Quadrature Down Converter

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***Abstract*—This project focuses on designing a Quadrature Down Converter for RF signal processing. It converts high-frequency RF signals to low-frequency baseband signals using I/Q demodulation. The system employs mixers, local oscillators, and low-pass filters to extract in-phase and quadrature components.**

***Index Terms*—Quadrature Oscillator, Mixer, Low Pass Filter, MOSFETs**

I. Introduction

In wireless communication systems, signals are often transmitted at very high frequencies. However, processing these signals directly at such high frequencies is difficult and inefficient. To make signal processing easier, we use down-conversion, which shifts the signal to a lower frequency. A Quadrature Down Converter is a special type of down converter that breaks the signal into two parts — called the In-phase (I) and Quadrature-phase (Q) components, which are 90 degrees out of phase with each other. This splitting helps store the information about the signal's phase, which is useful during signal reconstruction.

This project aims to design and implement a basic Quadrature Down Converter that can take a high-frequency input signal and convert it into its I and Q components.

As part of this project, we will study how mixers, local oscillators, and filters work together in a QDC to shift a signal's frequency and separate it into its I and Q parts. We will also explore how the correct phase relationship between the I and Q signals is crucial for accurate signal reconstruction and further processing.

II. Working of a Quadrature Down Converter

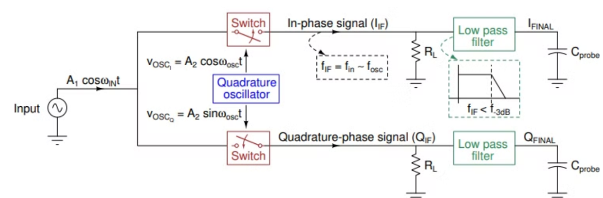

Fig. 1: Block diagram of the quadrature down-converter showing the I/Q paths, LO signals, and low-pass filtering.

A quadrature downconverter is an essential component in modern radio, RF, and communication systems. It is responsible for translating a high-frequency radio signal (RF) into a lower intermediate frequency (IF). This process preserves the original signal's amplitude and phase information, which is crucial for subsequent digital processing and demodulation.

The quadrature downconverter operates by mixing the incoming RF signal with a locally generated oscillator (LO) signal. The LO signal is derived from a quadrature oscillator, which generates two outputs:

- An in-phase signal: $v_{\mathrm{OSC}_1} = A_2 \cos(\omega_{\mathrm{OSC}} t)$
- A quadrature-phase signal: $v_{\mathrm{OSC}_Q} = A_2 \sin(\omega_{\mathrm{OSC}} t)$

These two signals are 90 degrees out of phase with each other. The input RF signal, represented as $A_1 \cos(\omega_{\mathrm{IN}} t)$, is mixed simultaneously with both components of the LO signal through two separate mixers.

- The mixer using the in-phase LO signal yields the in-phase intermediate frequency component (I_{IF}).
- The mixer using the quadrature LO signal produces the quadrature intermediate frequency component (Q_{IF}).

The outputs of both mixers are then passed through low-pass filters (LPFs), which eliminate high-frequency components resulting from the mixing process. This filtering ensures that only the desired baseband or IF signals remain:

- I_{FINAL} represents the filtered in phase signal.
- Q_{FINAL} represents the filtered quadrature phase signal.

The key advantage of this approach is that the combination of the I and Q outputs allows for a complete reconstruction of the amplitude and phase of the original signal. This characteristic makes quadrature down-conversion particularly suitable for use in software-defined radios, digital receivers, and radar systems, where precise signal representation is critical.
The three main components involved in this system are:

1) **Quadrature Oscillator (QO)** – generates orthogonal LO signals.
2) **Mixers** – perform the multiplication of RF and LO signals.
3) **Low-Pass Filters (LPFs)** – extract the desired IF/baseband signals.

III. Quadrature Oscillator

The canonical form of a feedback system is shown in Figure 2.

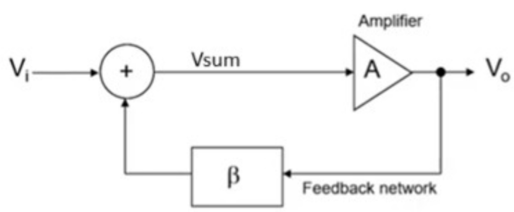

Fig. 2: Canonical form of feedback system

Here, A = open loop gain of the op-amp and $A\beta$ = loop gain.

$$V_o = A \cdot V_{\text{sum}} = A \cdot (V_i + \beta V_o)$$

$$\Rightarrow \frac{V_o}{V_i} = \frac{A}{1 - A\beta} \tag{1}$$

Oscillation results from an unstable state; i.e., the feedback system can't find a stable state because its transfer function can't be satisfied. Equation (1) becomes unstable when $(1 - A\beta) = 0$ because $A/0$ is an undefined state. So, the key to designing an oscillator is to ensure that $A\beta = 1$. This is called the Barkhausen criterion for oscillation.

In further analysis, we will see that loop gain, i.e. $A\beta$, is 1 only for one frequency due to reactive elements like capacitors. This is frequency selection. The Barkhausen criterion also says that the phase shift of the loop gain should be equal to 0° (or 360°).

Quadrature Oscillator Topology

The quadrature oscillator is used to produce two sinusoidal signals with a phase difference of 90°. The topology we used requires three op-amps, which produce two sinusoidal signals with a 90° phase difference.

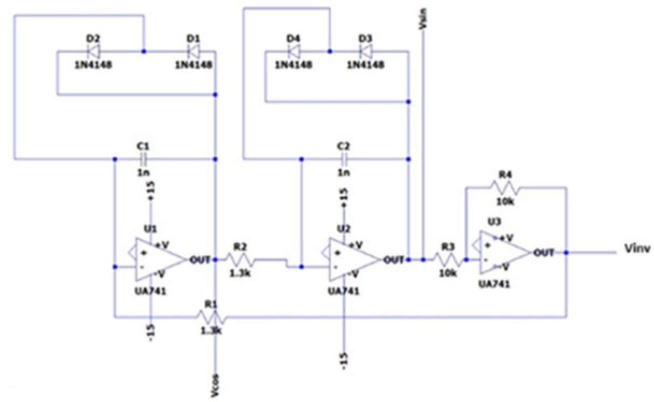

Fig. 3: Quadrature Oscillator Circuit Diagram

This topology can be divided into 3 sub-parts, each consisting of one op-amp. The oscillator uses 3 op-amp circuits. The first two (from left to right) act as integrators in negative feedback, while the third op-amp (on the extreme right) acts as an inverter (an inverting amplifier with gain = 1).

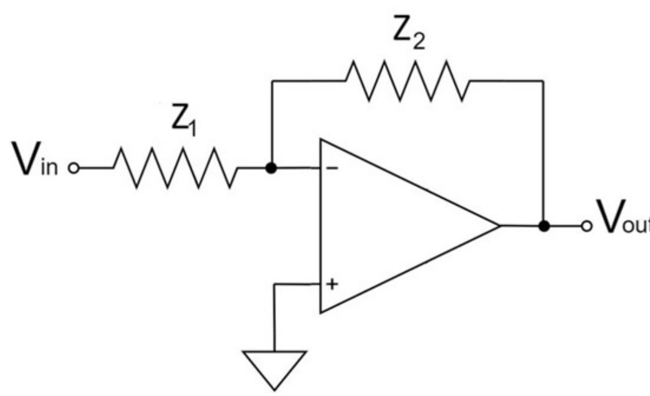

Fig. 4: Circuit diagram of an inverting amplifier

For an amplifier with the following setup:

$$V_{\text{out}} = -\left(\frac{Z_2}{Z_1}\right) \cdot V_{\text{in}}$$

For a capacitor with capacitance C, Z (impedance) = $\frac{1}{j\omega C}$
So, for the first 2 sub-parts, we get:

$$V_{\text{out}}(t) = -\frac{1}{RC}\int_0^t V_{\text{in}}(\tau)\,d\tau + V_{\text{out}}(0)$$

i.e., in Laplace domain:

$$\frac{V_{\text{out}}(s)}{V_{\text{in}}(s)} = -\frac{Z_2}{Z_1} = -\frac{1}{sRC}$$

These two integrators provide a phase shift of 180° $(90° + 90°)$. But this fails the Barkhausen criterion to get 0°. So, we add the third block, i.e., the inverter, adding another 180°

phase shift, resulting in the total phase shift for the loop gain to be $360°$, which is effectively $0°$.

