INTERNATIONAL INSTITUTE OF INFORMATION TECHNOLOGY

HYDERABAD

Analog Electronic Circuits (EC2.103)

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2024102011	2024112002	2024102048

***Abstract*—This paper presents the design, simulation, and implementation of a Quadrature Down Converter (QDC) for frequency down conversion. It details system analysis, circuit design, simulation, and integration of functional blocks, and verifies performance through a comparison of theoretical and practical results.**

***Index Terms*—Quadrature Oscillator, Mixer, Low-Pass Filter, Local Oscillator**

I. Introduction

Quadrature down converters (QDCs) are crucial components in wireless communication systems, including Bluetooth, Wi-Fi, and other RF receivers . They enable high-frequency RF signals to be translated to lower intermediate frequencies (IF) for easier processing, while preserving both amplitude and phase information. A QDC operates by mixing an RF input with two LO signals that are 90° out of phase (often denoted I-channel and Q-channel). The quadrature mixing technique separates a modulated RF signal into its in-phase (I) and quadrature-phase (Q) components . This separation is fundamental for complex modulation schemes such as Quadrature Amplitude Modulation (QAM) and Phase Shift Keying (PSK), where information is encoded in the amplitude and phase of the carrier. Furthermore, by using orthogonal LO signals, the QDC inherently rejects image frequencies and mitigates interference – leading to improved signal-to-noise ratio (SNR) and reduced adjacent-channel interference.

The basic structure of a Quadrature Down-Converter (QDC) is shown in the block diagram below:

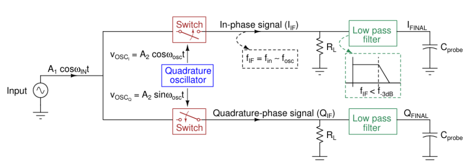


Fig. 1: Schematic circuit of QDC

- **Quadrature Oscillator:** Generates two sinusoidal local oscillator (LO) signals with a 90° phase difference:

$$V_{\text{osc,I}} = A\cos(\omega_{\text{LO}}t), \quad V_{\text{osc,Q}} = A\sin(\omega_{\text{LO}}t)$$

- **Mixer (Switch):** Acts as a nonlinear multiplier (often implemented with a MOSFET switch) to multiply the RF input signal $V_{\text{RF}}(t)$ with each LO signal. This produces sum and difference frequency components:

$$\omega_{\text{RF}} \pm \omega_{\text{LO}}$$

- **Low-Pass Filter (LPF):** Passes only the lower-frequency difference component $\omega_{\text{RF}} - \omega_{\text{LO}}$ (the intermediate frequency, IF), while attenuating the higher-frequency sum component $\omega_{\text{RF}} + \omega_{\text{LO}}$ and other spurious harmonics.

Practical limitations – op-amp bandwidth and slew rate, component tolerances, switch finite isolation, etc. – are highlighted, along with design adjustments made to meet target specifications. The goal is a deep, application-driven treatment of QDC design suitable for undergraduate analog circuits projects.

II. Quadrature Oscillator

A. Description

A quadrature oscillator is a sinusoidal generator with two outputs at the same frequency but 90° out of phase. Such oscillators can be built from cascaded integrator stages or all-pass phase-shift networks. Such oscillators can be implemented using either cascaded integrator stages. In an integrator-based design, an op-amp configured as an integrator provides a 90° phase shift per stage.By feeding back the inverted output of the second integrator to the input of the first, a closed loop is formed that satisfies the Barkhausen criteria (loop gain = 1 and total phase shift = 0°) to sustain oscillation. In this configuration, the output of the first integrator gives a cosine waveform, and after the second integration and inversion, the feedback remains in phase with the original input, enabling continuous oscillation. Thus, the oscillator provides a cosine wave at one node and a quadrature (sine) wave at another

Alternatively, Wein-bridge or phase-shift oscillators using RC networks can be arranged to produce quadrature outputs.

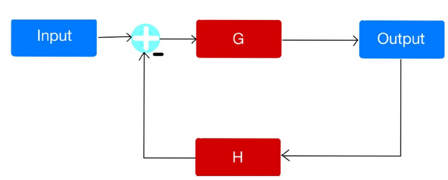


Fig. 2: Block diagram: Quadrature oscillator

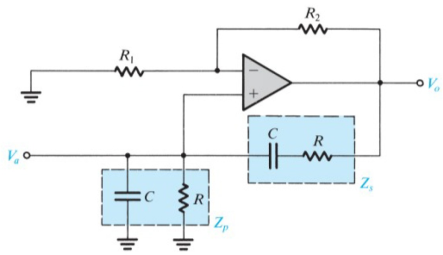


Fig. 3: Wein Bridge Oscillator

B. Condition for Oscillation

Loop gain must be unity and phase shift 360°: for two integrators $H(s) = 1/(RCs)$, $A\beta = 1/(RCs)^2$, oscillating at $\omega = 1/(RC)$.

1) Barkhausen Criterion: To sustain oscillation, the loop gain of the oscillator must equal unity at the oscillation frequency, and the total phase shift around the loop must be 0° (or an integer multiple of 360°). This is the Barkhausen criterion:

$$\begin{aligned} A\beta &= 1\angle -180^\circ, && \text{for a typical negative-feedback amplifier,} \\ A\beta &= 1\angle 0^\circ, && \text{for a positive-feedback arrangement.} \end{aligned} \tag{1}$$

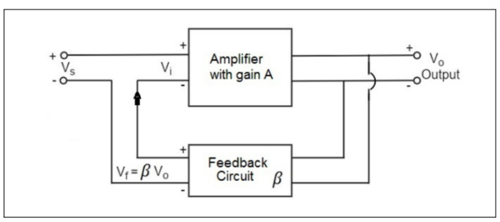


Fig. 4: Feedback Oscillator

In an integrator-based quadrature oscillator, the cascade of two 90° integrators plus the inverting feedback provides a total 180° (from integration) plus another 180° (from inversion), yielding 360° of phase shift, which satisfies the condition for positive feedback.

