic

Figure.3.1 shows the schematic of the OTA used as part of this Thesis. As mentioned above, the differential amplifier is formed by the NMOS pair. The bias current  $I_{bias}$  to the differential pair is provided by the current mirror formed by  $M_{nB1}$  and  $M_{nB2}$  which is in turn controlled by the PMOS transistor  $M_9$  whose gate voltage  $V_{bias}$  is the programmable variable of the OTA and the entire system. This bias voltage is inversely proportional to the bias current flowing through each branch of the OTA.

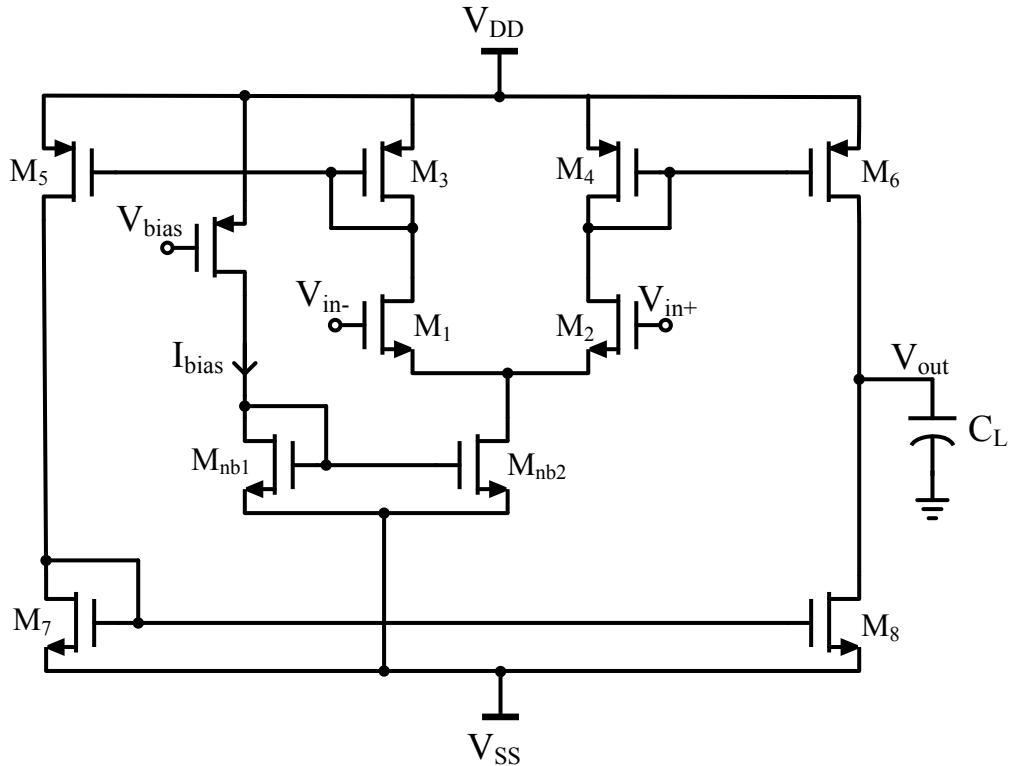


Figure 3.1.: Schematic of the OTA Designed

Transistor	Width	Length	Multiplier
M1	8u	500n	5
M2	8u	500n	5
M3	35u	500n	1
M4	35u	500n	1
M5	28u	500n	3
M6	35u	500n	3
M7	35u	500n	18
M8	33u	500n	18
M9	10u	500n	4
MnB1	20u	500n	8
MnB2	20u	500n	8

Table 3.1.: Dimensions of the Transistors of the designed OTA

From the Table.3.1, it can be seen that the current mirror gain for the current mirrors formed by  $M_3$ ,  $M_5$  and  $M_4$ ,  $M_6$  is 3. It can also be noted that the transistors  $M_5$  and  $M_8$  are not symmetric with respect to their counterparts. This is designed so to make sure the DC bias point at the output of the OTA is close to 0V, which is the mid point of  $V_{DD}$  and  $V_{SS}$ . Changing the bias current of an amplifier, will automatically change the DC bias point at its output. Since we are using the OTA as a programmable block, it is important to have an output which is symmetric over 0. Even though it is not entirely possible,

making these two transistors a little assymetric or uneven would close the gap between the bias points for highest and lowest bias currents.

### 3.1.2 Test Setup

In the following subsections, let us look at how this  $V_{bias}$  affects various paramters of the OTA by performing DC, AC, Transient and Noise Analyses.

#### DC Analysis

The circuit in Figure.3.2, is a common test bench to measure various parameters. The AC source in  $V_{ac}$  does not impact the DC analysis in any way. So for the DC analysis, the differential pair transistors will have a voltage of  $V_{cm}$  at its gates.  $V_{bias}$  is the variable supply used as a programmable parameter and a capacitive load to measure the outputs.

$$V_{DD} = 2.5V; V_{SS} = -2.5V; V_{cm} = 1.95V; V_{bias} = 150mV \text{ to } 700mV; V_{ac} = 1V; C_L = 50fF.$$

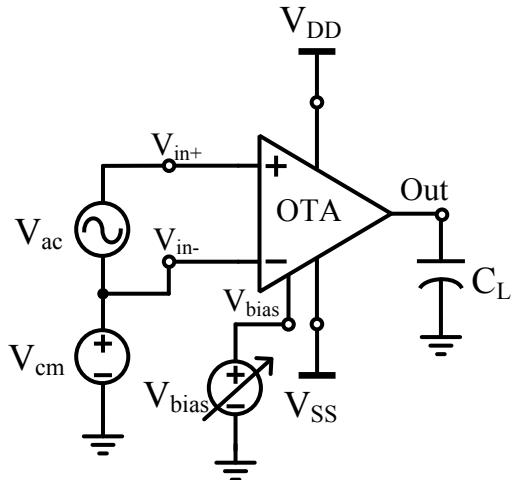


Figure 3.2.: OTA Test setup for AC, DC and Noise Analysis

Vbias (mV)	150	200	300	400	500	600	700
Output DC Bias (mV)	13.68	-16.43	-78.96	-144.6	-213.3	-285.2	-360.3

Table 3.2.: Output DC Bias Point of the OTA

The output DC bias points for various values of  $V_{bias}$  is tabulated in Table3.2. The DC voltage at the output decreses with the increase in bias voltage or decrease in bias current.

#### AC Analysis

Figure. 3.2 is used for AC analysis as well.

$$V_{DD} = 2.5V; V_{SS} = -2.5V; V_{cm} = 1.95V; V_{bias} = 150mV \text{ to } 700mV; V_{ac} = 1V; C_L = 50fF.$$

The same parameters hold good for AC analysis as well. Here, the value of  $V_{ac}$  becomes significant. The value of the open loop gain is given by the ratio of output voltage to input voltage.

$$\text{Gain} = \frac{V_{out}}{V_{in}}$$

$$\text{Gain in dB} = 20\log_{10} \frac{V_{out}}{V_{in}}$$

Figure 3.3 shows the semilog plot of Gain for maximum and minimum  $V_{bias}$ . The plots for the other values of  $V_{bias}$  are not shown here as that would make the plot less legible for understanding and more so because those plots would be in between the plots already shown.

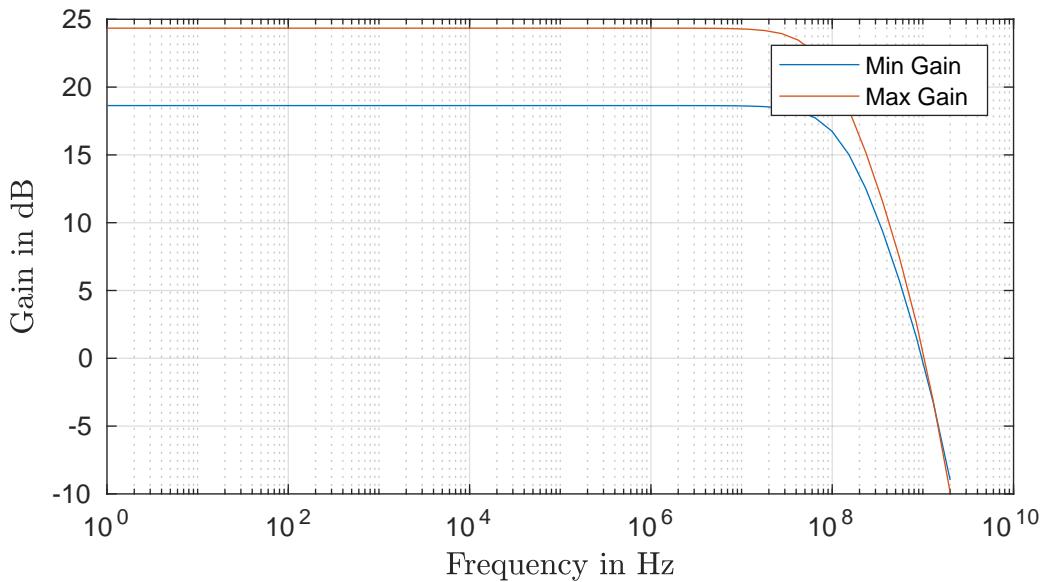


Figure 3.3.: OTA Plot of Gain vs Frequency for different  $V_{bias}$

The variation of open loop gain of the OTA, its bandwidth and the phase margin with respect to  $V_{bias}$  is tabulated in Table.???. The bandwidth is kept at a high value in order to make sure that the dominant poles of the first and second stages are sufficiently far from each other so as to obtain a stable system for operation. The maximum open loop gain for this system is 23.7dB. Any value beyond this will cause the amplifier to saturate at the crests and troughs of the output voltage.

$V_{bias}$ (mV)	150	200	300	400	500	600	700
Open Loop Gain (dB)	18.64	19.46	20.9	22.05	22.96	23.7	24.35
Phase Margin (degrees)	63.3	60.63	55.96	52.42	49.91	48.12	46.72
Bandwidth (MHz)	135	130.7	122.2	113.7	105	96.63	89.19

Table 3.3.: Open Loop Gain, Phase Margin and Bandwidth of the OTA

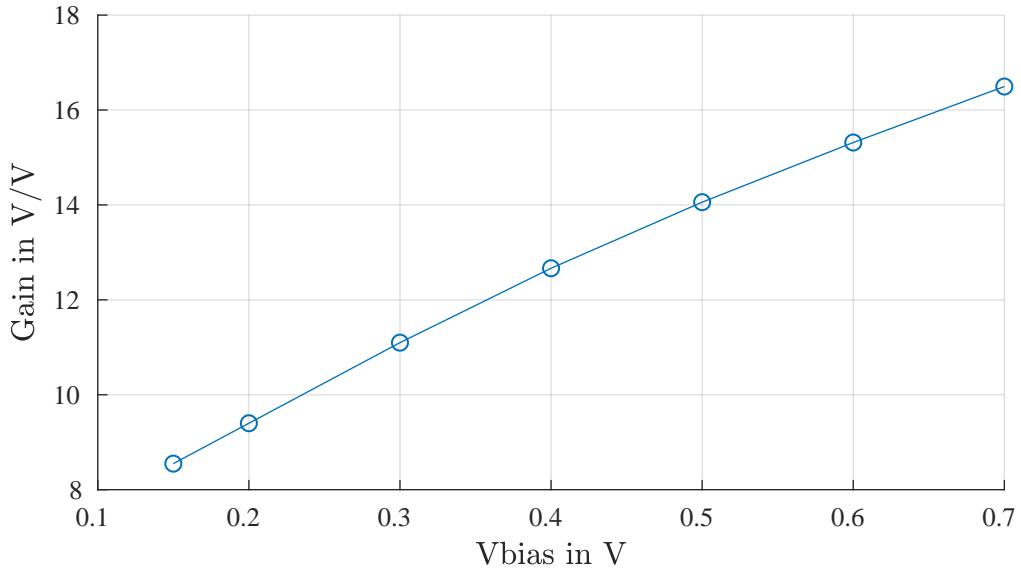


Figure 3.4.: OTA Plot of Gain vs Vbias

On the same lines, the variation of absolute value of gain with respect to different values of  $V_{bias}$  is seen in Figure 3.4 tabulated in Table 3.4. The reason to choose the values of  $V_{bias}$  are evident from this table and that is because it provides us a gain in the ratio of almost 1:2 for those values of  $V_{bias}$ . As seen from the table of specifications that we need a transconductance in the ratio of 1:1.8 or something close to that.