Analysis of the Circuit

We see that the output of the third op-amp is the feedback input for the first op-amp. The output of the first op-amp is the input for the second op-amp, and so on. Applying the above results:

$$V_{\text{COS}} = -\frac{1}{R_1 j\omega C_1} \cdot V_{\text{INV}}$$

$$V_{\text{SIN}} = -\frac{1}{R_2 j\omega C_2} \cdot V_{\text{COS}}$$

$$V_{\text{INV}} = -\frac{R_4}{R_3} \cdot V_{\text{SIN}}$$

So, the loop gain for the circuit is:

$$A\beta = \frac{R_4}{R_1 R_2 R_3 C_1 C_2 \omega^2}$$

Taking $R_1C_1 = R_2C_2 = RC$ and $R_4 = R_3$ for inverter action, the loop gain equation simplifies to:

$$A\beta = \frac{1}{(RC \cdot \omega)^2}$$

From the Barkhausen criterion for sustained oscillation, $|A\beta| = 1$. So,

$$\omega = \frac{1}{RC}, \quad \text{and hence,} \quad f = \frac{1}{2\pi RC}$$

Now, the desired frequency is 100 kHz. Choosing an arbitrary capacitor with capacitance $C = 1$ nF, we get $R = 1.59$ kΩ.

But when simulated on LTspice, we observed a frequency lower than the expected 100 kHz due to the op-amp bandwidth and phase shift. The UA741 op-amps in the circuit have a finite gain-bandwidth product and phase delay. These factors modify the effective frequency response of the circuit. The UA741 op-amp is not ideal and can introduce deviations in the oscillator's behavior at high frequencies. The UA741 op-amp bandwidth is around 1 MHz, and 100 kHz is close to this limit. So, stray capacitances, op-amp slew rates, etc., can cause the frequency to shift slightly from the theoretical value.

On performing various simulations on LTSpice we found out that for resistance value of $R = 1.3$ kΩ, we obtain an output frequency of 100KHz.

Automatic Gain Control (AGC)

During the initial phase, the nonlinear clipping of the diodes determines amplitude stabilization. This nonlinear behavior can indirectly affect the oscillation frequency.

We used diodes for Automatic Gain Control (AGC). AGC is used in circuits to maintain a consistent output level despite input variations. It ensures oscillation amplitude does not grow uncontrollably or decay.

The oscillator consists of two integrator stages followed by an inverting amplifier. In each integrator, a pair of diodes (e.g., D1 and D2, D3 and D4) are connected in a back-to-back direction in the feedback loop, forming a symmetrical nonlinear feedback path.

At low output amplitudes, the diodes are reverse-biased and do not conduct, allowing the integrators to operate linearly. As amplitude increases, the voltage exceeds the diode threshold ($\tilde{0}$.7V for silicon diodes), and the diodes begin to conduct. This limits the gain and stabilizes the amplitude without significant distortion.

Advantages of This AGC Method

- Requires no additional active components
- Preserves phase relationships needed for oscillation
- Automatically adapts to component tolerances and power supply variations

Limitations

- Small capacitances and nonlinearities
- Minor deviations in frequency and phase due to non-idealities

These factors cause the observed phase shift to deviate slightly from the ideal $90°$.

Simulation Results

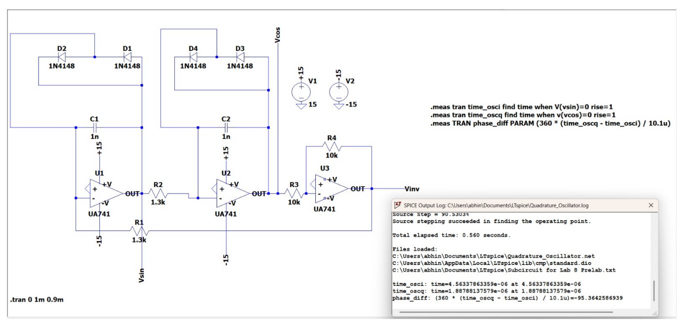

Fig. 5: Phase difference representation in SPICE Error LOG

Simulated phase difference = $95.36°$

This is very close to $90°$, but not exactly due to:

- Limited gain-bandwidth product of UA741
- Nonlinear clipping by diodes introducing slight signal delay