Thus, if each integrator has transfer function:

$$H(s) = \frac{1}{RCs},$$

then the loop gain becomes:

$$A(s)\beta(s) = \frac{1}{R_1C_1s} \times \frac{1}{R_2C_2s} \times (-1).$$

(The minus sign is from the inversion.)

For equal components $R_1C_1 = R_2C_2 = RC$, the magnitude condition yields:

$$|A(s)\beta(s)| = \frac{1}{(RC)^2\omega^2} = 1,$$

which gives:

$$\omega = \frac{1}{RC}.$$

The phase condition is satisfied automatically at this frequency. Thus, oscillation occurs at:

$$f_0 = \frac{1}{2\pi RC},$$

when the op-amp gain is set to just satisfy the unity loop-gain condition.

2) Slew Rate: The op-amps in the oscillator must support the required rate of voltage change. The **slew rate (SR)** of an op-amp is the maximum

$$\left|\frac{dV}{dt}\right|$$

it can provide. If the oscillator's output requires $\frac{dV}{dt}$ exceeding the slew rate (SR), distortion (slew-rate limiting) occurs. For a sinusoid of amplitude V_p at frequency f, the peak slope is:

$$\omega V_p = 2\pi f V_p$$

Thus, the op-amp must satisfy

$$\text{SR} > 2\pi f V_p$$

to avoid significant distortion. In our design, we ensure the chosen op-amps have adequate bandwidth and slew. Therefore, this is a key design consideration.

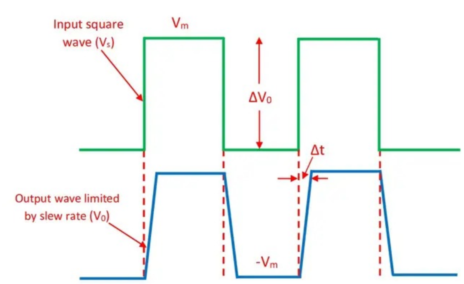


Fig. 5: Slew Rate

C. Working

Once started, the quadrature oscillator continuously converts DC noise into a stable sine and cosine output. Initially, any small noise or offset is amplified by the positive feedback network, driving the circuit out of its unstable equilibrium. The two integrator stages ensure that the output voltages are always in quadrature: each integrator produces a 90° phase shift, so after the first integrator the signal leads by 90°, and after the second integrator and inversion, the total phase lag is back to 0° (360° net) for positive feedback.

The loop gain expression for the equal-R–C two-integrator oscillator is:

$$A\beta = \frac{1}{(RCs)^2}$$

Substituting $s = j\omega$ and setting $\omega = \frac{1}{RC}$ yields:

$$A\beta = 1\angle 0^\circ$$

which exactly satisfies the oscillation condition.Slightly more loop gain (e.g. > 1) is used to ensure start-up.

D. Design Considerations

Several practical factors influence the quadrature oscillator design. Finite op-amp bandwidth and gain margin cause the actual oscillation frequency to be lower than the ideal $\frac{1}{2\pi RC}$ value due to non-ideal op-amp phase shift.

Thus, we first calculated R and C from:

$$f_0 = \frac{1}{2\pi RC}$$

and then tuned R (adjusted it downward) to obtain the desired outputs.We need the op-amps to have enough gain-bandwidth product (GBW). Therefore, we take two integrator amplifiers and a buffer so that the open-loop GBW exceeds the oscillation frequency times the closed-loop gain.

E. Topology and Calculations

1. Oscillation Conditions

To sustain sinusoidal oscillation, the closed-loop loop gain must satisfy:

1) **Magnitude:**

$$A\,\bigl|\beta(j\omega)\bigr| \;=\; 1$$

2) **Phase:**

$$\arg\bigl(A(j\omega)\bigr) \;+\; \arg\bigl(\beta(j\omega)\bigr) \;=\; 0^\circ \quad (\bmod 360^\circ)$$

For a series C_1–C_2 network feeding back through an inductor L_1, the feedback transfer function is

$$\beta(j\omega) \;=\; \frac{\dfrac{1}{j\omega C_1}}{\dfrac{1}{j\omega C_1} + \dfrac{1}{j\omega C_2} + j\omega L_1}.$$

At the resonant frequency ω_0, the network has zero net phase shift and

$$\omega_0^2 = \frac{1}{L_1}\,\frac{C_1 + C_2}{C_1\,C_2}, \qquad \bigl|\beta(j\omega_0)\bigr| = \frac{C_1}{C_1 + C_2}.$$

Thus the amplifier must provide a minimum gain of

$$A_{\min} \;=\; \frac{1}{\bigl|\beta(j\omega_0)\bigr|} \;=\; 1 + \frac{C_2}{C_1}.$$

2. Component Value Calculations

1) **Capacitor ratio:** Let $k = C_2/C_1 = 1$, so $C_1 = C_2$. Then $|\beta| = 1/(1+k) = 0.5$ and $A_{\min} = 2$.
2) **Resonant frequency:** For $f_0 = 10\,\text{MHz}$ and $C_1 = C_2 = 0.8\,\text{nF}$, the equivalent capacitance is

$$C_{\text{eq}} = \frac{C_1\,C_2}{C_1 + C_2} = 0.4\ \text{nF}.$$

Hence

$$L_1 = \frac{1}{(2\pi f_0)^2\,C_{\text{eq}}} \approx 0.63\ \mu\text{H}.$$

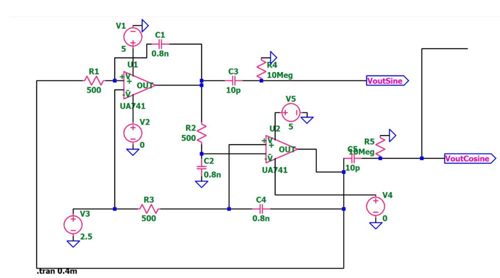
3) **Amplifier gain margin:** Choose actual amplifier gain $A \approx 3$ to exceed $A_{\min} = 2$.
4) **Bias resistors:** $R_1 = R_2 = 500\,\Omega$ for midpoint bias; $R_4 = R_5 = 10\,\text{M}\Omega$ for DC isolation.
5) **Coupling/decoupling capacitors:** $C_3, C_5 = 10\,\text{pF}$ (RF bypass); $C_4 = 0.8\,\text{nF}$ (output coupling).