Vbias (mV)	150	200	300	400	500	600	700
DC Gain (V/V)	8.548	9.4	11.1	12.67	14.06	15.31	16.49

Table 3.4.: Absolute values of DC Gain of the OTA

### Noise Analysis

Once again, we use the same test bench as shown in Figure 3.2. As mentioned in the previous section, the NMOS differential pair exhibits high flicker noise, also known as 1/f noise. This noise is significant at low frequencies. At moderate and high frequencies, the effect of thermal or white noise is much more dominant and hence flicker noise becomes less significant at those frequencies.

The variation of the input referred noise with respect to  $V_{bias}$  is tabulated in Table 3.5. With increase in  $V_{bias}$ , the input referred noise decreases. This is because the gain increases with increase in  $V_{bias}$  and consequently input referred noise decreases. This also converges to the fact that the input referred noise decreases with bigger dimensions of the differential pair, and thereby contributing to a higher gain.

Vbias (mV)	150	200	300	400	500	600	700
Input Referred Noise ( $nV/\sqrt{Hz}$ )	46.36	43.73	39.83	37.29	35.57	34.31	33.28

Table 3.5.: Input Referred Noise of the OTA

---

### Transient Analysis - Sine Input

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The test bench in Figure.3.5 is used to measure the transient parameters for a sine wave input. An ideal balun is used at the input to provide a differential voltage to the terminals. The voltages are  $180^0$  out of phase with each other.

$V_{DD} = 2.5V$ ;  $V_{SS} = -2.5V$ ;  $V_{cm} = 1.95V$ ;  $V_{bias} = 150mV$  to  $700mV$ ;  $V_{sin} amplitude = 100mV$ ;  $frequency = 1MHz$   $C_L = 50fF$ .

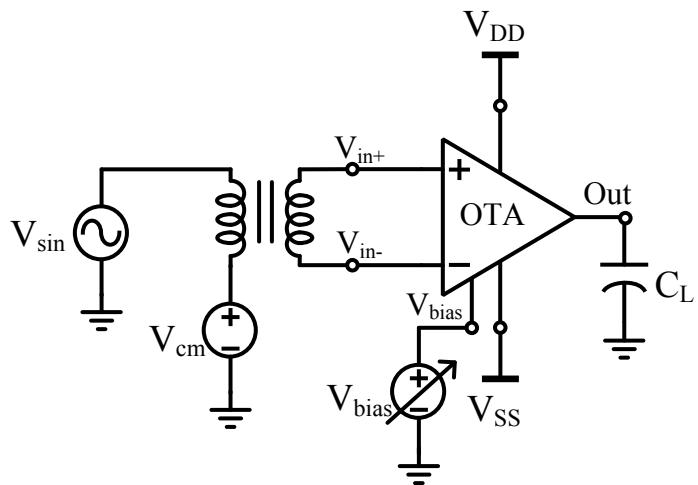


Figure 3.5.: OTA Test setup for Transient Analysis - Sine Wave Input

The transient simulation is carried out for  $3\mu s$ . The sinusoidal output with respect to time is shown in Figure.3.6 for different values of  $V_{bias}$ . As we know from the AC analysis, the gain varies in the ratio of 1:2 and thereby, here we have the peak-to-peak voltages varying in the same ratio. And as discussed in the DC Analysis, the DC bias points at the output are close to zero but not exactly at 0 for all the  $V_{bias}$  values.

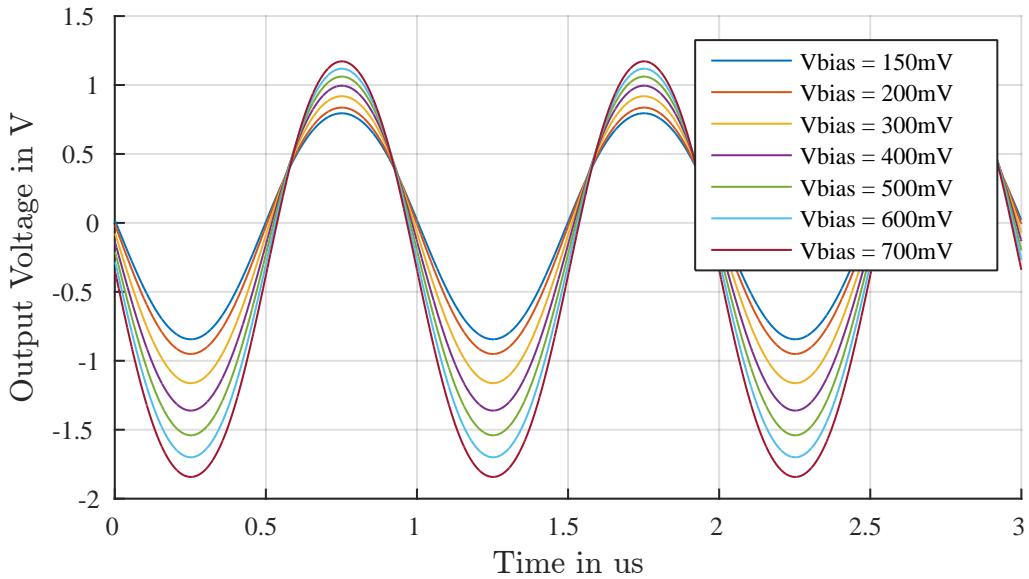


Figure 3.6.: OTA Output Voltage for vs time for different  $V_{bias}$

Table 3.6 shows the variation of different transient parameters with respect to  $V_{bias}$ . The voltage swing varies from 1.639 to 3.014 which is almost in the ratio of 1:2. HD3 worsens with increase in  $V_{bias}$  and on the contrast, improves. Although, it is seen that the HD2 again starts to worsen beyond 600mV of  $V_{bias}$ .

$V_{bias}$ (mV)	150	200	300	400	500	600	700
Vout Max (V)	0.7949	0.8362	0.9188	0.995	1.061	1.119	1.172
Vout Min (V)	-0.844	-0.9505	-1.162	-1.361	-1.54	-1.7	-1.842
Vout Swing (V)	1.639	1.787	2.081	2.356	2.601	2.818	3.014
HD2 (dBc)	-44.83	-44.28	-43.86	-44.74	-47.97	-60.86	-48.75
HD3 (dBc)	-52.15	-49.69	-46.13	-43.87	-42.24	-40.78	-39.32

Table 3.6.: Transient Parameters of the OTA

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### Transient Analysis - Square Input

---

The test bench in Figure 3.7 is used to measure the transient parameters for a square wave input. Slew rate is the rate of change of output voltage. It is also defined as the ratio of bias current to load capacitance.

$V_{DD} = 2.5V$ ;  $V_{SS} = -2.5V$ ;  $V_{cm} = 1.95V$ ;  $V_{bias} = 150mV$  to  $700mV$ ;  $V_{pulse}V_1 = 100mV$ ;  $V_2 = -100mV$ ;  $PulseWidth = 49.7ns$ ;  $Period = 100ns$ ;  $RiseTime = 300ps$ ;  $Falltime = 300ps$ ;  $C_L = 50fF$ .

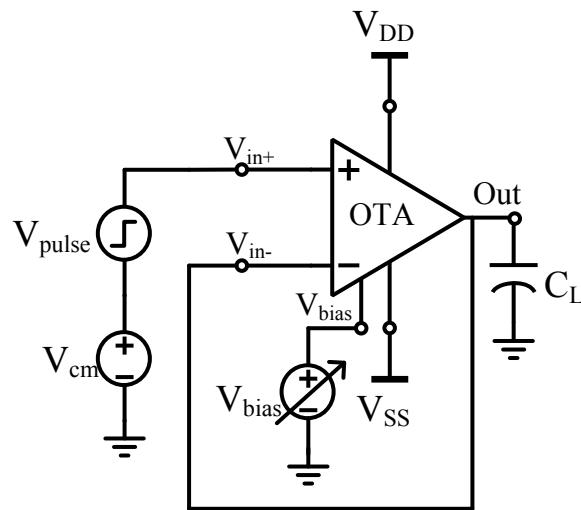


Figure 3.7.: OTA Test setup for Transient Analysis - Square Wave Input

The slew rate, both for the rising edge and the falling edge is tabulated in the Table.3.7.

Vbias (mV)	150	200	300	400	500	600	700
Slew Rate Rising Edge (V/us)	386.9	407.8	452.9	501.9	547.6	585.4	609.3
Slew Rate Falling Edge (V/us)	-389	-413.8	-469.4	-523.5	-570.5	-605.7	-626.4

Table 3.7.: Slew Rate of the OTA

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### AC Analysis - PSRR

---

PSRR or Power Supply Signal Ratio is defined as the ability of an amplifier to maintain its output as its DC power supply is varied. It is the ratio of change in the output current to change in power supply.

$$PSRR = \frac{\Delta I_{out}}{\Delta V_{DD}}$$

$V_{DD} = 2.5V$ ;  $AC\text{Magnitude of } V_{DD} = 1V$ ;  $V_{SS} = -2.5V$ ;  $V_{cm} = 1.95V$ ;  $V_{bias} = 150\text{mV to } 700\text{mV}$ ;  $C_L = 50\text{fF}$ .

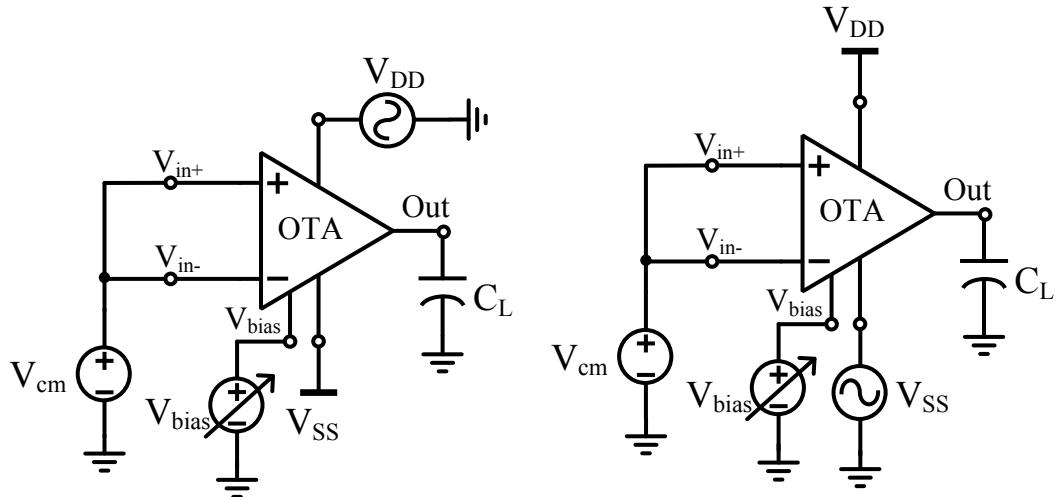


Figure 3.8.: OTA Test setup for calculating PSRR

$V_{DD} = 2.5V$ ;  $V_{SS} = -2.5V$ ;  $AC\text{Magnitude of } V_{SS} = 1V$ ;  $V_{cm} = 1.95V$ ;  $V_{bias} = 150\text{mV to } 700\text{mV}$ ;  $C_L = 50\text{fF}$ .