LT SPICE Simulations

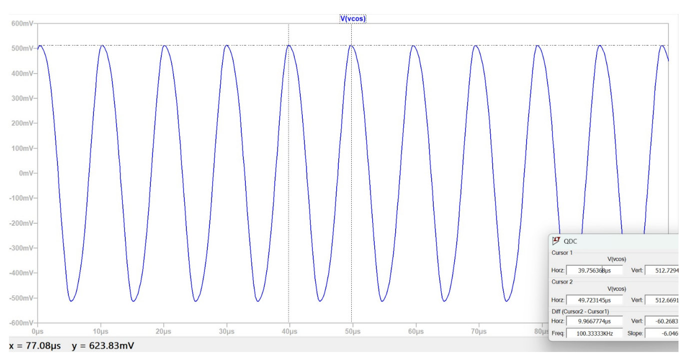

Fig. 6: Transient analysis of in-phase component

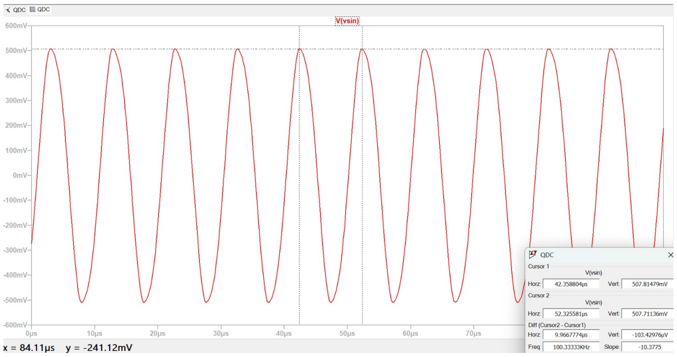

Fig. 7: Transient analysis of q-phase component

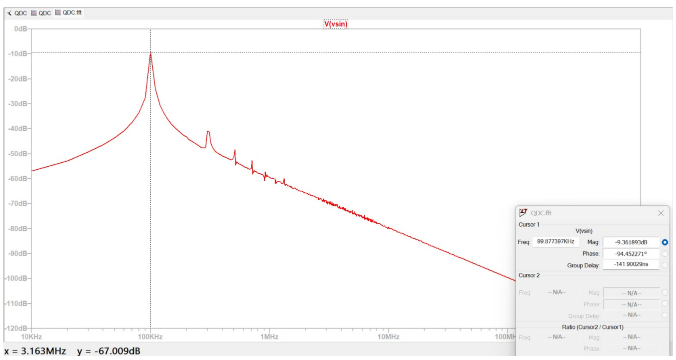

Fig. 8: FFT of q-phase component

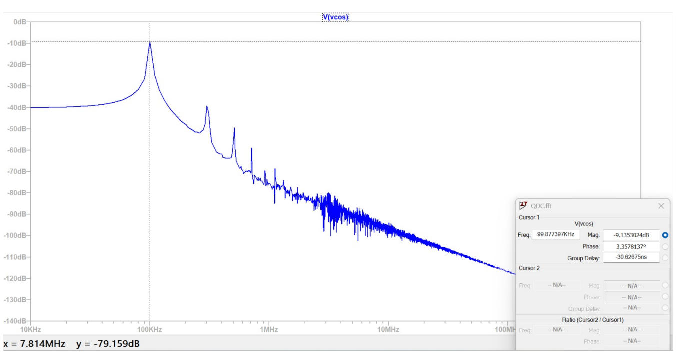

Fig. 9: FFT of in-phase component

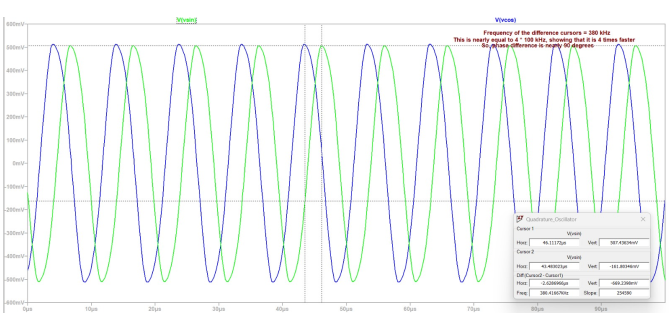

Fig. 10: Comparing phase difference of Vsin and Vcos

IV. SWITCH/MIXER

Mixer Calculations

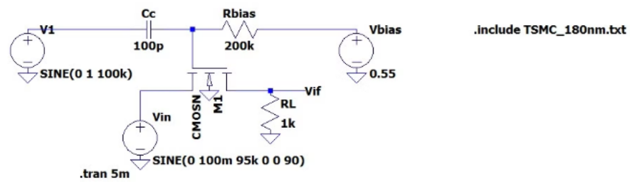

Fig. 11: Mixer circuit diagram

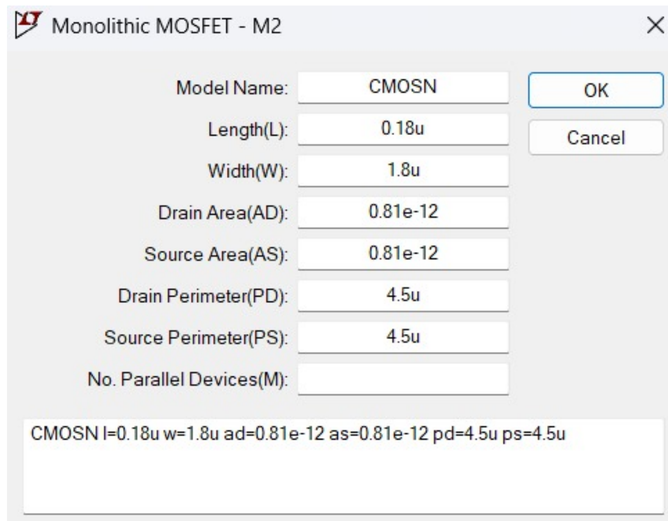

Fig. 12: L,W values for the NMOS

Mixer circuits function by multiplying two input signals, with the output being the product of these inputs. An NMOS device operates effectively as a mixer when it is biased in the linear region, where one input signal is applied to the gate, the other to the source, and the output is extracted from the drain. To achieve this condition, a bias voltage V_{bias} is applied to the gate such that it is close to but not equal to the threshold voltage V_{TH}, effectively linearizing the term $V_{\text{gs}} - V_{\text{TH}}$ in the drain current equation. When V_{bias} is set appropriately, the term $V_{\text{gs}} - V_{\text{TH}}$ can be approximated as $V_{\text{osc}} - V_{\text{in}}$, allowing for efficient signal mixing.

$$
\begin{aligned}
v_{IF_I} &= v_{in} \times v_{OSC_I} \\
&= \frac{A_1 A_2}{2} \left(\cos(\omega_{in}t - \omega_{OSC}t) + \cos(\omega_{in}t + \omega_{OSC}t)\right) \\
v_{IF_Q} &= v_{in} \times v_{OSC_Q} \\
&= \frac{A_1 A_2}{2} \left(\sin(\omega_{in}t + \omega_{OSC}t) - \sin(\omega_{in}t - \omega_{OSC}t)\right)
\end{aligned}
$$

- **$\mathbf{C_C}$**: AC coupling capacitor that isolates the DC level of the oscillator from the drain.
- **$\mathbf{R_{BIAS}}$**: Resistor that provides DC bias to the drain.

Purpose of R_{BIAS}

- Provides a DC path for biasing the drain with V_{BIAS}.
- Ensures that the drain of the NMOS has a proper quiescent point for correct operation.
- Blocks the AC signal from the oscillator from flowing into V_{BIAS}, since it forms a high impedance path at high frequencies.

MIXER CALCULATIONS

The threshold voltage V_{TH} was calculated to be 0.57 V from LTspice simulations.

The coupling capacitance C_c is used to block the DC component of the output from the quadrature oscillator. A value of 100 pF was chosen. The impedance due to this capacitance is given by:

$$X_{C_c} = \frac{1}{2\pi f C} \approx 15.9\,k\Omega$$

To ensure minimal signal loss through the bias resistor R_{bias} (so that most of the AC signal goes into the NMOS), we require:

$$R_{\text{bias}} \geq 10 \times X_{C_c}$$

Hence, we chose:

$$R_{\text{bias}} = 200\,k\Omega$$

Now, consider the signal relationships in the mixer circuit. The gate voltage is given by $V_g = V_{\text{BIAS}} + V_{\text{OSC}}$, and the source voltage is $V_s = V_{\text{in}}$. Therefore, the effective gate-source voltage relative to threshold is:

$$V_{gs} - V_{TH} = V_{\text{BIAS}} + V_{\text{OSC}} - V_{\text{in}} - V_{TH}$$

Assuming $V_{\text{BIAS}} = V_{TH}$, we simplify this to:

$$V_{gs} - V_{TH} = V_{\text{OSC}} - V_{\text{in}}$$

If $V_{\text{OSC}} - V_{\text{in}} < 0$, then $I_{DS} = 0$, and the MOSFET enters cutoff. If $V_{\text{OSC}} - V_{\text{in}} > 0$, the MOSFET operates in the triode region.

The drain-source voltage is given by:

$$V_{DS} = V_{\text{out}} - V_{\text{BIAS}} - V_{\text{OSC}}$$

For the MOSFET to operate in saturation, the condition $V_{DS} > V_{gs} - V_{TH}$ must be met, which is not satisfied in this case. Hence, the MOSFET remains in the triode region.

The drain current in the triode region is:

$$I_{DS} = \mu_n C_{OX} \frac{W}{L} \left[(V_{gs} - V_{TH}) V_{DS} \right]$$

Substituting into the expression for V_{out}:

$$V_{\text{out}} = I_{DS} R_L = k_n R_L \left[(V_{gs} - V_{TH}) V_{DS} \right]$$

Substituting $V_{gs} - V_{TH} = V_{\text{OSC}} - V_{\text{in}}$ and $V_{DS} = V_{\text{out}} - V_{\text{in}}$:

$$V_{\text{out}} = k_n R_L \left[(V_{\text{OSC}} - V_{\text{in}})(V_{\text{out}} - V_{\text{in}}) \right]$$

Assuming $V_{\text{OSC}} \gg V_{\text{in}} \gg V_{\text{out}}$, the output simplifies to:

$$V_{\text{out}} \approx k_n R_L V_{\text{OSC}} V_{\text{in}}$$

When $V_{\text{osc}} > 0$ irrespective of the V_{in} values $[+ve| - ve]$, $I_D = k \cdot R_i \cdot V_{\text{in}} V_{\text{osc}}$. So, we get negative values of the multiplied signal when $V_{\text{in}} < 0$ and $V_{\text{osc}} > 0$. When $V_{\text{osc}} < 0$, $I_D = 0$, so the output is indeed 0 since the MOSFET goes into cutoff region. But the information about the signal is not lost since the difference in frequencies of V_{in} and V_{osc} makes sure that the signal is slow/fast enough to capture the multiplication of all phases of the input signal! So, even if the MOSFET goes to cutoff region sometimes, the information of the input signal is not lost.