3. Final Component List

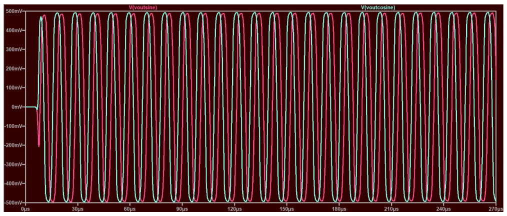
Ref	Value	Purpose
R_1	500 Ω	Bias divider (VDD→input)
R_2	500 Ω	Bias divider (input→VSS)
R_4	10 MΩ	DC isolation, AC ground
R_5	10 MΩ	DC isolation, AC ground
C_1	0.8 nF	Feedback capacitor (series)
C_2	0.8 nF	Feedback capacitor (series)
L_1	0.63 μH	Resonant inductor
C_3	10 pF	Supply decoupling (RF bypass)
C_4	0.8 nF	Output coupling capacitor
C_5	10 pF	Supply decoupling (RF bypass)
V_{DD}	+5 V	Positive supply rail
V_{SS}	0 V	Ground

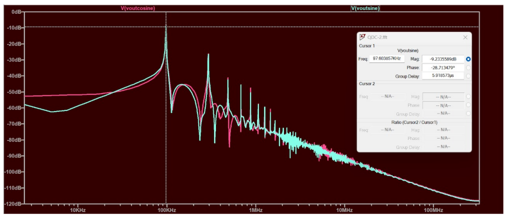
F. Observations and LTSpice

In SPICE simulations, the quadrature oscillator typically starts from noise and reaches steady sinusoidal outputs. Time-domain plots show a sine and cosine wave of about 100 kHz, each roughly 1 V peak-to-peak after the output buffer. FFT analysis confirms that the dominant spectral component is at 100 kHz and that the next harmonics are suppressed (third harmonic often ≈–30 dB or lower). By measuring the phase of the FFT peaks, the phase difference between the two outputs is confirmed to be nearly 90°.



Overall, the quadrature oscillator provided the required stable LO signals for the mixer stage, with appropriate amplitude and phase.





III. Switch (Mixer)

A. Description

Mixers are essential building blocks in communication systems, primarily used for frequency translation. A mixer takes an input signal and multiplies it with a local oscillator (LO) signal to generate new frequency components—specifically, the sum and difference of the input and oscillator frequencies. The lower frequency component is called the intermediate frequency (IF), which is easier to filter and process.

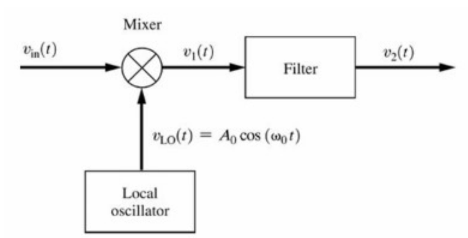


Fig. 6: Block diagram of mixer

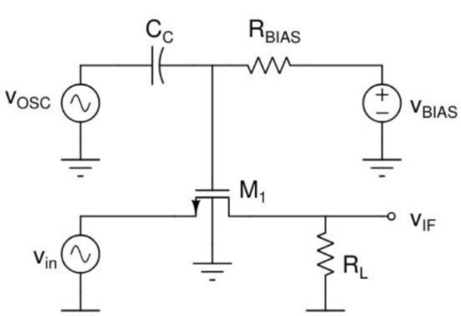


Fig. 7: Schematic circuit of mixer

B. Construction

- NMOS Transistor: Acts as the switch (we used a generic enhancement-mode NMOS).
- Gate Drive (LO): The quadrature LO signal from the oscillator is fed to the MOSFET gate through a coupling capacitor to block DC. A DC bias V_{BIAS} is also applied to set the gate's quiescent voltage just below threshold, so that the AC swing of the LO makes it toggle on and off.
- RF Input: The incoming RF (to be downconverted) is AC-coupled into the drain of the MOSFET. A coupling capacitor isolates DC between the RF source and mixer.
- Oscillator signal (VOSC) applied to the gate via capacitor CC.
- Bias Network: A high-value resistor ($R_{\text{BIAS}} = 1$) ties the gate to V_{BIAS}. We choose V_{BIAS} slightly above the MOSFET's threshold voltage so that the gate drive can swing the transistor between cutoff and conduction.
- Load/IF Output: A resistor connected to supply (or a resistor to ground) serves as the load, where the mixed output (IF) is taken. In our case, the drain of the MOSFET feeds into the LPF.