The test bench to measure the PSRR of the OTA is as shown in the Figure.3.8. The one on the left is used to measure PSRR for a change in  $V_{DD}$ . And similarly, the one on the right side is used to measure PSRR for a change in  $V_{SS}$ .

The variation of PSRR with respect to  $V_{bias}$  is tabulated in Table.3.8. The PSRR is fairly low in the range of nA/V and increases only slightly with increase in  $V_{bias}$ .

Vbias (mV)	150	200	300	400	500	600	700
PSRR (VDD Supply) (nA/V)	234.6	236.9	241.6	246.7	252.2	258	264.3
PSRR (VSS Supply) (nA/V)	254.5	256.5	260.4	264.2	267.9	271.5	274.9

Table 3.8.: Power Supply Rejection Ratio of the OTA

---

### AC Analysis - Input Impedance

---

OTAs generally exhibit a very high input impedance. And to measure this parameter, an AC current source with a magnitude of 1A is connected to the non-inverting terminal of the OTA. And the input impedance is given by the ratio of the AC voltage to the AC current at the input of the OTA. Since the current at the input is 1A, the magnitude of the input voltage will be the value of the input impedance in Ohms at that particular frequency. Since the OTA operation is carried out at 1MHz, the voltage at 1MHz is considered to obtain the value of Input Impedance. Figure.3.9 shows the test bench to measure the input impedance.

$V_{DD} = 2.5V$ ;  $V_{SS} = -2.5V$ ;  $V_{bias} = 150\text{mV to } 700\text{mV}$ ;  $C_L = 50\text{fF}$ ;  $I_{sinmagnitude} = 1A$ .

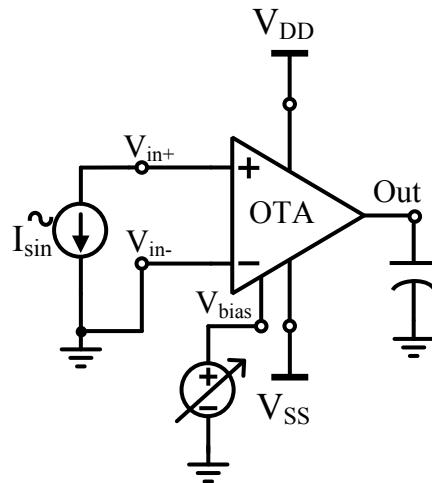


Figure 3.9.: OTA Test setup for calculating Input Impedance

The variation of input impedance with respect to  $V_{bias}$  is tabulated in Table 3.9. The variation is quite small and it increases slightly with an increase in bias voltage and is in the order of Mega Ohms.

Vbias (mV)	150	200	300	400	500	600	700
Input Impedance ( $M\Omega$ )	3.38	3.388	3.403	3.419	3.434	3.45	3.467

Table 3.9.: Input Impedance of the OTA

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### AC Analysis - Output Impedance

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In a concept similar to measuring the input impedance, the output impedance too, is measured with a help of a current source with unity magnitude at the output instead of a Capacitive load. The differential inputs are connected in a common mode configuration. The test bench to measure the output impedance of the OTA is as shown in Figure 3.10.

$$V_{DD} = 2.5V; V_{SS} = -2.5V; V_{cm} = 1.95V; V_{bias} = 150mV \text{ to } 700mV; I_{sin} \text{ magnitude} = 1A.$$

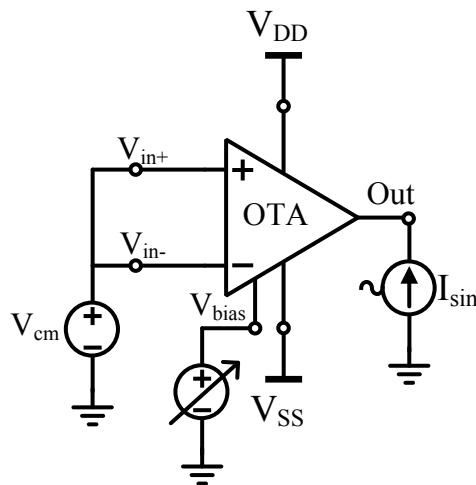


Figure 3.10.: OTA Test setup for calculating Output Impedance

As discussed in the theory chapter, OTAs generally have high output impedance. And its variation with respect to  $V_{bias}$  is tabulated in the Table.3.10

Vbias (mV)	150	200	300	400	500	600	700
Output Impedance ( $k\Omega$ )	1.262	1.3	1.382	1.473	1.577	1.697	1.837

Table 3.10.: Output Impedance of the OTA

## 3.2 OP AMP Design

The second stage of the design is the Operational Amplifier. This stage too, like the first stage uses a bipolar power supply of 2.5V and -2.5V. A two-stage Miller compensated op amp is designed and it is used as a Voltage Buffer for the OTA.

### 3.2.1 Schematic

The schematic of the Miller compensation op amp is as shown in the Figure.3.11. The bias current to the differential pair is provided through the current mirror pair of  $M_5$  and  $M_8$ . The bias current is controlled through the transistor  $M_9$  whose gate voltage is provided by the voltage divider formed by the resistors  $R_{b1}$  and  $R_{b2}$ .  $C_C$  is the compensation capacitance between the two stages of the amplifier.

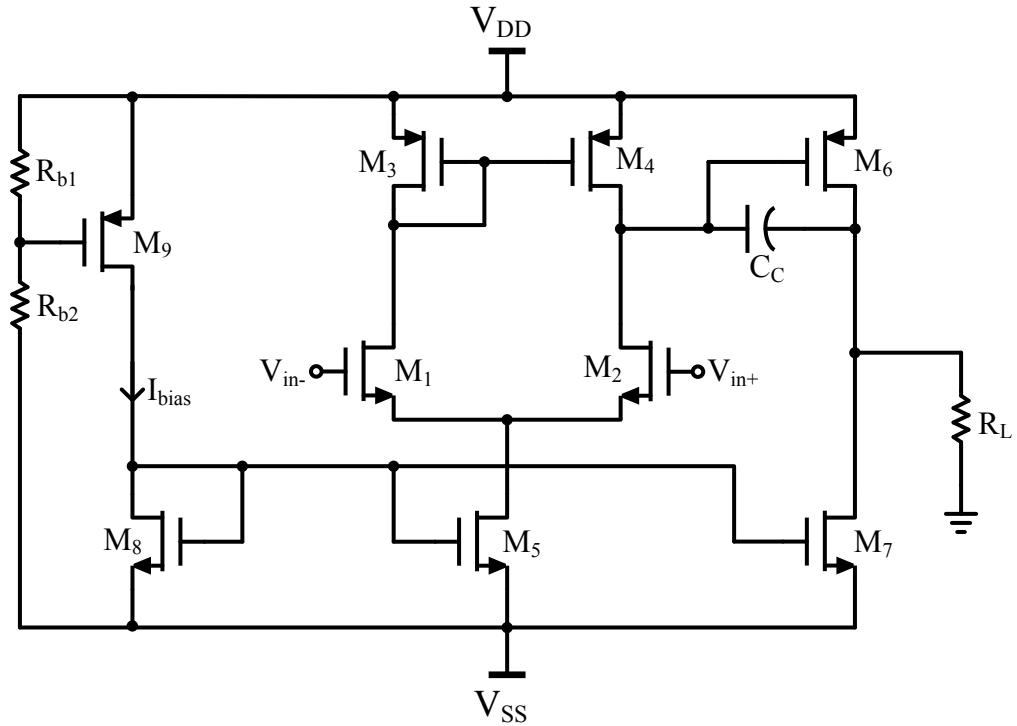


Figure 3.11.: Schematic of the OPAMP Designed

The dimensions of the transistors are tabulated in Table 3.11. One can easily recognize that the transistors at the output are large. This is designed so to make sure the op amp provides a very high current and thereby avoiding a common collector amplifier at the output of the op amp to amplify the output current. The compensation capacitance value is 50fF.

Transistor	Width	Length	Multiplier
M1	5u	500n	2
M2	5u	500n	2
M3	30u	500n	1
M4	30u	500n	1
M5	2u	500n	1
M6	85u	500n	55
M7	50u	500n	48
M8	2u	500n	1
M9	700n	500n	1

Table 3.11.: Dimensions of the Transistors of the designed OPAMP

### 3.2.2 Test Setup

In the following subsections, the typical parameters of an op amp are simulated and analysed.

## DC Analysis

Input Common Mode Range or ICMR of an operational amplifier is the range of the common mode input voltage over which the amplifier behaves as a linear amplifier for differential input signals. In simpler words, the ICMR is defined by the input voltages at which all the transistors remain in saturation. Figure.3.12 shows the test setup to measure the ICMR of the op amp designed.

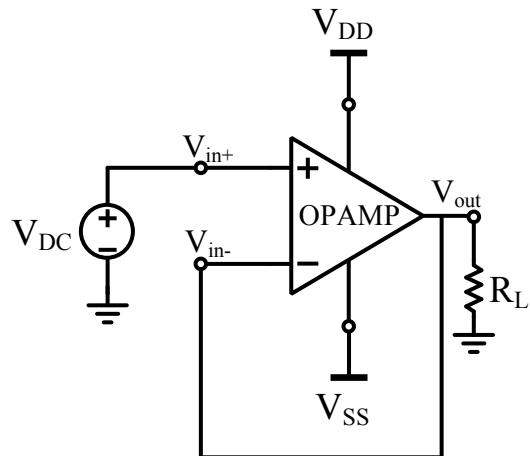


Figure 3.12.: OPAMP Test setup for calculating ICMR

The transistors which impact the ICMR are  $M_3$ ,  $M_4$  and  $M_5$ . These transistors were designed in such a way that the op amp ICMR lies between -2V and 2V. The DC input voltage at the non-inverting terminal is swept in the range of  $V_{DD}$  and  $V_{SS}$ , i.e., 2.5V to -2.5V. The plot of DC output voltage for a varying DC input voltage is as shown in Figure.3.13. It can be noted that the minimum value of the ICMR is -2.19V and the maximum value of the ICMR is 2.089V.

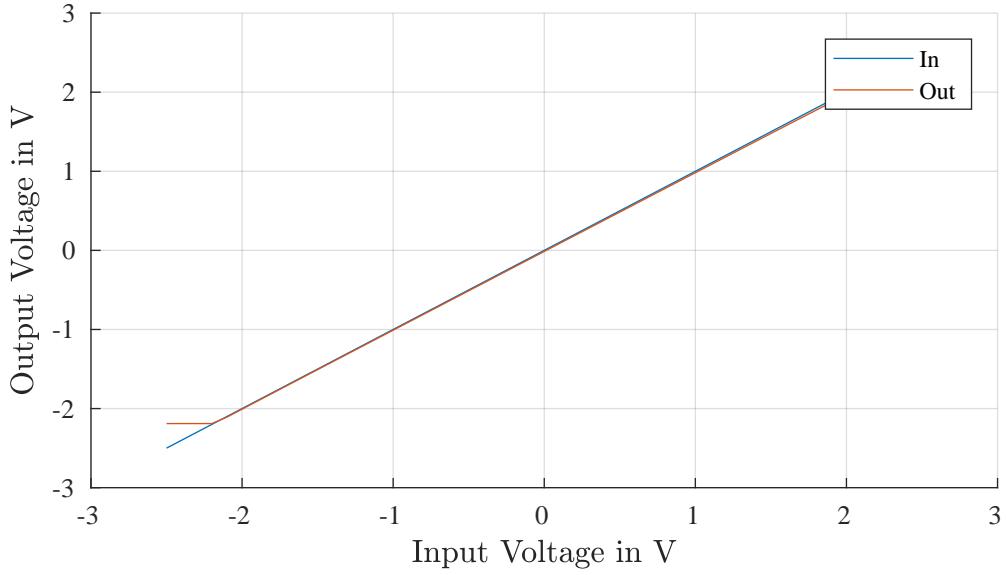


Figure 3.13.: OPAMP Plot of ICMR vs  $V_{in}$

The output voltage swing is the range of voltage that an opamp can physically provide at its output. On the same lines as ICMR but without feedback, the output voltage swing is measured and plotted with respect to the input voltage. And the output voltage ranges from -2.19 to 2.323. The plot is as shown in Figure 3.14.