LT SPICE SIMULATION

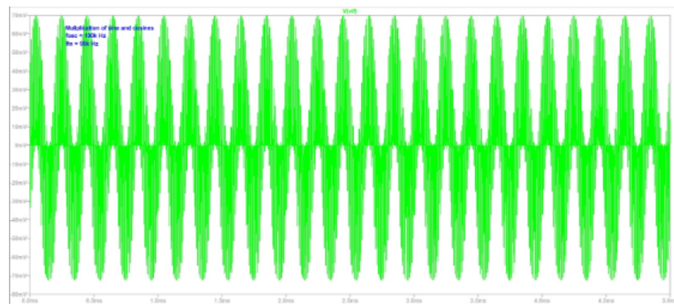

Fig. 13: Transient analysis for f_{in} = 95kHz

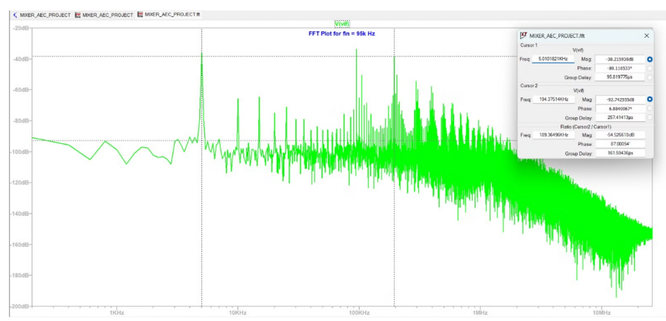

Fig. 14: FFT for f_{in} = 95kHz

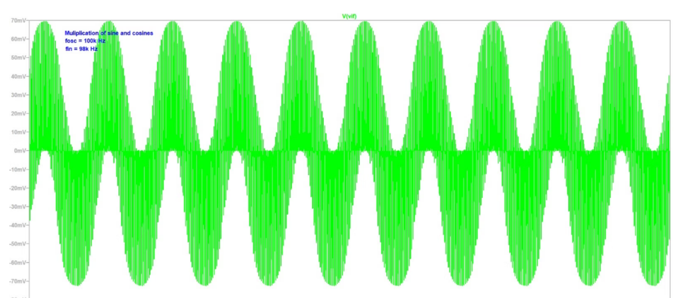

Fig. 15: Transient analysis for f_{in} = 98kHz

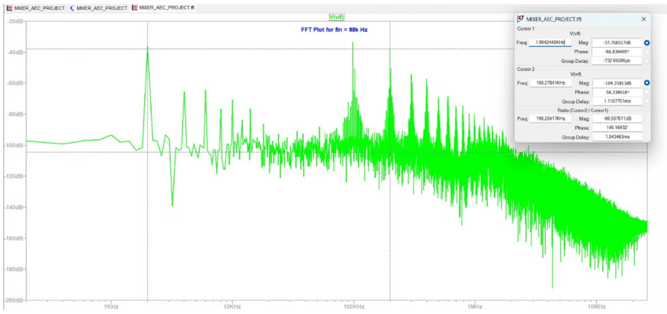

Fig. 16: FFT for f_{in} = 98kHz

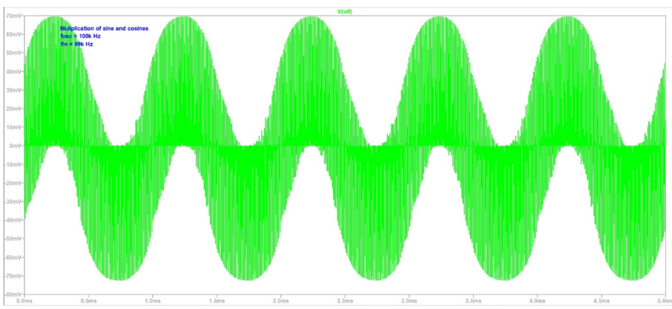

Fig. 17: Transient analysis for f_{in} = 99kHz

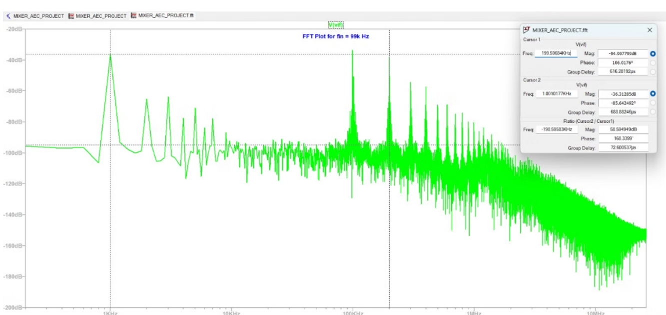

Fig. 18: FFT for f_{in} = 99kHz

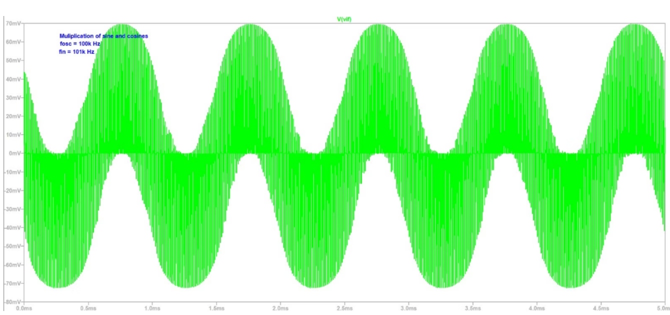

Fig. 19: Transient analysis for f_{in} = 101kHz

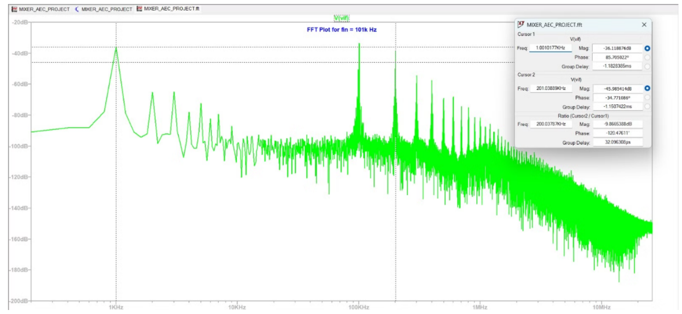

Fig. 20: FFT for f_{in} = 101kHz

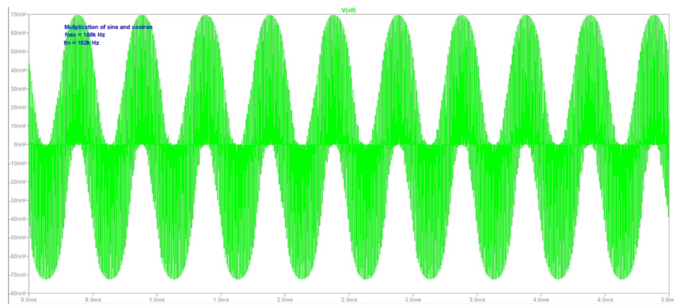

Fig. 21: Transient analysis for f_{in} = 102kHz

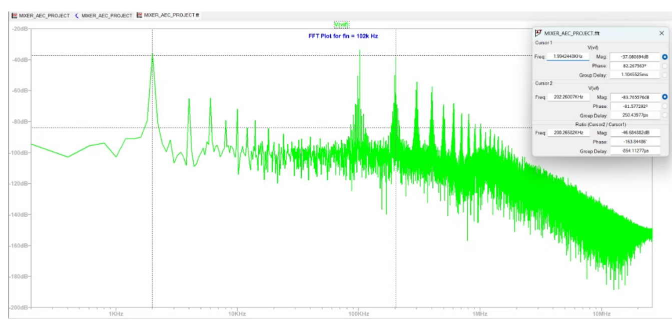

Fig. 22: FFT for f_{in} = 102kHz

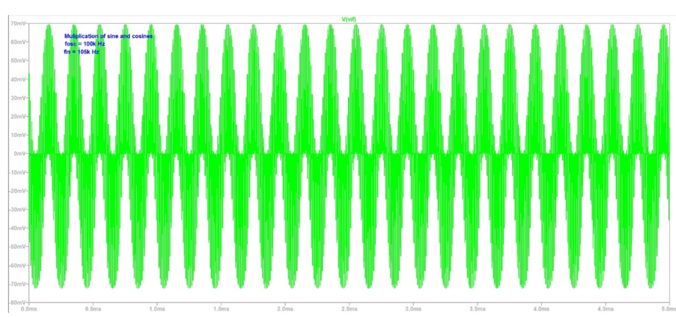

Fig. 23: Transient analysis for f_{in} = 105kHz

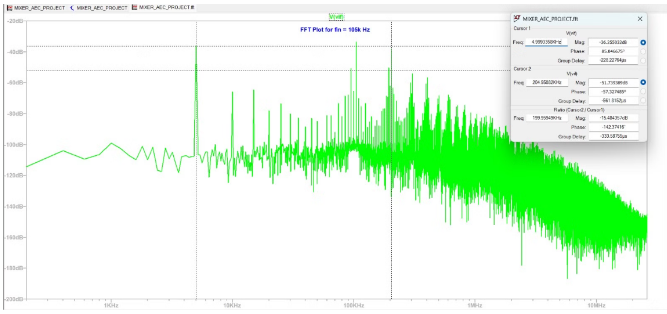

Fig. 24: FFT for f_{in} = 105kHz

V. Low Pass Filter

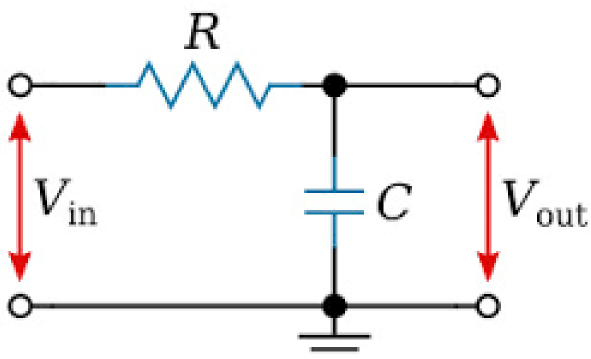

Fig. 25: Circuit diagram of Low Pass Filter

In the context of the project, the low pass filter (LPF) is assigned a specific bandwidth to allow only the required signal to pass through it. The LPF is placed after the mixing stage to eliminate high-frequency components, particularly the sum frequency products generated from the multiplication of the input signal with the local oscillator (LO). The cutoff frequency of the LPF defines the bandwidth of the down-converted signal. Since our project involves a quadrature down-converter, the LPF is used to extract the desired difference frequency component while attenuating the undesired higher frequency components. In other applications, the LPF can also be configured to isolate different frequency components based on specific requirements. Additionally, the LPF plays a crucial role in minimizing aliasing by limiting the frequency content of the signal before any sampling process. Aliasing can lead to errors in signal interpretation and reconstruction, particularly affecting the in-phase (I) and quadrature (Q) components in our design. Moreover, the LPF contributes to signal integrity by enhancing the signal-to-noise ratio, thus compensating for noise introduced from external sources. For our implementation, the required cutoff frequency is 2kHz, which is determined using the standard formula for an RC low pass filter:

$$\boxed{f_c = \frac{1}{2\pi RC}}$$

LT SPICE Simulations

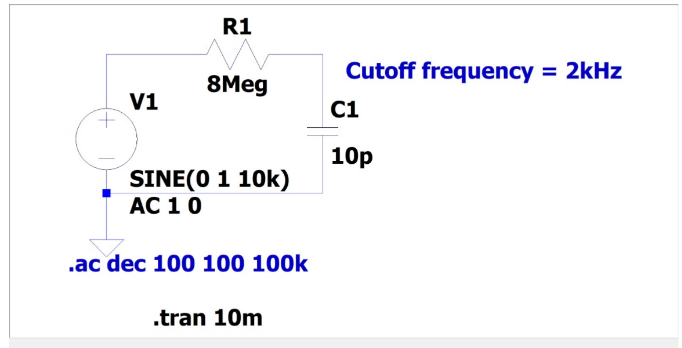

Fig. 26: Circuit diagram

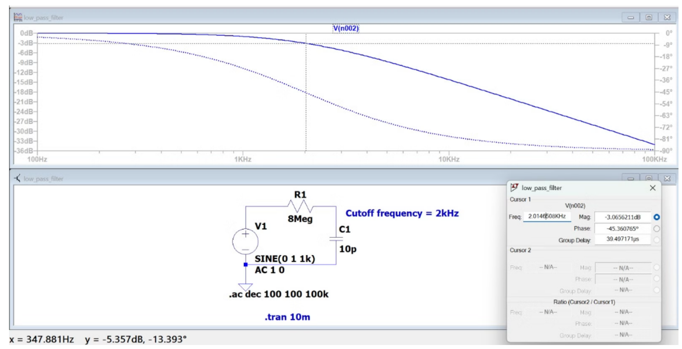

Fig. 27: AC Analysis of Low Pass Filter

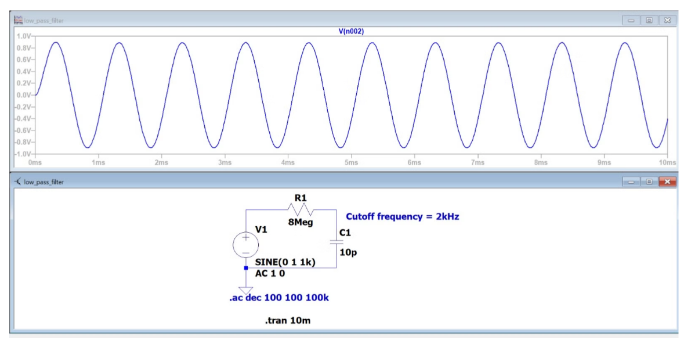

Fig. 28: Transient Analysis at 1kHz

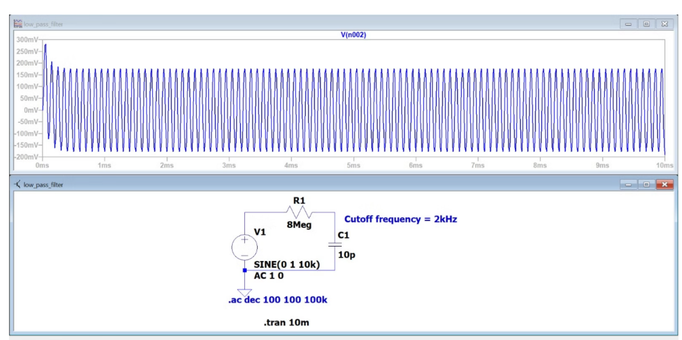

Fig. 29: Transient Analysis at 10kHz

VI. MIXER + LPF

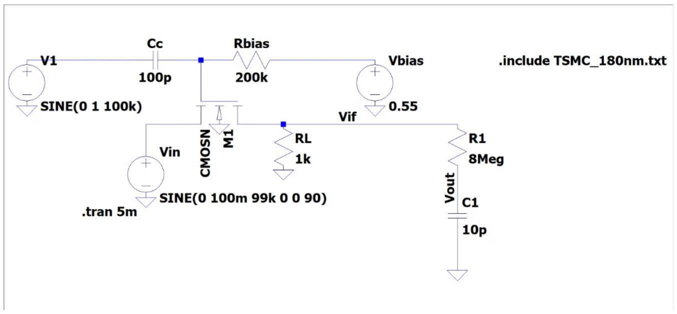

Fig. 30: Combined circuit diagram of Mixer and LPF

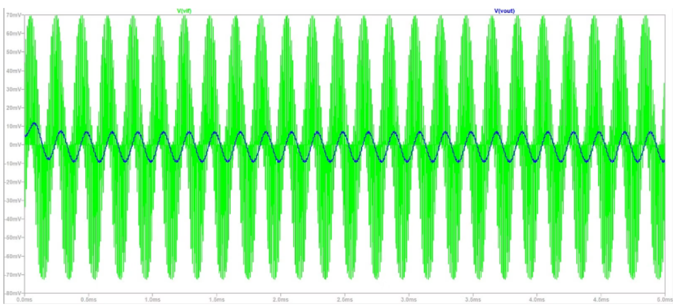

Fig. 31: Transient analysis for f_{in} = 95kHz

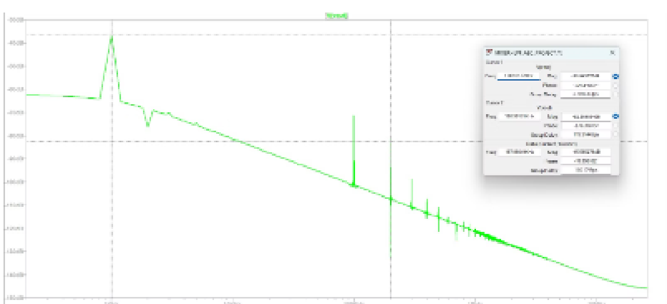

Fig. 32: FFT for f_{in} = 95kHz

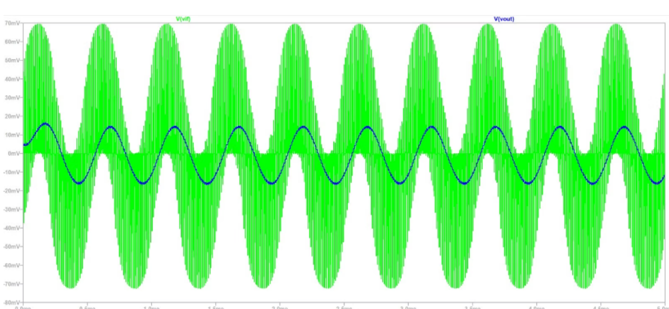

Fig. 33: Transient analysis for f_{in} = 98kHz

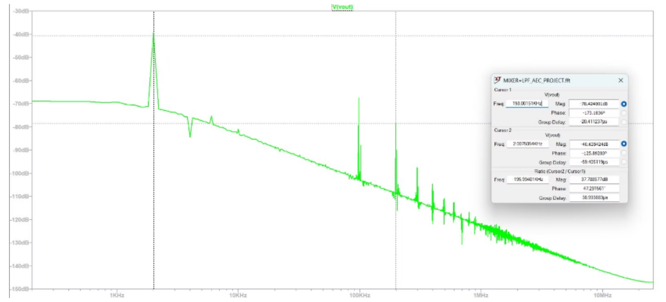

Fig. 34: FFT for f_{in} = 98kHz

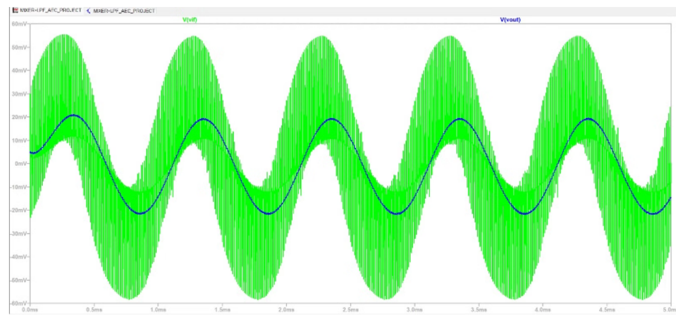

Fig. 35: Transient analysis for f_{in} = 99kHz

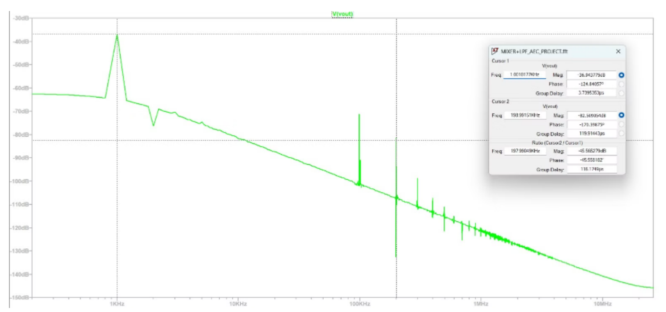

Fig. 36: FFT for f_{in} = 99kHz

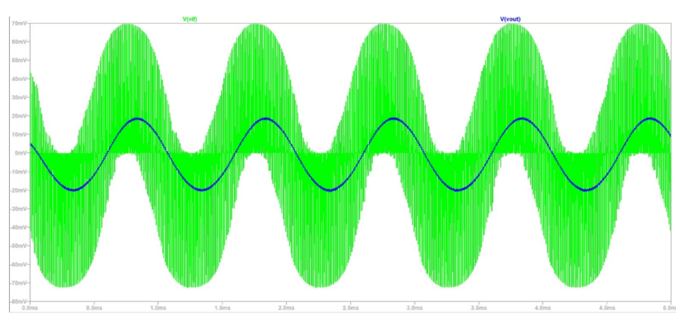

Fig. 37: Transient analysis for f_{in} = 101kHz

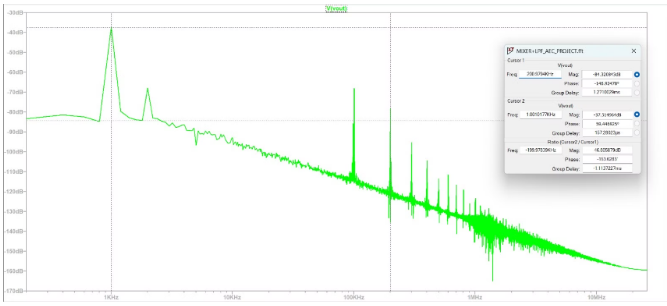

Fig. 38: FFT for f_{in} = 101kHz

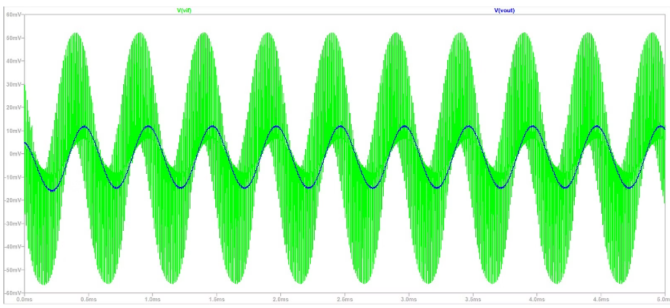

Fig. 39: Transient analysis for f_{in} = 102kHz

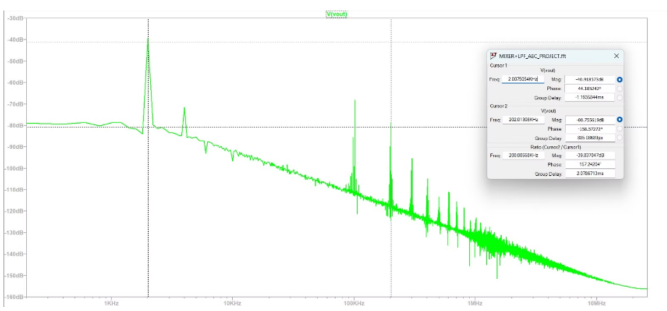

Fig. 40: FFT for f_{in} = 102kHz

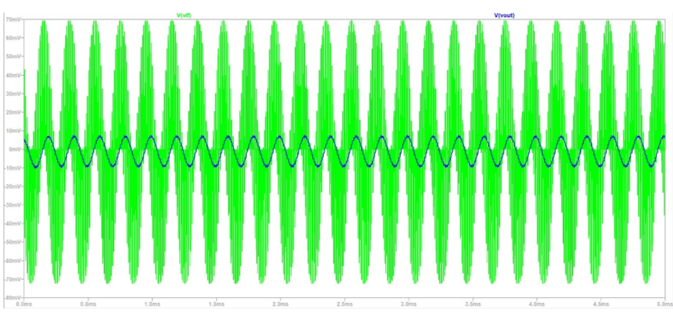

Fig. 41: Transient analysis for f_{in} = 105kHz

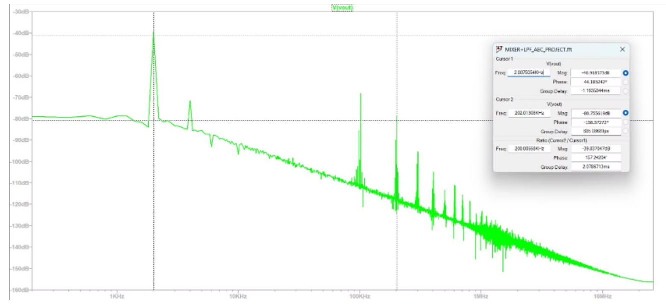