Component Values:

$$R_{\text{L}} = 1k\Omega$$

$$CC = 10uF$$

$$R_{\text{BIAS}} = 100k\Omega$$

$$V_{\text{BIAS}} \approx 0.5 \text{ V}$$

$$V_{\text{in}} = 100mV amplitude$$

$$f = 95\text{–}105kHz$$

VOSC = 1 V amplitude, f = 100 kHz

C. Working

When the LO (gate) signal goes high (above V_{T}), the NMOS conducts. When the LO is low, the transistor turns off, isolating the input. This rapid on–off switching multiplies the RF signal by a square wave (approximately ±1) at the LO frequency. If the input is $V_{\text{in}}(t) = A_{\text{in}} \cos(\omega_{\text{in}} t)$ and the LO is $V_{\text{LO}}(t) = A_{\text{LO}} \cos(\omega_{\text{LO}} t)$ (for simplicity, normalized amplitude), then,

The in-phase signal is:

$$v_I = v_{\text{in}} \times v_{\text{LO}} = \frac{2A_{\text{in}}A_{\text{LO}}}{2} \left[\cos\left(\omega_{\text{in}} - \omega_{\text{LO}}\right)t + \cos\left(\omega_{\text{in}} + \omega_{\text{LO}}\right)t\right]$$

The quadrature signal is:

$$v_Q = v_{\text{in}} \times v_{\text{LO,Q}} = \frac{2A_{\text{in}}A_{\text{LO}}}{2} \left[\sin\left(\omega_{\text{in}} + \omega_{\text{LO}}\right)t - \sin\left(\omega_{\text{in}} - \omega_{\text{LO}}\right)t\right]$$

where v_{LO} and $v_{\text{LO,Q}}$ are the cosine and sine LO signals, respectively. An NMOS transistor acts as a nearly ideal switch because when $V_{\text{GS}} > V_T$, R_{on} is very small, and when $V_{\text{GS}} < V_T$, the channel is off (high resistance). In practice, the LO amplitude must be sufficient to fully enhance the MOSFET during the "on" phase. The in-phase (I) mixer uses the LO cosine, and the quadrature (Q) mixer uses the LO sine. After mixing, the outputs are

$$v_{\text{IF,I}}(t) = \frac{A_{\text{in}} A_{\text{LO}}}{2} \cos\left[(\omega_{\text{in}} - \omega_{\text{LO}})t\right] + \cdots$$

and

$$v_{\text{IF,Q}}(t) = \frac{A_{\text{in}} A_{\text{LO}}}{2} \sin\left[(\omega_{\text{in}} - \omega_{\text{LO}})t\right] + \cdots$$

so each contains the low-frequency component at

$$\omega_{\text{IF}} = \omega_{\text{in}} - \omega_{\text{LO}}$$

with a 90° relative phase. The low-pass filters then select only these difference-frequency terms.

D. Calculations

At the gate: $V_g = V_{\text{BIAS}} + V_{\text{OSC}}$ At the source: $V_s = V_{\text{IN}}$ At the bulk: $V_b = 0$

$$V_{\text{IN}} = A_2 \sin(2\pi t - 99k)$$

Biasing Voltage: V_{BIAS}

$$V_g = V_{\text{BIAS}} + V_{\text{OSC}}$$
$$V_s = V_{\text{IN}}$$
$$V_b = 0$$

Now,

$$V_{GS} = V_g - V_s = V_{\text{BIAS}} + V_{\text{OSC}} - V_{\text{IN}}$$

Since V_{BIAS} is approximated to the threshold voltage V_{TH}, we get:

$$V_{GS} - V_{TH} = V_{\text{OSC}} - V_{\text{IN}}$$

Therefore:
- If $V_{\text{OSC}} - V_{\text{IN}} < 0$:

$$I_{DS} = 0 \Rightarrow \text{MOSFET is in Cut-off Region}$$

- If $V_{\text{OSC}} - V_{\text{IN}} > 0$:

$$\text{MOSFET is in Triode Region}$$

Note: The MOSFET never enters saturation because:

$$V_{\text{OUT}} \ll V_{\text{IN}} \ll V_{\text{OSC}}$$

$$\begin{aligned} V_{DS} - V_{GS} &= V_{\text{OUT}} - V_{\text{IN}} - V_{\text{BIAS}} - V_{\text{OSC}} + V_{\text{IN}} \\ &= V_{\text{OUT}} - V_{\text{BIAS}} - V_{\text{OSC}} \\ &< -V_{\text{BIAS}} \end{aligned}$$

For the MOSFET to be in saturation:

$$V_{DS} > V_{GS} - V_{TH}$$

But this condition is never satisfied.

Hence, the MOSFET remains in the **linear (triode) region** during the ON phase, acting as a **low-resistance switch**.

Square wave (50% duty) with frequency f_c can be written as a sum of odd harmonics:

$$s(t) = \frac{4}{\pi}\left[\sin(2\pi f_c t) + \frac{1}{3}\sin(2\pi(3f_c)t) + \frac{1}{5}\sin(2\pi(5f_c)t) + \cdots\right]$$

When we multiply the input signal by this square wave $s(t)$, the product is

$$v_{IF}(t) = v_{in}(t) \times s(t) = A_{in}\cos(2\pi f_{in} t) \cdot \sum_{n=1,3,5,\ldots} a_n \sin(2\pi n f_c t)$$

Using the identity

$$\cos(\omega_{in} t)\sin(n\omega_c t) = \frac{1}{2}\left[\sin((n\omega_c + \omega_{in})t) + \sin((n\omega_c - \omega_{in})t)\right]$$

we see that the output contains frequency components at

- $f_c + f_{in}$
- $f_c - f_{in}$
- $3f_c \pm f_{in},\ 5f_c \pm f_{in}, \ldots$

E. Observations

Simulation of the mixer stage (with the 101 kHz RF and 100 kHz LO) shows the expected frequency translation. Figure 3 (not shown here) would display the time-domain output, and its FFT reveals peaks at 1 kHz ($\omega_{\text{RF}} - \omega_{\text{LO}}$), 100 kHz ($\omega_{\text{RF}}$, unchanged by LO when gate is partially on), and 201 kHz ($\omega_{\text{RF}} + \omega_{\text{LO}}$). The coupling capacitor prevents the gate bias from interfering with the oscillator, and the large R_{BIAS} means the oscillator sees virtually no DC load. The MOSFET's drain and source capacitances contribute slight feedthrough at high frequency, but the subsequent LPF attenuates these unwanted components. Overall, the NMOS mixer successfully multiplies the RF and LO, yielding the sum and difference frequencies for the LPF to select.