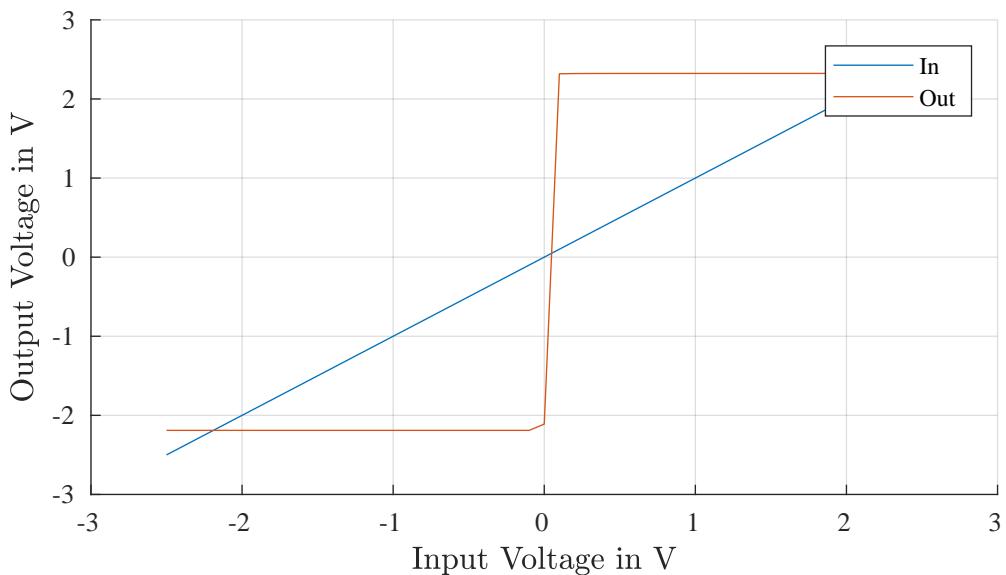


Figure 3.14.: OPAMP Plot of Output Voltage Swing vs  $V_{in}$

## AC Analysis

The open loop gain and the gain bandwidth product are inarguably the most important parameters of an operational amplifier. Before the op amp was actually designed, these parameters were simulated with the help of an ideal op amp from the ahdlLib in Cadence. In order to have a stable system, i.e., an OTA with an OP AMP buffer in a cascade configuration, it is necessary to have the dominant poles far from each other. So that is the reason for the OTA to have a high bandwidth and correspondingly the OP AMP bandwidth can be adjusted so as to reach the specification and also to have a stable system.

The test bench used to measure these parameters is as shown in Figure.3.15. The op amp is configured in an open loop. The DC biasing to the differential transistor pair is provided by the output of the OTA, which is centred around 0V.

$$V_{DD} = 2.5V; V_{SS} = -2.5V; V_{ac\text{magnitude}} = 1 \text{ V}; R_L = 50\Omega.$$

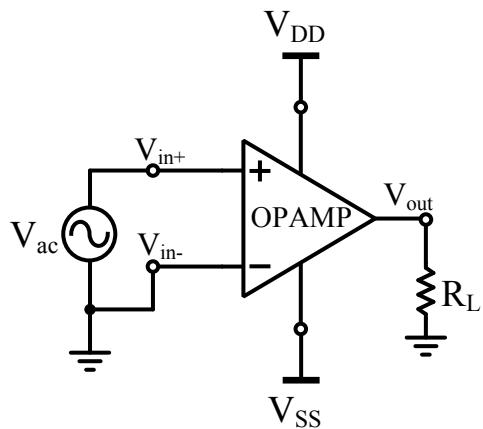


Figure 3.15.: OPAMP Test setup for AC, DC and Noise Analysis

The plot of the open loop gain and the phase of the op amp is as shown in the Figure.3.16. The open loop gain of the op amp is 30.8dB. The gain bandwidth product is 6.2MHz and correspondingly the phase margin is  $76.79^0$ .

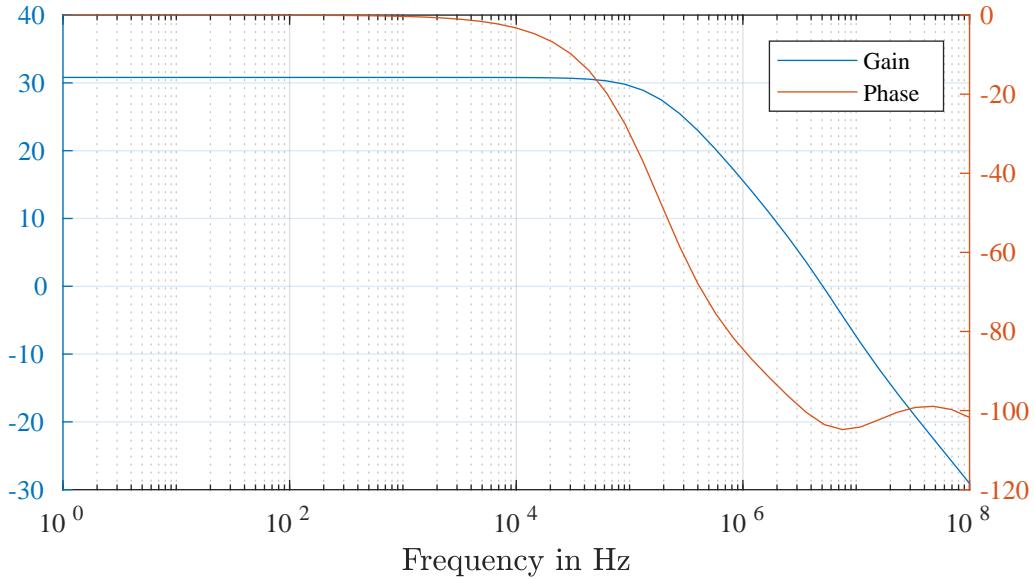


Figure 3.16.: OPAMP Plot of Gain and Phase vs Frequency

The PSRR of the OTA in the first stage was in the range of  $nA/V$  and hence negligible. However, the op amp designed exhibits a relatively high value of PSRR. This is due to the bulky transistors at the output stage. The test bench to measure the PSRR of the op amp is as shown in the Figure.3.17. The one on the left is used to measure PSRR for a change in  $V_{DD}$ . And similarly, the one on the right side is used to measure PSRR for a change in  $V_{SS}$ .

$V_{DD} = 2.5V$ ;  $V_{SS} = -2.5V$ ;  $V_{ac\,magnitude} = 1 V$ ;  $R_L = 50\Omega$ .

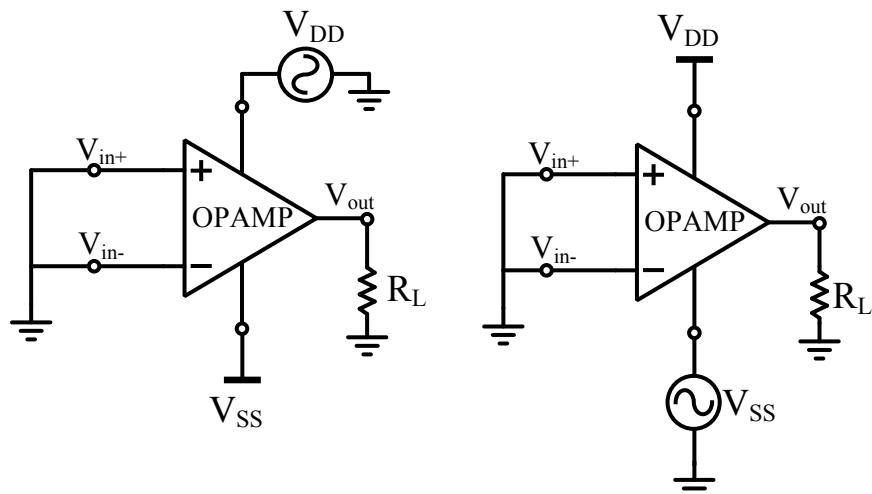


Figure 3.17.: Test setup for calculating PSRR

The PSRR ( $V_{DD}$ ) of the op amp is 30.96 uA/V, and PSRR ( $V_{SS}$ ) is 130.96 uA/V. In the next chapter, it is seen that the assymetry in the two PSRR values doesn't affect the PSRR of the whole system.

The test bench to calculate the input impedance follows a similar concept to that of the OTA. i.e., a unitz current source is connected to the non-inverting terminal of the op amp and the voltage at that terminal is measured which in turn is the magnitude of the input impedance of the op amp.

$$V_{DD} = 2.5V; V_{SS} = -2.5V; I_{sinmagnitude} = 1A; R_L = 50\Omega.$$

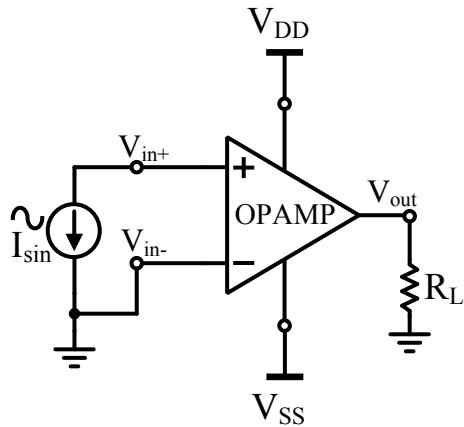


Figure 3.18.: Test setup for calculating Input Impedance

To have a stable system, it is important that input impedance of the second stage is higher than the output impedance of the first stage. Given that the output impedance of the OTA is generally high, the input impedance of the op amp becomes a very important parameter. And th input impedance is found to be 9.04MΩ.

On the same lines, the test bench to measure the output impedance of the op amp is as shown in Figure.3.19. The resistive load is replaced by a current source with unity magnitude. And hence the voltage at the output will be the magnitude of the output impedance of th op amp.

$$V_{DD} = 2.5V; V_{SS} = -2.5V; I_{sinmagnitude} = 1A.$$

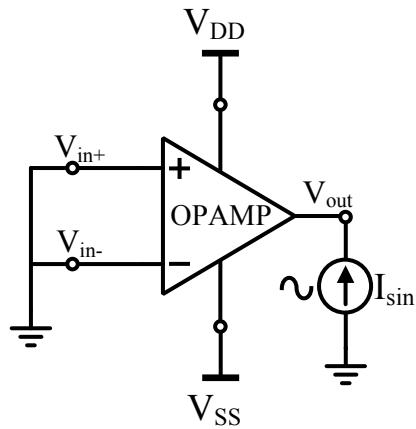


Figure 3.19.: Test setup for calculating Output Impedance

Op amps are known to have zero or very low output impedance. Along with this theoretical fact, the op amp in this work is designed to retrieve very high currents, so it is expected to have a very low output impedance. The value of the output impedance is observed to be  $4.167\Omega$ .

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#### Transient Analysis

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The test bench to perform the transient analysis is as shown in the Figure.3.20. The op amp is connected in a negative feedback configuration so as to analyse its behavior as a voltage buffer.

$V_{DD} = 2.5V$ ;  $V_{SS} = -2.5V$ ;  $V_{sin}$  amplitude =  $1.5V$ ; frequency =  $1MHz$ ;  $R_L = 50\Omega$ .