Fig. 42: FFT for f_{in} = 105kHz

VII. Overall Circuit

We take $R_3 = R_4 = 1.3\,\text{k}\Omega$ for the resistances of the inverting amplifier. After extensive testing, we found that the circuit only worked reliably when both resistors in this stage (input and feedback resistors) were set to $1.3\,\text{k}\Omega$(to achieve impedance matching). This specific choice was critical for achieving the correct output frequency and amplitude.

Using resistor values that match the oscillator's internal RC stages ensures impedance continuity throughout the circuit. Mismatched values (like $1\,\text{k}\Omega$ or $1.59\,\text{k}\Omega$) led to incorrect filtering at the output. At high frequencies, even slight mismatches in digestibility can affect the stability and accuracy of the output waveform.

For values other than 1.3kΩ for R_3 and R_4, a buffer could be added at both oscillator output terminals. This can be achieved using a non-inverting op-amp without resistances, but it would increase the circuit's cost and complexity. Therefore, we chose 1.3kΩ for simplicity.

Phase Analysis of I and Q Components

Yes, we can process the FFT plots to find the phase of final I and Q components. If we closely look at the FFT plots in the simulations below, it has both magnitude and the phase components (look at the cursors). The difference between the phases of both these cursors (i.e., 1st is I FFT and 2nd is Q FFT), using these we can get the required phase difference.