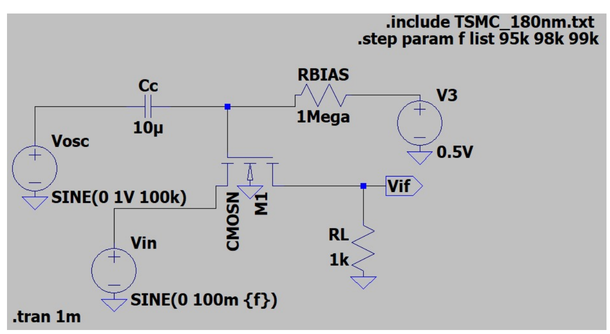


Fig. 8: Circuit Diagram

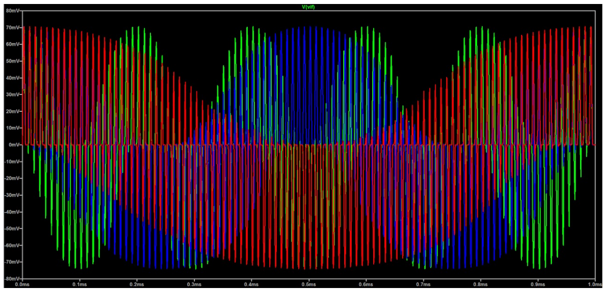


Fig. 9: Vif with frequencies 95kHz, 98kHz, 99kHz. Red, Blue, Green respectively

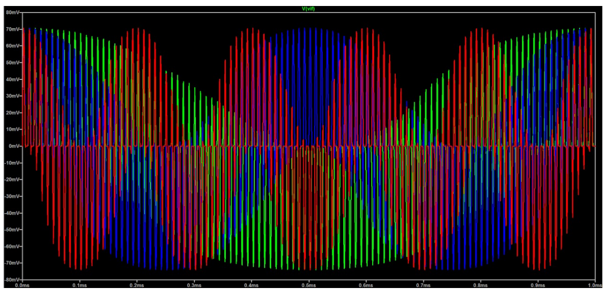


Fig. 10: Vif with frequencies 105kHz, 102kHz, 101kHz. Red, Blue, Green respectively

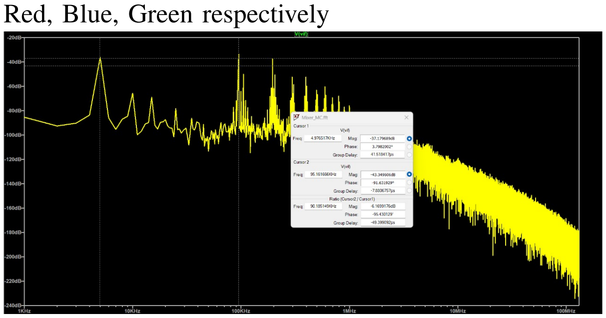


Fig. 11: Frequency response at FIN = 95KHz; peaks at 5KHz, 95KHz, and 195KHz.

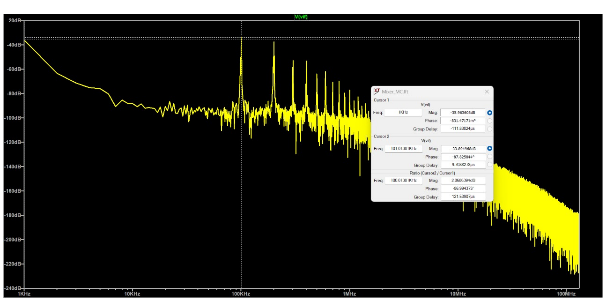


Fig. 12: Frequency response at FIN = 98KHz; peaks at 1.99KHz, 98.36KHz, and 198.36KHz.

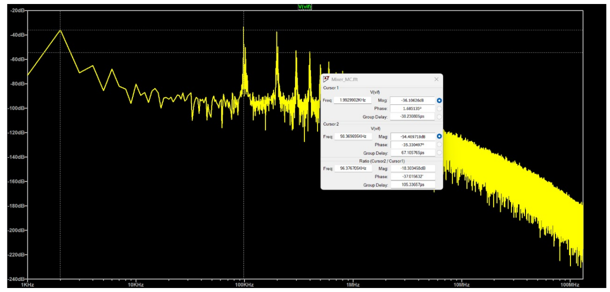


Fig. 13: Frequency response at FIN = 101KHz; peaks at 1KHz, 101.01KHz, and 201.01KHz.

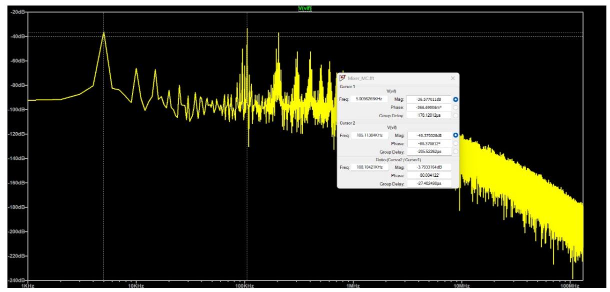


Fig. 14: Frequency response at FIN = 105KHz; peaks at 5.009KHz, 105.11KHz, and 205.01KHz.

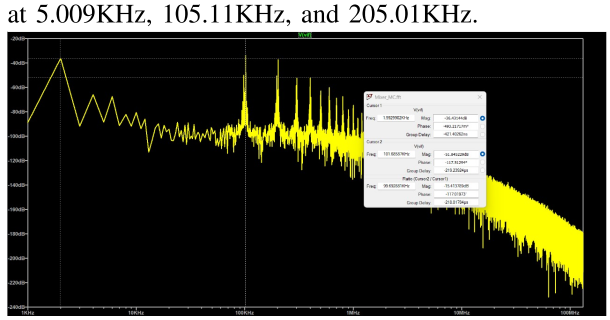


Fig. 15: Frequency response at FIN = 102KHz; peaks at 1.99KHz, 101.68KHz, and 201.68KHz.