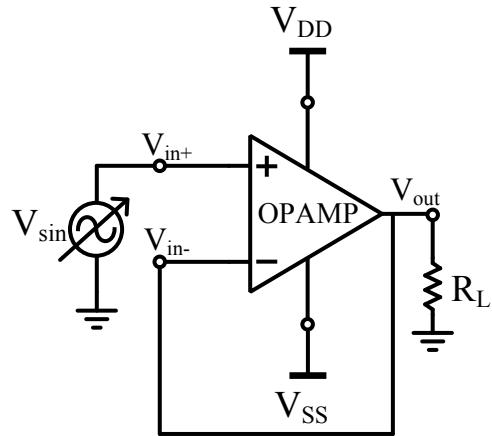


Figure 3.20.: Test setup for Transient Analysis - Sine Wave Input

The plot of input voltage and output voltage with respect to time is as shown in the Figure.3.21. The output follows the input only with a very small delay.

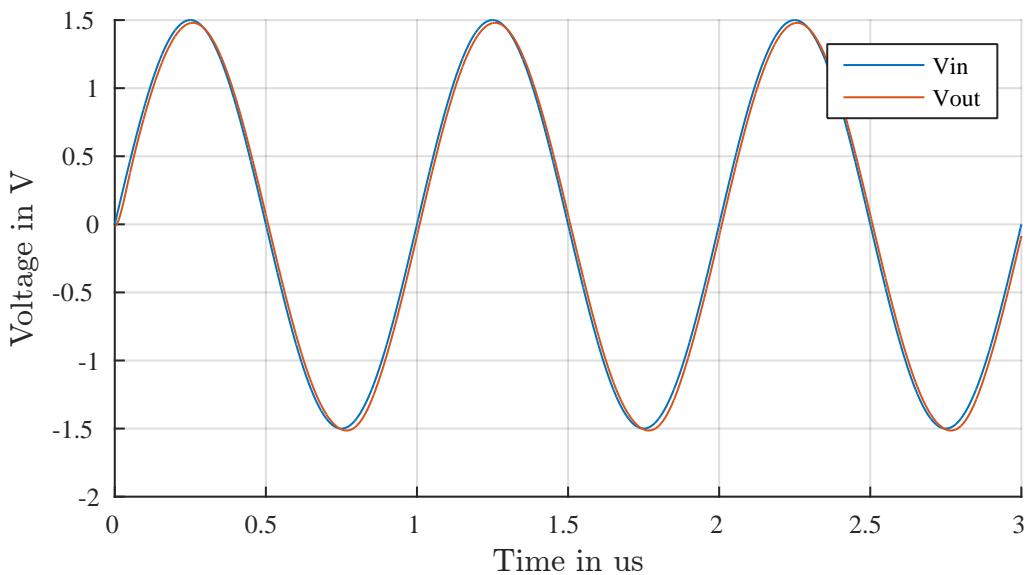


Figure 3.21.: OPAMP Plot of Output Voltage vs time

The output current of the op amp is effectively the current flowing through the load resistor  $R_L$ . The plot of current versus time is as shown in the Figure 3.22. The current is  $180^\circ$  out of phase with the voltage and it ranges from -30mA to 30mA for a  $50\Omega$  load and 3V peak to peak input voltage.

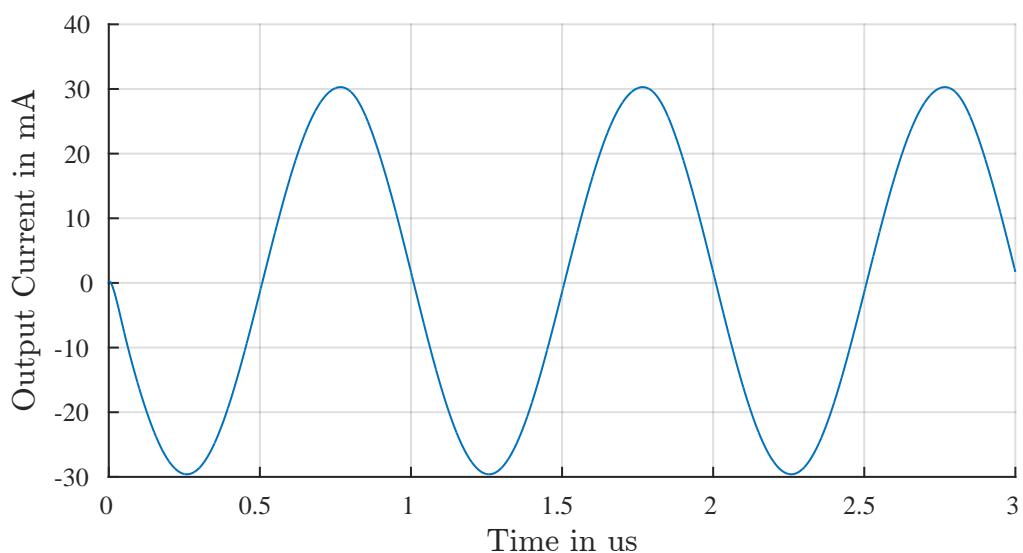


Figure 3.22.: OPAMP Plot of Ourput Current vs time

The only change from the previous test bench to this is that we need to replace the sinusoidal source by a pulse input source.

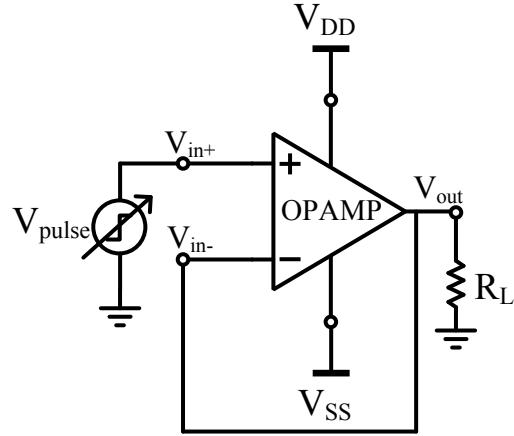


Figure 3.23.: Test setup for Transient Analysis - Square Wave Input

The values of all the parameters of the op amp discussed so far have been tabulated in Table 3.12

Parameter	Value
Open Loop Gain	30.8 dB
Gain Bandwidth Product	6.2 MHz
Phase Margin	76.79
ICMR (min)	-2.19 V
ICMR (max)	2.089 V
Output Current (max)	-29.6 mA
Output Current (min)	30.28 mA
Output Voltage Swing	-2.19 .. 2.323
Slew Rate	10 V/us
PSRR (VDD)	30.96 uA/V
PSRR (VSS)	138.8 uA/V
Input Impedance	9.04 MOhms
Output Impedance	4.167 Ohms

Table 3.12.: Simulation Results of the OPAMP

### 3.3 The Complete Design

The block diagram of the two stage design with an OTA and an OP AMP is as shown in Figure 3.24. The other end of the capacitor of the first stage is connected to the negative power supply instead of ground since we have only two levels in the IC (-2.5V and 2.5V). Both the stages use the same power supply. The output of the OTA is directly connected to the non-inverting terminal of the OP AMP without any DC isolation via a coupling capacitor.

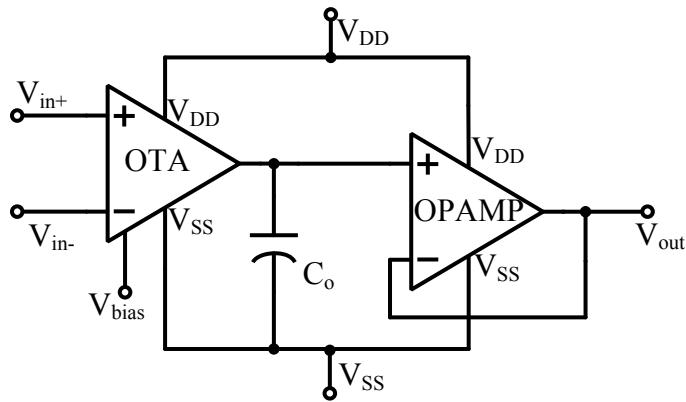


Figure 3.24.: Block Diagram of the Overall System

The schematic symbol for the overall system is as indicated in Figure 3.25. The block diagram of the system is drawn inside to get an understanding of the configuration.

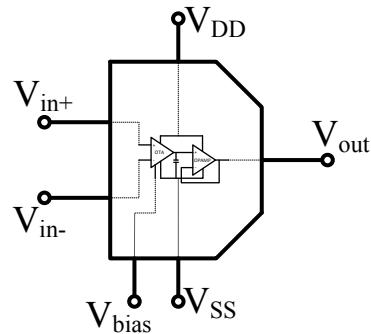


Figure 3.25.: Schematic Symbol for the Overall System

### 3.3.1 Schematic

The transistor level schematic of the overall system is as shown in the Figure 3.26. The upper portion of the schematic is the OTA and the bottom portion is the OP AMP. Note the feedback connection of the OP AMP and the connection between the OTA and the OP AMP. To summarize the physical aspects of the system, there are 3 input terminals, 2 power supply levels and 1 output pin.

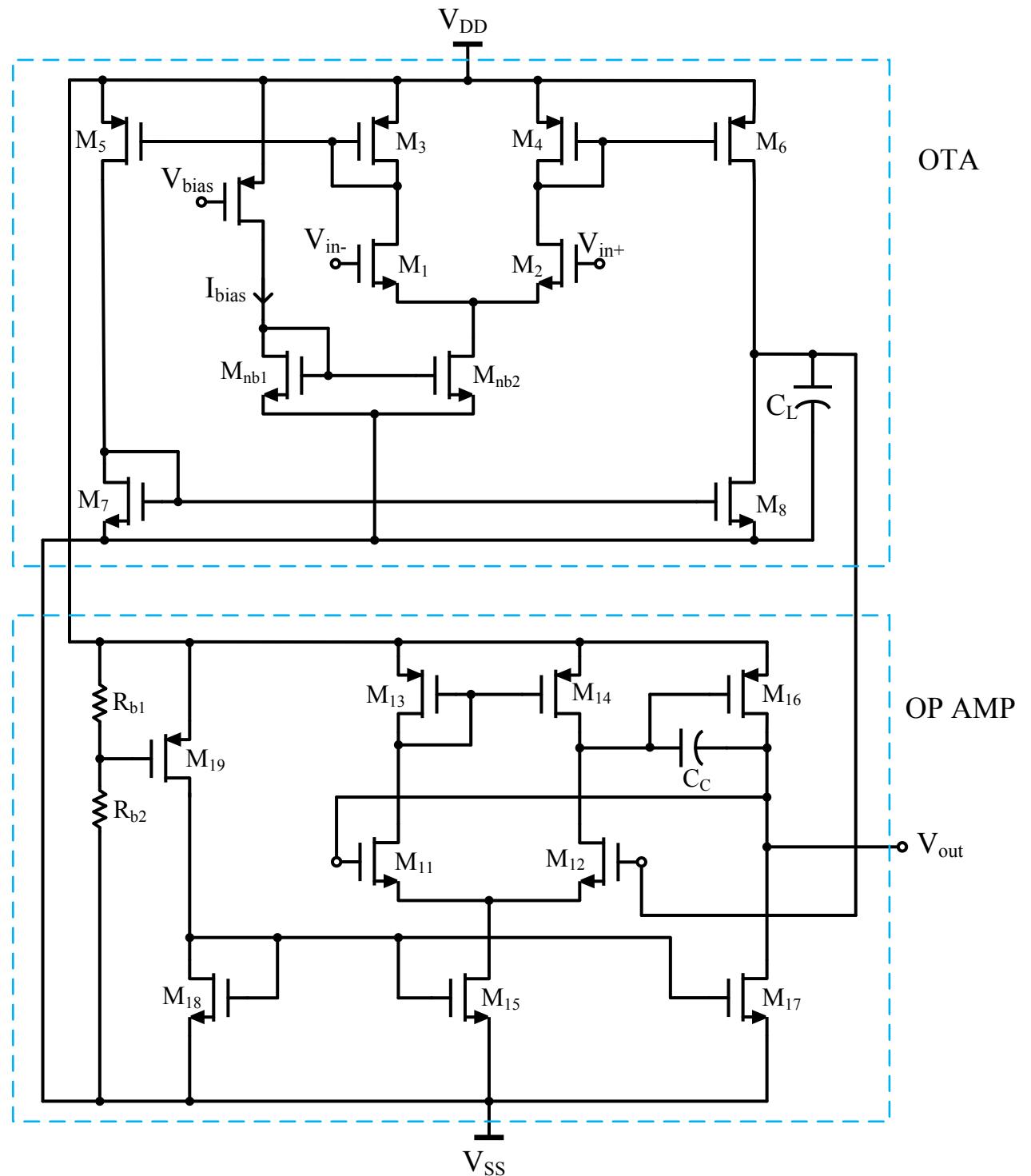


Figure 3.26.: Schematic Diagram for the Overall System

## 4 Simulation Results

In this chapter, the simulation results for the overall system designed are discussed. And also the behavior of the amplifier for process variation, power supply variation and temperature variation is discussed.