We can find the phase of the input signal by just finding the difference between the phase of the I_{FINAL} and Q_{FINAL} and subtract 90° from it, giving the input phase of the signal flag just finding the difference between the phase of the I_{FINAL} because of subbing the phase:

Let input = $A\cos(\omega_{\text{IN}}t + \phi)$.

$$\theta = (\omega_{\text{IN}} - \omega_{\text{OSC}})\,t + \phi$$

So, I_{FINAL} signal has the term $\cos(\theta)$ while the Q_{FINAL} signal has the $\sin(\theta)$ which is nothing but $\cos(90^\circ + \theta)$.

$$(\omega_{\text{IN}} - \omega_{\text{OSC}})\,t = 2\pi(k)$$

So, just finding the difference between the phases of both I_{FINAL} and Q_{FINAL} will give the required value of ϕ.

Thus, taking both In-phase and Quadrature-Phase components of will help us in preserving the phase of the input signal.

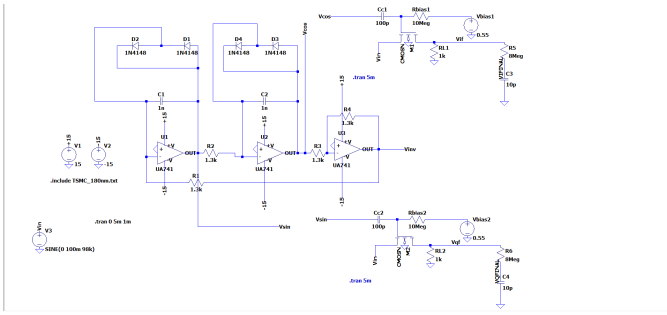

Fig. 43: Overall circuit diagram of Quadrature Down Converter

for f_{in} = 98kHz

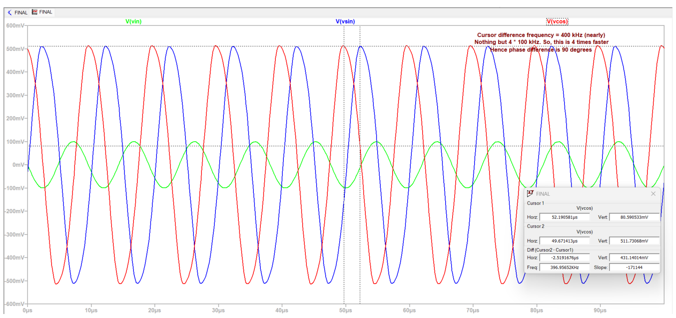

Fig. 44: Transient analysis of V_{in}, V_{OSC_I}, and V_{OSC_Q}

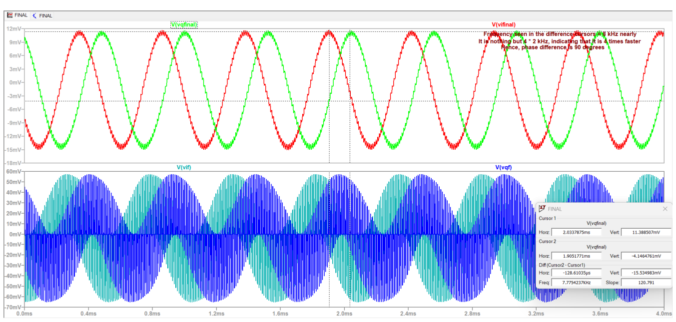

Fig. 45: Transient analysis of V_{IF_I}, V_{IF_Q}, $V_{IF_{FINAL_I}}$, and $V_{IF_{FINAL_Q}}$

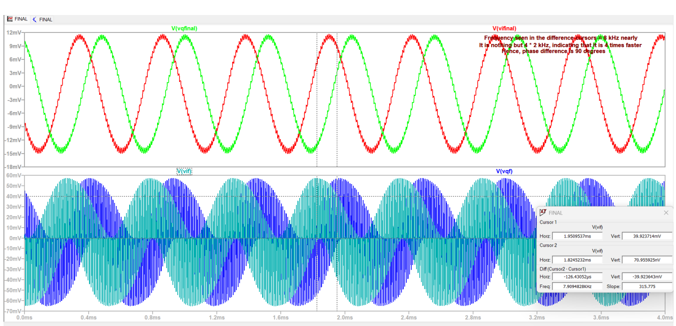

Fig. 46: Transient analysis of V_{IF_I}, V_{IF_Q}, $V_{IF_{FINAL_I}}$, and $V_{IF_{FINAL_Q}}$