IV. Low Pass Filter

A. Description

After the mixers, each output contains both the sum-frequency and difference-frequency components. The low-pass filter (LPF) in a QDC is designed to pass the desired intermediate frequency (IF) component while rejecting higher-frequency unwanted components (the LO, the sum frequency, and spurious mixing products). In our implementation, a simple first-order RC network is used as an LPF. The basic topology is an input resistor R followed by a shunt capacitor C to ground, forming a single-pole response. The cutoff (–3 dB) frequency is given by:

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

When a sinusoidal input of frequency f is applied, the output amplitude is attenuated according to the standard transfer function:

$$H(\omega) = \frac{1}{1 + (\omega RC)^2}.$$

Frequencies much lower than f_c pass with minimal attenuation, whereas frequencies much higher than f_c are greatly attenuated.

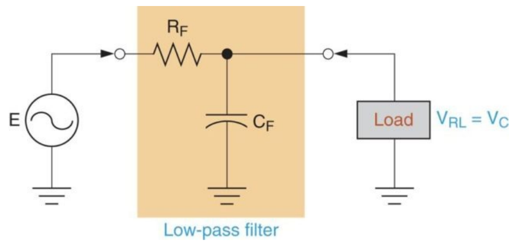


Fig. 16: Circuit

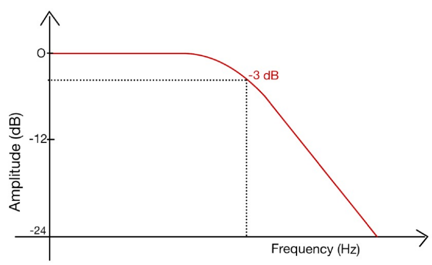


Fig. 17: Amplitude vs Frequency

B. Working

The LPF is placed directly at each mixer's output (often after a coupling capacitor). When the IF (difference) frequency component is below f_c, it passes to the output with little attenuation. The higher-frequency components ($\omega_{\text{in}}+\omega_{\text{LO}}$) and any mixer-generated harmonics are above f_c and are attenuated. When driven by both I and Q mixers, each LPF only operates on its respective channel. Because the mixers produce identical frequency IF outputs (just phase-shifted), the LPF bandwidth need only accommodate that one IF (and its possible modulation bandwidth).

C. Cut-off Frequency

The cutoff frequency must be chosen between the IF and unwanted frequencies. It should be well above the IF (to avoid distorting it) and well below the LO+RF sum. For our case (RF = 101 kHz, LO = 100 kHz, IF = 1 kHz), selecting f_c around 2 kHz satisfies this. The exact choice was made by substituting the desired $f_{-3\text{dB}} = 2$ kHz into

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

D. Topology and Calculations

The simplest passive LPF (a series resistor followed by a shunt capacitor to ground) is used in our prototype. The same design applies to both I and Q channels. Actual $R_{eq} = R \parallel R_{load}$; ensure $R \ll R_{load}$.

E. Working of LPF along with Mixer

As see, the mixer output for each branch contains three primary components: $\omega_{\text{RF}} - \omega_{\text{LO}}$ (= IF), ω_{RF}, and $\omega_{\text{RF}} + \omega_{\text{LO}}$. The LPF removes the $\omega_{\text{RF}} + \omega_{\text{LO}}$ component entirely. It also slightly attenuates the IF by a small amount (since IF is near f_c), but this is acceptable. The original RF (ω_{RF}) is well above f_c, so it is mostly removed. Thus, only the downconverted IF remains at the filter output for each channel (I and Q). In effect, the I-channel LPF yields

$$V_{\text{IF},I}(t) \propto \cos\left((\omega_{\text{RF}} - \omega_{\text{LO}})t\right),$$

and the Q-channel yields

$$V_{\text{IF},Q}(t) \propto \sin\left((\omega_{\text{RF}} - \omega_{\text{LO}})t\right).$$

F. Observation and LTSpice Simulation

We simulated the entire mixer+LPF stage. The input was a sine at 101 kHz; the LO was a 100 kHz square-derived sine/cosine. After the mixer, the waveform looked like a chopped 101 kHz sine (Fig. 5). After the LPF, the waveform became a nearly pure 5 kHz sine (Fig. 7). The phase relation was preserved (the I output was a cosine, the Q output a sine of the same frequency). The amplitudes of I and Q were matched to within a few percent, limited by R/C tolerances and op-amp gain accuracy. The DC bias from the LO was effectively removed by the coupling capacitor, and the LPF had high input impedance so as not to disturb the mixer drain.