### 4.1 DC Analysis

The test bench to perform the DC analysis is as shown in the Figure.4.1.

$V_{DD} = 2.5V$ ;  $V_{SS} = -2.5V$ ;  $V_{cm} = 1.95V$ ;  $V_{bias} = 150mV$  to  $700mV$ ;  $V_{ac} = 1V$ ;  $R_L = 50\Omega$ .

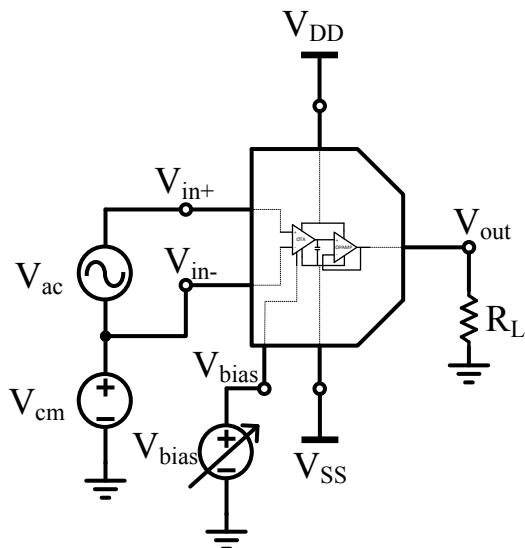


Figure 4.1.: Test setup for DC Analysis

The DC Bias point at the output of the system is tabulated in the Table.4.1. Output DC bias points are almost symmetric around 0V and vary from 180.1mV to -219.3mV.

Vbias (mV)	Output DC Bias (mV)
150	180.1
200	148.4
250	115.8
300	82.28
350	47.83
400	12.45
450	-23.86
500	-61.08
550	-99.24
600	-138.3
650	-178.4
700	-219.3

Table 4.1.: DC Bias Point at the output of the circuit

## 4.2 AC Analysis

The test bench in Figure.4.1 is used for the AC analysis too. The magnitude of the AC source is set to 1V and as mentioned in the previous chapter, the voltage at the output will be the gain of the overall system.

The plot of gain for different values of  $V_{bias}$  is as shown in the Figure.4.2. The value of the open loop gain is between 18dB and 25dB.

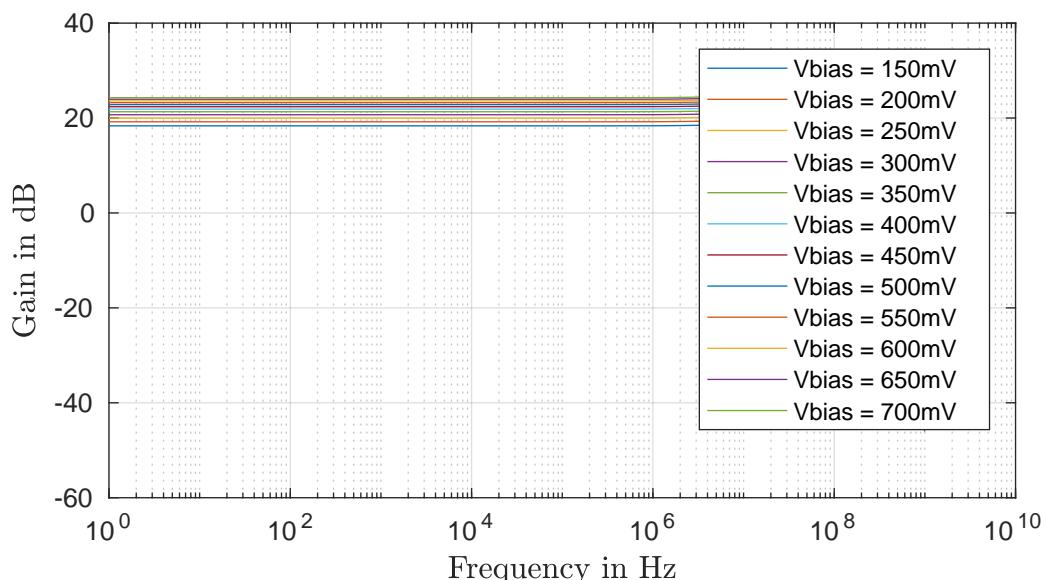


Figure 4.2.: Plot of Gain vs Frequency for different Vbias

The results of the AC analysis are tabulated below in Table.4.2. The bandwidth is fairly constant at around 14MHz. But the phase margin starting to drop just a tad below 30 with a  $V_{bias}$  of 700mV and beyond.

Vbias (mV)	DC Gain (dB)	Bandwidth (MHz)	Phase Margin
150	18.5	14.43	55.49
200	19.34	14.38	51.37
250	20.13	14.34	47.76
300	20.83	14.29	45.08
350	21.46	14.24	42.55
400	22.02	14.19	40.19
450	22.52	14.14	38
500	22.97	14.08	35.97
550	23.38	14.01	34.08
600	23.75	13.95	32.31
650	24.1	13.88	30.63
700	24.43	13.8	29.02

Table 4.2.: DC Gain, Bandwidth and Phase Margin of the Overall System

The test bench in Figure 4.3 is used to measure the input impedance of the system. The concept for measurement is same as in the case of the OTA. The voltage at the non-inverting terminal will be the magnitude of the input impedance of the overall system because the magnitude of the current from the AC source is 1A.

$$V_{DD} = 2.5V; V_{SS} = -2.5V; V_{bias} = 150mV \text{ to } 700mV; I_{sin magnitude} = 1A; R_L = 50\Omega.$$

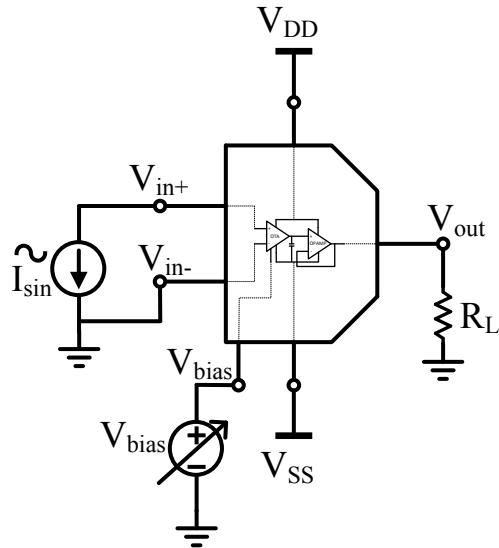


Figure 4.3.: Test setup to measure Input Impedance

The value of input impedance with respect to Vbias is tabulated in Table 4.3. It is clear that the input impedance of the overall system is same as the input impedance of the OTA in the first stage.

Vbias (mV)	Input Impedance ( $M\Omega$ )
150	3.394
200	3.402
250	3.41
300	3.418
350	3.425
400	3.433
450	3.44
500	3.448
550	3.456
600	3.464
650	3.473
700	3.482

Table 4.3.: Input Impedance of the Overall System

On the same lines, the test bench for measuring the output impedance is as shown in the Figure.4.4. The resistive load is replaced by a current source which tries to pull out 1A of current from the circuit and thereby the voltage at the output turning out to be the magnitude of the output impedance of the system.

$V_{DD} = 2.5V$ ;  $V_{SS} = -2.5V$ ;  $V_{cm} = 1.95V$ ;  $V_{bias} = 150mV$  to  $700mV$ ;  $I_{sin magnitude} = 1A$ .

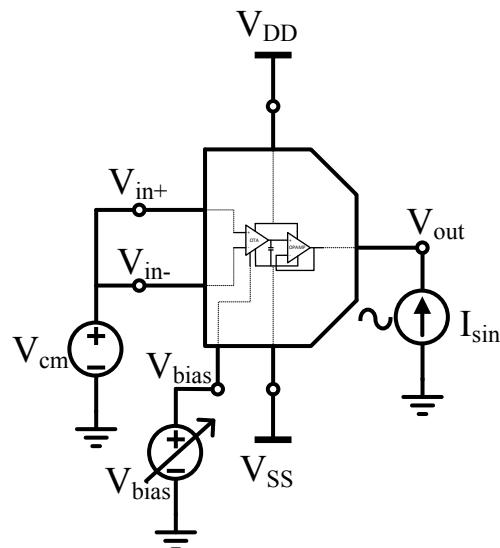


Figure 4.4.: Test setup to measure Output Impedance

The value of the output impedance is very low because of the output buffer and there is a small variation of output impedance with change in  $V_{bias}$  and is tabulated in Table.4.4. The values are between  $0.94\Omega$  and  $1\Omega$ .

Vbias (mV)	Output Impedance ( $\Omega$ )
150	0.9415
200	0.9434
250	0.9453
300	0.9474
350	0.9495
400	0.9517
450	0.9541
500	0.9565
550	0.9591
600	0.9618
650	0.9646
700	0.9675

Table 4.4.: Output Impedance of the Overall System

The next important parameter is the PSRR or the Power Supply Rejection Ratio. It was seen in the previous chapter that the PSRR of the second stage is much more significant than the first stage. Figure 4.5 shows the test bench used to measure the PSRR of the whole system. The schematic on the left is used to measure PSRR for  $V_{DD}$  and likewise on the right side for  $V_{SS}$ .

$V_{DD} = 2.5V$ ;  $V_{SS} = -2.5V$ ;  $V_{cm} = 1.95V$ ;  $V_{bias} = 150mV$  to  $700mV$ ;  $V_{ac\ magnitude} = 1V$ ;  $R_L = 50\Omega$ .

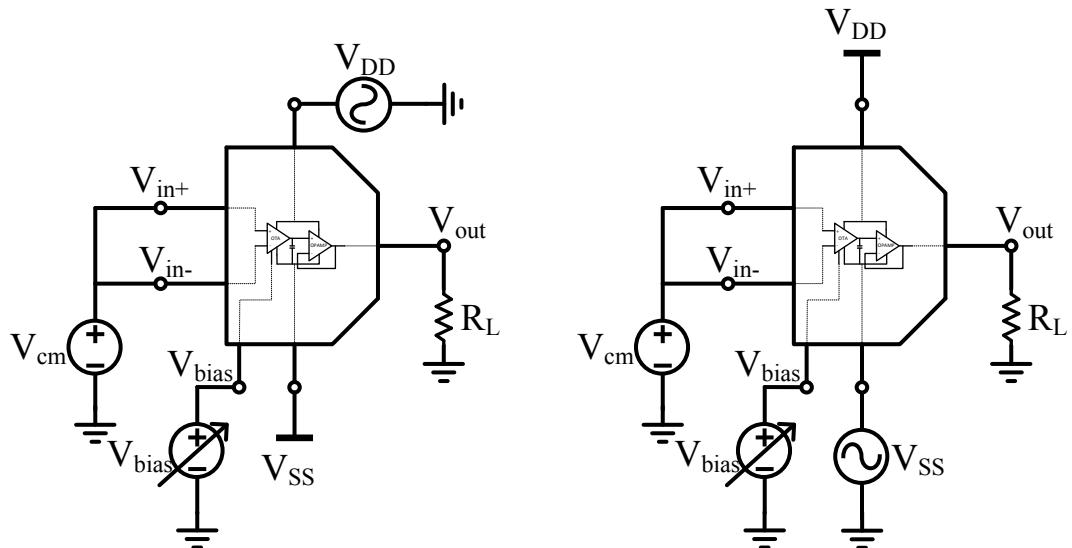


Figure 4.5.: Test setup to measure PSRR

The variation of PSRR against  $V_{bias}$  is tabulated below in Table 4.5. PSRR is high for high bias currents or low values of  $V_{bias}$  and it decreases with increase in  $V_{bias}$ . PSRR is generally expressed in terms of dB. But from the specifications, it is seen that the parameter has to be expressed in uA/V as part of this work.