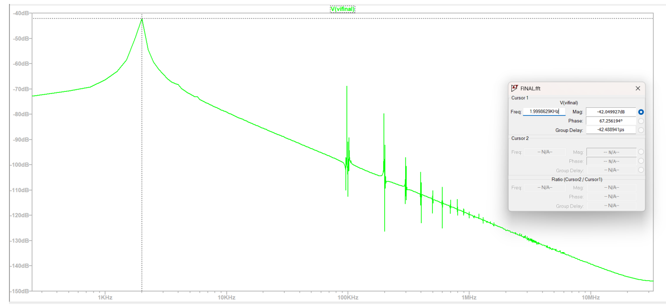

Fig. 47: FFT plot for $V_{IF_{FINAL_I}}$

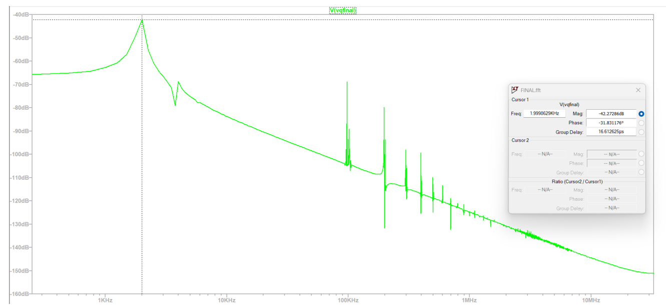

Fig. 48: FFT plot for $V_{IF_{FINAL_Q}}$

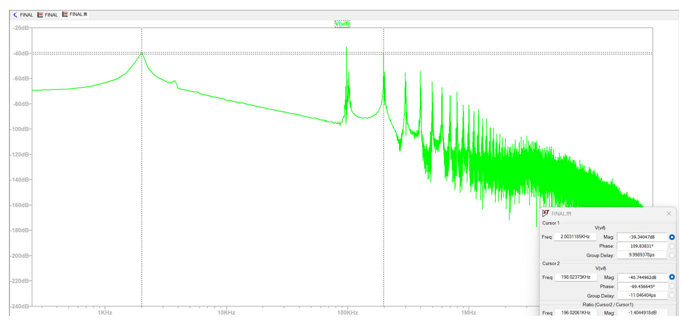

Fig. 49: FFT plot for V_{IF_I}

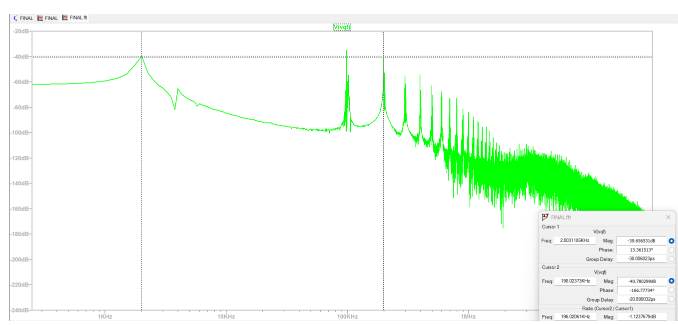

Fig. 50: FFT plot for V_{IF_Q}

for f_{in} = 105kHz

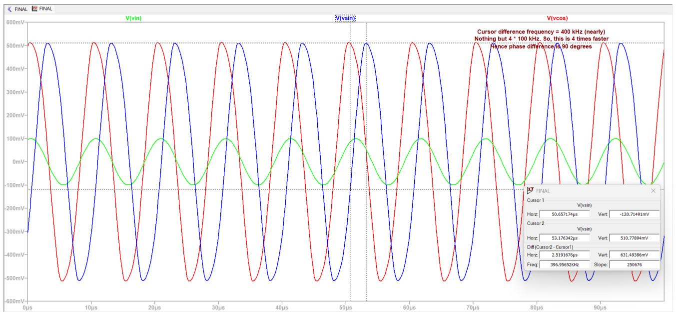

Fig. 51: Transient analysis of V_{in}, V_{OSC_I}, and V_{OSC_Q}

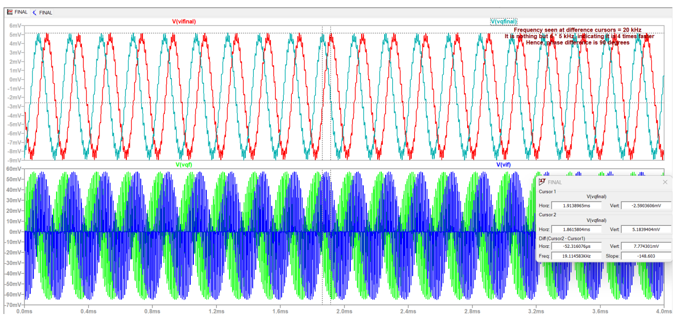

Fig. 52: Transient analysis of V_{IF_I}, V_{IF_Q}, $V_{IF_{FINAL_I}}$, and $V_{IF_{FINAL_Q}}$

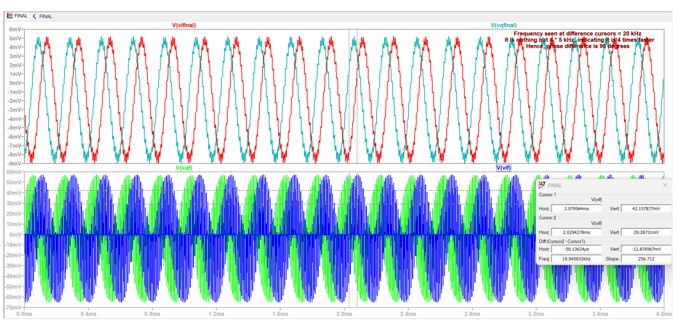

Fig. 53: Transient analysis of V_{IF_I}, V_{IF_Q}, $V_{IF_{FINAL_I}}$, and $V_{IF_{FINAL_Q}}$

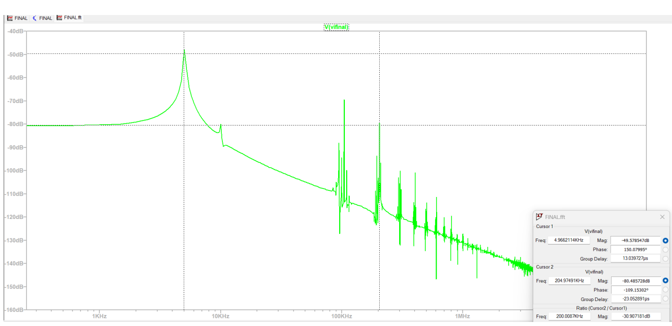

Fig. 54: FFT plot for $V_{IF_{FINAL_I}}$

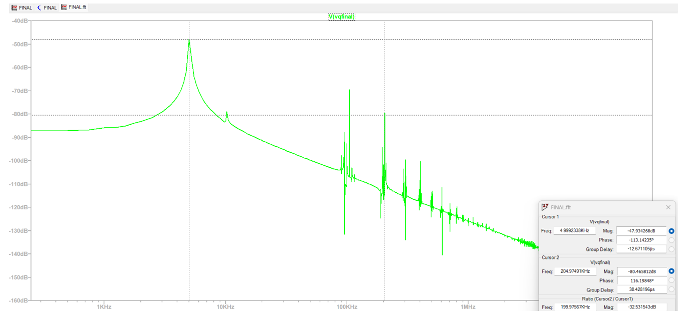

Fig. 55: FFT plot for $V_{IF_{FINAL_Q}}$

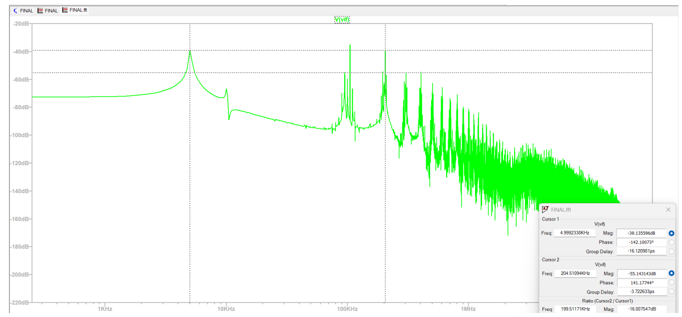

Fig. 56: FFT plot for V_{IF_I}

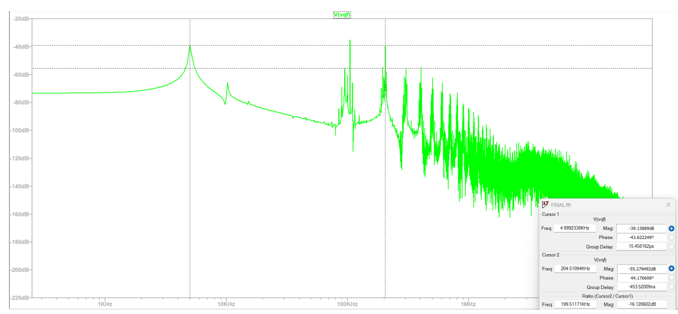

Fig. 57: FFT plot for V_{IF_Q}

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