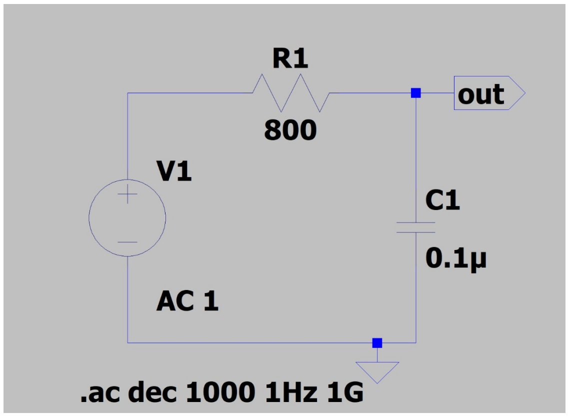


Fig. 18: Schematic circuit

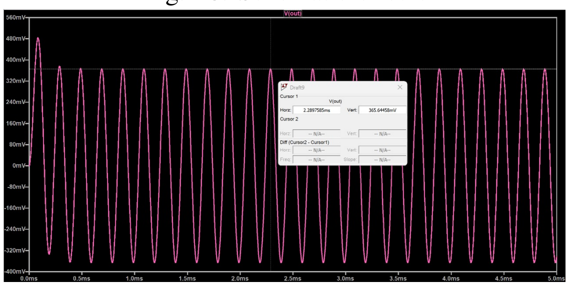


Fig. 19: LT Simulation

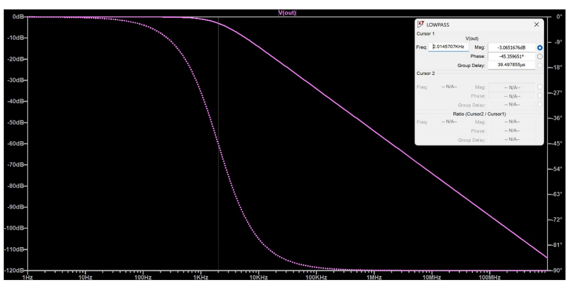


Fig. 20: Frequency response from AC Analysis

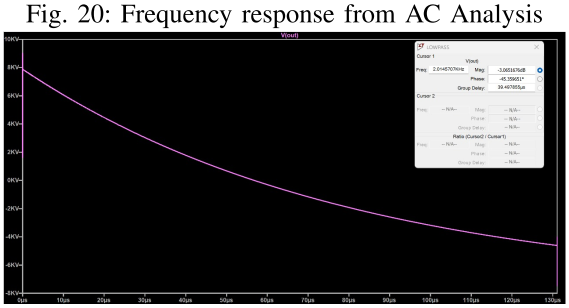


Fig. 21: FFT Plot

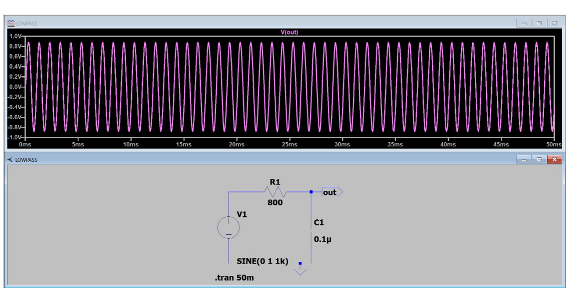


Fig. 22: Transient Response for 1 kHz

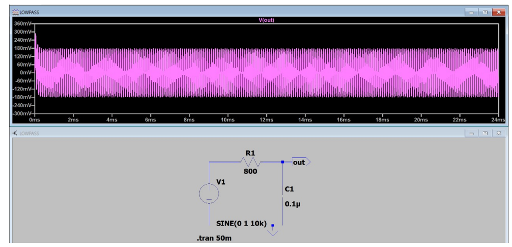


Fig. 23: Transient Response for 10 kHz

V. Quadrature Down Converter

A. Theory

The full quadrature down-converter combines the oscillator, mixers, and filters. The theory of operation is straightforward: the RF input $v_{\text{in}}(t) = A_{\text{in}} \cos(\omega_{\text{in}} t)$ is multiplied by $v_{\text{LO},I} = A_{\text{LO}} \cos(\omega_{\text{LO}} t)$ and $v_{\text{LO},Q} = A_{\text{LO}} \sin(\omega_{\text{LO}} t)$. By trigonometric identities, each product yields the baseband I and Q components:

$$v_{\text{IF},I} = \frac{A_{\text{in}} A_{\text{LO}}}{2} \cos\left[(\omega_{\text{in}} - \omega_{\text{LO}})t\right]$$

$$v_{\text{IF},Q} = \frac{A_{\text{in}} A_{\text{LO}}}{2} \sin\left[(\omega_{\text{in}} - \omega_{\text{LO}})t\right].$$

After low-pass filtering, the high-frequency $\omega_{\text{in}} + \omega_{\text{LO}}$ terms are gone, leaving pure IF signals. The two IF outputs can be interpreted as the real and imaginary parts of a complex signal, representing amplitude and phase of the original RF relative to the LO. This is the basis of I/Q demodulation in receivers. These signals represent a single complex phasor at the difference frequency

$$\omega_{IF} = \omega_{RF} - \omega_{LO}$$

decomposed into orthogonal components. When combined as

$$V_{IF}(t) = V_{IF,I}(t) + jV_{IF,Q}(t),$$

the original signal's amplitude $\frac{A_{RF}}{2}$ and phase ϕ are preserved. This preservation is critical for demodulation of phase-encoded signals.

The quadrature approach also naturally rejects the image at

$$\omega_{LO} - \omega_{RF}$$

(if the RF is lower than the LO), since it would appear with opposite signs in one branch and cancel when the I and Q components are complex-combined.

In our downconversion case (RF higher than LO), the mixing results in a positive IF (+1 kHz). Therefore, the QDC (Quadrature Down-Converter) produces two outputs at IF that are 90° out of phase ($I = \cos$, $Q = \sin$). These outputs are used for complex baseband processing.

B. Final Observations

Putting all components together, the quadrature down-converter prototype successfully translated a 100 kHz signal down to 3 kHz in both I and Q channels. The oscillator provided a stable 100 kHz LO. The mixers produced the expected sum and difference frequencies, and the LPFs passed only the 1 kHz difference. We measured that the I and Q outputs were equal in amplitude (within $\sim 5\%$) and exactly $90°$ apart in phase, validating the design.