Vbias (mV)	PSRR (VDD)(uA/V)	PSRR (VSS)(uA/V)
150	97.76	93.69
200	89.66	85.77
250	82.86	79.11
300	77.26	73.59
350	72.68	69.04
400	68.93	65.28
450	65.84	62.18
500	63.27	59.53
550	61.1	57.26
600	59.22	55.28
650	57.59	53.52
700	56.13	51.92

Table 4.5.: PSRR of the Overall System

#### 4.3 Transient Analysis

The transient analysis is divided in to two parts. The first part involves simulation with a sine wave input and the second part involves simulation with a square wave input. The test bench for transient analysis with a sine wave input for the system is as shown in Figure.4.6.

$$V_{DD} = 2.5V; V_{SS} = -2.5V; V_{cm} = 1.95V; V_{bias} = 150mV \text{ to } 700mV; V_{sin amplitude} = 100mV; R_L = 50\Omega.$$

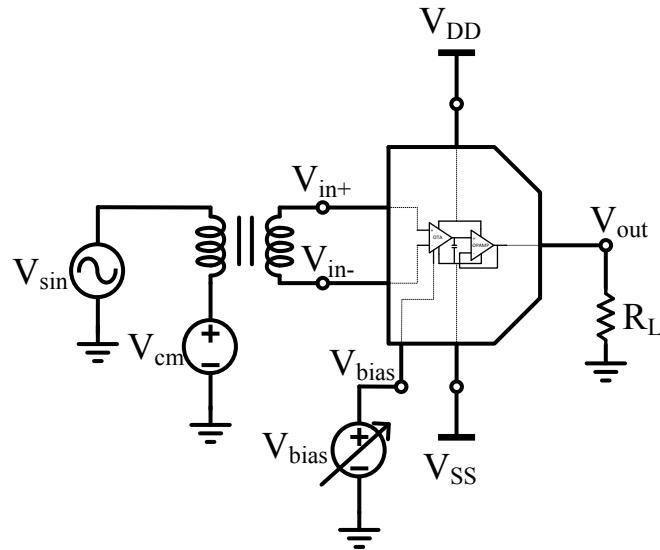


Figure 4.6.: Test setup for Transient Analysis - Sine Wave input

The plot of output current against time for different values of  $V_{bias}$  is as shown in the Figure.4.7. The peaks are not symmetric around zero, however it is made sure that the peak-to-peak current values range from 30mA to 60mA.

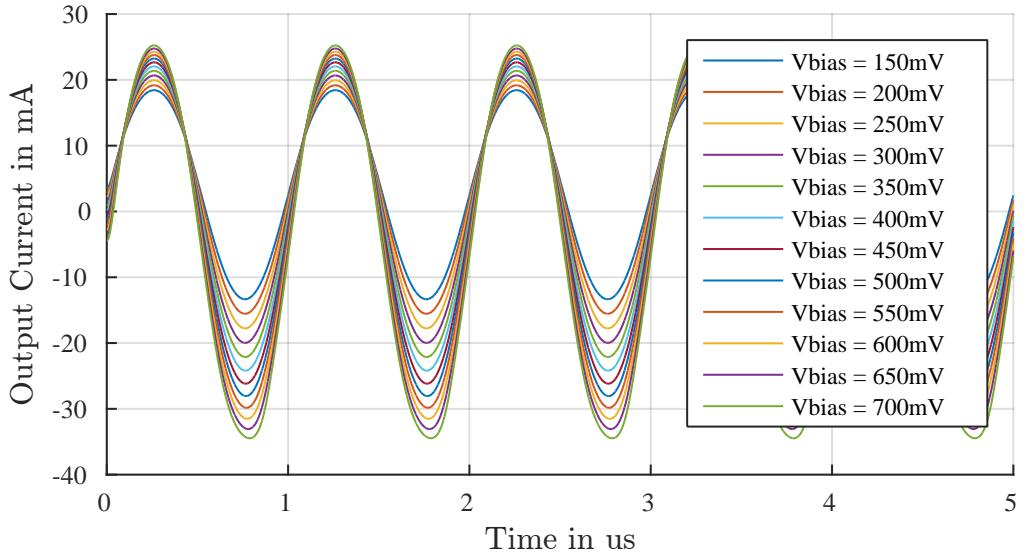


Figure 4.7.: Plot of Output Current vs time for different  $V_{bias}$

Table 4.6 gives a better picture of the sinusoidal plots shown above. The HD2 and HD3 values are less than 30 dBc. However, the HD3 seems to take a turn towards the worse with increase in  $V_{bias}$ . So for  $V_{bias}$  values beyond 700mV, the output starts clipping at the trough of the current wave.

$V_{bias}$ (mV)	Iout Max(mA)	Iout Min(mA)	Iout p2p(mA)	HD2 (dBc)	HD3 (dBc)
150	18.43	-13.34	31.77	-37.14	-48.17
200	19.17	-15.55	34.72	-36.62	-45.67
250	19.93	-17.77	37.7	-35.65	-43.65
300	20.67	-19.97	40.64	-35.15	-42.07
350	21.38	-22.12	43.49	-34.84	-40.83
400	22.04	-24.19	46.23	-34.72	-39.85
450	22.66	-26.17	48.83	-34.8	-39.01
500	23.24	-28.05	51.29	-35.08	-38.24
550	23.77	-29.83	53.6	-35.54	-37.42
600	24.28	-31.51	55.79	-36.13	-36.47
650	24.76	-33.07	57.83	-36.61	-35.24
700	25.24	-34.47	59.71	-36.44	-33.61

Table 4.6.: Transient Parameters of the Overall System

Once the peak-to-peak value of current is obtained, we can use that to calculate the transconductance of the system for different  $V_{bias}$  values. Transconductance is the ratio of output current to the input voltage. Since the output current is not symmetric around 0A, the the value half of peak-to-peak current is considered. That divided by the input voltage (100mV peak) would give us the required transconductance value. Figure 4.8 shows the plot of transconductance against  $V_{bias}$ . And it rightfully follows the behavior of the output current.

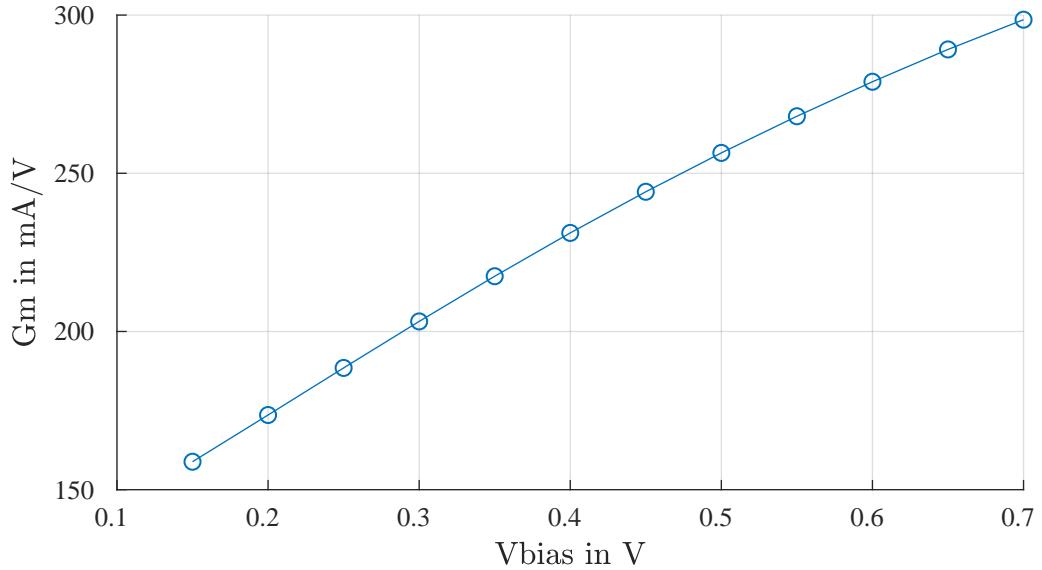


Figure 4.8.: Plot of Gm vs Vbias

The transconductance value for each  $V_{bias}$  used to plot the Figure.4.8 is tabulated in Table.4.7.

Vbias (mV)	Transconductance (mA/V)
150	158.8
200	173.6
250	188.5
300	203.2
350	217.5
400	231.2
450	244.2
500	256.4
550	268
600	278.9
650	289.1
700	298.5

Table 4.7.: Transconductance Gain of the Overall System

Now, the second part of the transient analysis is performed with a square wave input. Figure.4.9 shows the test bench for transient analysis for a square wave input. The system is connected in a unity gain configuration.

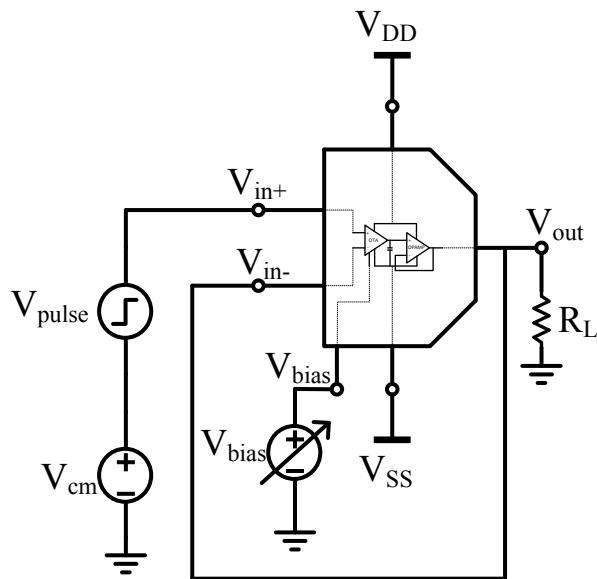


Figure 4.9.: Test setup for Transient Analysis - Square Wave input

The slew rate for the rising edge and the falling edge is tabulated in the Table 4.8. The slew rate is at a low value due to the fact that the OP AMP designed exhibits a very low slew rate.

Vbias (mV)	Slew Rate (Rising Edge)(V/us)	Slew Rate (Falling Edge)(V/us)
150	6.985	-8.73
200	7.505	-9.779
250	8.042	-10.68
300	8.581	-11.34
350	9.095	-11.66
400	9.561	-11.7
450	9.954	-11.6
500	10.27	-11.45
550	10.49	-11.3
600	10.65	-11.14
650	10.76	-10.99
700	10.82	-10.84

Table 4.8.: Slew Rate of the Overall System

#### 4.4 Noise Analysis

The test bench to perform the noise analysis is same as the test bench used for AC analysis in Figure 4.1. The input referred noise of the first stage is far more significant for the overall system than the second stage.

The Bode plot of the input referred noise for different values of  $V_{bias}$  is shown in the Figure 4.10. It is clear that the type of noise is 1/f noise or the Flicker noise. The magnitude of the input referred noise is very high for low frequencies and it reduces significantly in moderate and high frequencies when White noise starts to dominate.