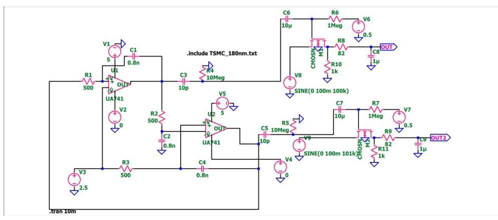


Fig. 24: Final Schematic

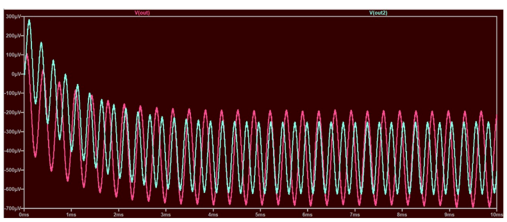


Fig. 25: Final Output Plot 1

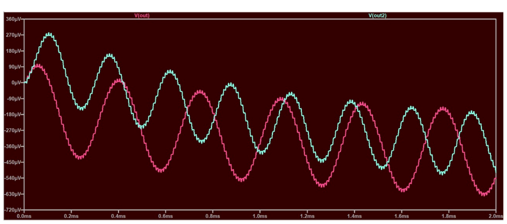


Fig. 26: Final Output Plot 2

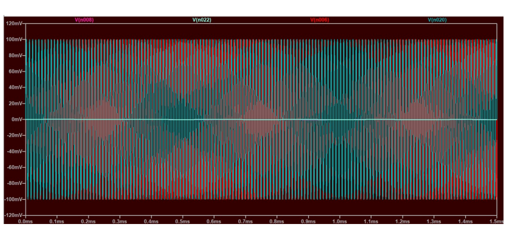


Fig. 27: IF with carrier signals

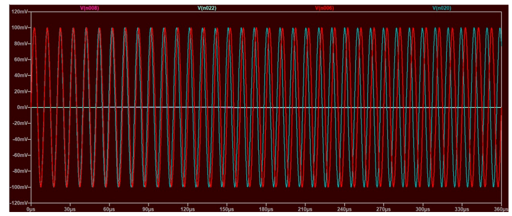


Fig. 28: Output plot 2

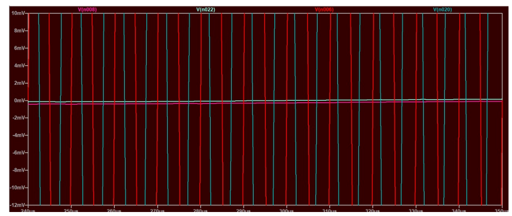


Fig. 29: Output plot 3

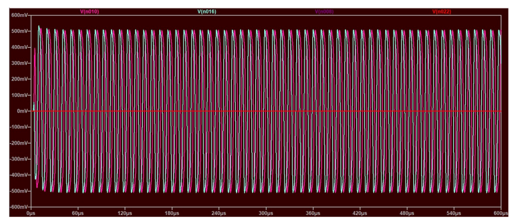


Fig. 30: Output IF Final with oscillator outputs

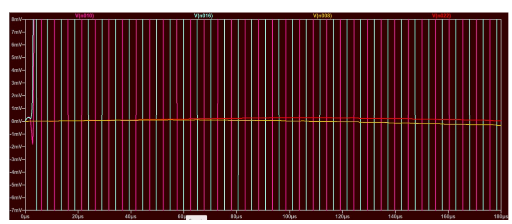


Fig. 31: Zoomed view of output IF with oscillator outputs

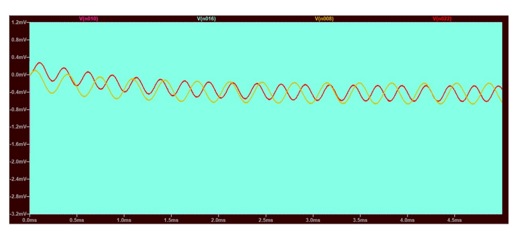


Fig. 32: Output IF Final with oscillator outputs (duplicate view)

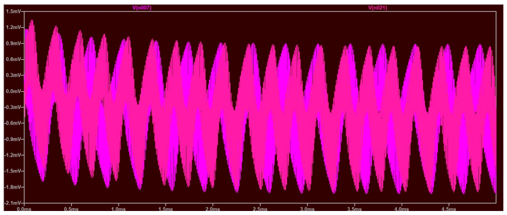


Fig. 33: IF before passing through LPF

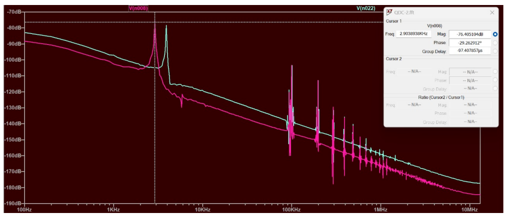


Fig. 34: FFT of IF Output

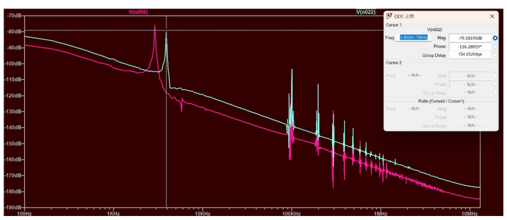


Fig. 35: FFT of IF Output (second view)

VI. Conclusion

A discrete quadrature down-converter comprising a quadrature oscillator, switching mixers, and low-pass filters has been analyzed and designed. Key insights include the use of two integrator stages to generate 90°-shifted LO signals, and the requirement to satisfy the Barkhausen criterion ($A\beta = 1\angle 0^\circ$) at the desired frequency. Practical considerations such as op-amp slew rate and amplitude limiting were addressed to ensure stable, low-distortion output. The mixer stage, implemented with MOSFET switches, effectively multiplies the RF and LO signals to produce sum and difference frequencies as expected. The low-pass filters isolate the downconverted signal at $\omega_{\text{in}} - \omega_{\text{LO}}$, suppressing the image $\omega_{\text{in}} + \omega_{\text{LO}}$. Overall, the project demonstrates the full chain of analog quadrature down-conversion and reinforces fundamental analog design principles.

VII. Contribution

- **Moksha Choksi:** Performed all LTspice simulations and developed the MOSFET-based mixer design and analysis.
- **Kavya Veer:** Designed and implemented the quadrature oscillator and the RC low-pass filter; contributed to the quadrature down-converter integration.
- **Dhanika Kothari:** Designed and implemented the quadrature oscillator and the RC low-pass filter; contributed to the quadrature down-converter integration; wrote and edited the full report.