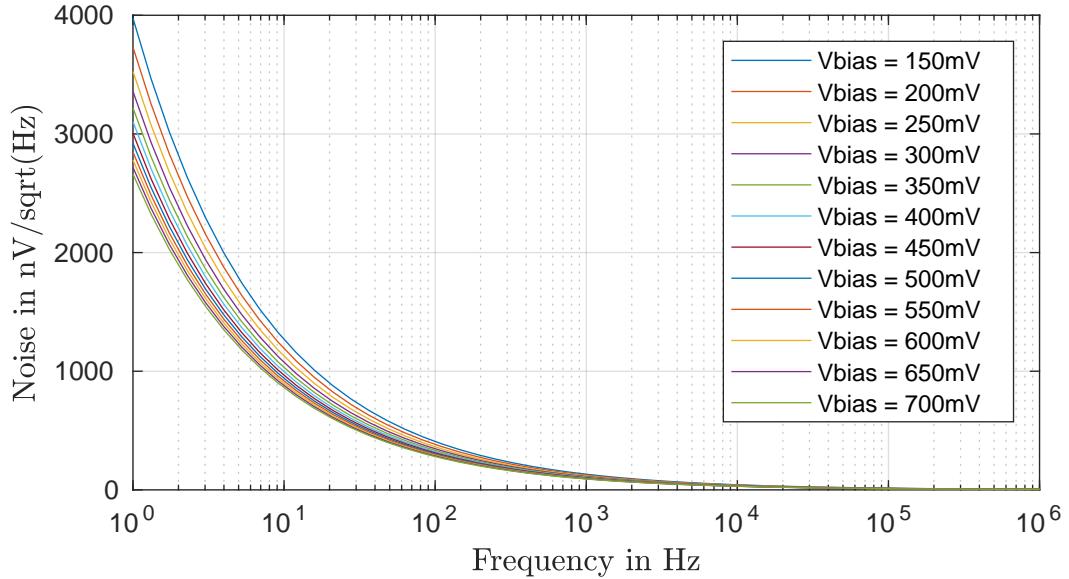


Figure 4.10.: Plot of Input Referred Noise vs Frequency for different  $V_{bias}$

The value of input referred noise at 1MHz for different  $V_{bias}$  is tabulated in the Table 4.9.

$V_{bias}$ (mV)	Input Referred Noise (nV/sqrt(Hz))
150	46.64
200	43.93
250	41.71
300	39.91
350	38.46
400	37.29
450	36.33
500	35.52
550	34.82
600	34.21
650	33.66
700	33.16

Table 4.9.: Input Referred Noise of the Overall System

## 4.5 Programmable Load

Another way to make the overall system programmable is to have a programmable load. Having a Resistive load that varies would imply that the current flowing through the load is controlled. So, for the following analyses, the  $V_{bias}$  is made constant and set a value of 425mV.

Table 4.10 shows the variation of DC Gain, bandwidth, phase margin and transconductance with varying load. The gain hardly changes with the changing load. The variation in bandwidth is just 1.8MHz between the minimum and maximum load. The transconductance however, shows variation in a range that is expected.

RL ( $\Omega$ )	DC Gain(dB)	Bandwidth(MHz)	Phase Margin	Transconductance(mA/V)
35	22.2	13.13	40.97	338.9
40	22.22	13.55	40.22	296.8
45	22.25	13.88	39.6	264
50	22.28	14.16	39.08	237.7
55	22.3	14.4	38.63	216.2
60	22.32	14.6	38.24	198.3
65	22.34	14.77	37.9	183.1
70	22.35	14.92	37.6	170

Table 4.10.: AC Parameters of the Overall System for a programmable load - 1

Table 4.11 shows the tabulation of values for the parameters that are independent of the variation in load resistance.

Parameter	Value
Input Impedance M( $\Omega$ )	3.436
Output Impedance ( $\Omega$ )	0.953
PSRR (VDD)(uA/V)	67.32
PSRR (VSS)(uA/V)	63.65
Input Referred Noise (nV/ $\sqrt{Hz}$ )	36.79

Table 4.11.: AC Parameters of the Overall System for a programmable load - 2

The transient analysis of the system with a programmable load shows a clear picture of how effective the system is. Since the bias voltage or the bias current doesn't change, the DC bias point at the output of the system remains unchanged. Therefore, we get a symmetrical waveform around 0A. The plot in Figure 4.11 shows the output current plot for different values of load resistance.

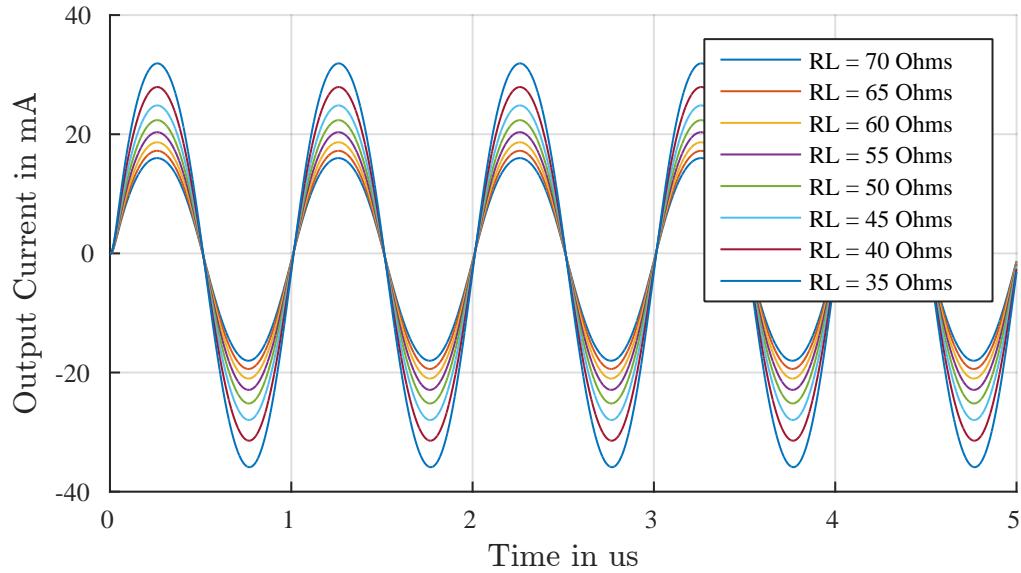


Figure 4.11.: Plot of Output Current vs Time for different RL

Table.4.12 contains tabulation from the plot in Figure.4.11. Output current is the most significant parameter that gets affected by a variable load.

RL ( $\Omega$ )	Iout(max)(mA)	Iout(min)(mA)	Iout(p2p)(mA)
35	31.89	-35.89	67.77
40	27.92	-31.44	59.36
45	24.83	-27.97	52.8
50	22.36	-25.19	47.55
55	20.33	-22.91	43.24
60	18.65	-21.01	39.65
65	17.22	-19.4	36.61
70	15.99	-18.02	34.01

Table 4.12.: Maximum and Minimum Output Currents for a programmable load

Table.4.13 shows the values of the parameters that are constant with a varying load. HD2 and HD3 are less than -30dBc.

Parameter	Value
HD2 (dBc)	-34.6
HD3 (dBc)	-39.4
Slew Rate (Rising Edge)(V/us)	10.28
Slew Rate (Falling Edge)(V/us)	-10.45

Table 4.13.: Transient Parameters of the Overall System for a programmable load

And finally, we have the plot of transconductance versus the load resistance as shown in Figure 4.12

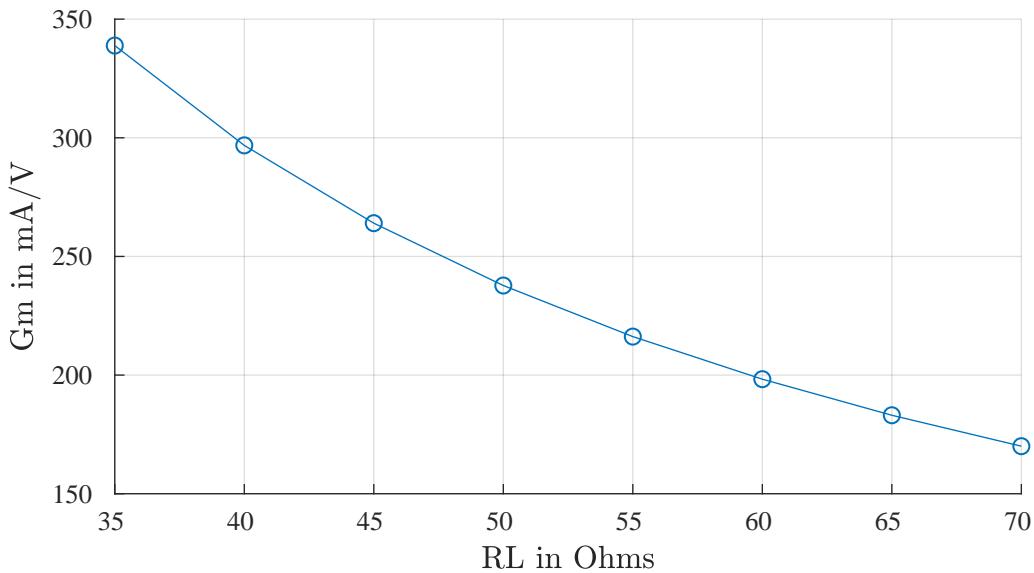


Figure 4.12.: Plot of Transconductance vs RL

## 4.6 Corner Simulation

Corner simulations are performed to verify that the circuit works at all possible conditions of process, supply voltage and temperature. Analog circuit designs that perform sufficiently well across these corners are to be considered a robust design.

### 4.6.1 Process Variation

Process variation is the naturally occurring variation in the attributes of transistors (length, widths, oxide thickness) when ICs are fabricated. In general there are 5 corners - tt, ff, ss, fs, sf. Transistors can be typical, fast or slow. The first letter indicates the PMOS and the second letter indicates the NMOS. However, as part of XT018 technology, the corners are termed as - tm, wp, ws, wo, wz. The meaning and their translation to the general conventions are described in Table 4.14. The process corners of Capacitors and Resistors too follow a similar naming convention but with only 3 corners - tm (typical), wp(minimum), ws(maximum).

Corner	Expansion	Translation
tm	Typical Mean Condition	tt
wp	Worst Case Power Condition	ff
ws	Worst Case Speed Condition	ss
wo	Worst Case One Condition	fs
wz	Worst Case Zero Condition	sf

Table 4.14.: Different Corners for MOS Transistors in XT018 Technology

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A total of 45 corners are simulated for 3 different values of  $V_{bias}$  - 150mV, 400mV and 700mV. And the results are provided in the subsequent sections.

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#### 4.6.2 Process and Supply Variation

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In addition to process variation, the system responds to power supply variations too. 3 different corners are obtained by evaluating the power supply with  $\pm 10\%$  from the nominal value. Therefore, we get 3 corners for  $V_{DD}$  - 2.5V, 2.25V, 2.75V and 3 corners for  $V_{SS}$  - -2.5V, -2.25V, -2.75V. The total number of corners in this case are 405. And the results are discussed in the subsequent sections.

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#### 4.6.3 Process, Voltage and Temperature (PVT) variation

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Another possible variation is seen in the temperature parameter. The PVT corner simulations are performed at the following corners:

- All Process corners mentioned in Section 4.6.1.
- $\pm 10\%$  variations of  $V_{DD}$  i.e., 2.5V, 2.25V, 2.75V and  $V_{SS}$  i.e., -2.5V, -2.25V, -2.75V.
- Three temperature conditions:  $-25^0C$ ,  $25^0C$  and  $80^0C$ .

Therefore there are 1215 corners in total. The results are provided in the subsequent sections.

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#### 4.6.4 Summary of PVT Corner Analysis

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# 5 Conclusion

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5.1 Summary of Results

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5.2 Outlook

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## A Appendix

