Lecture 7: Voltage-Controlled Oscillators
Announcements

- HW3 is due Oct 25
Agenda

- VCO Fundamentals
- VCO Examples
- VCO Phase Noise
  - Phase Noise Definition and Impact
  - Ideal Oscillator Phase Noise
  - Leeson Model
  - Hajimiri Model
  - Phasor-Based Phase Noise Analysis
- VCO Jitter
Charge-Pump PLL Circuits

- Phase Detector
- Charge-Pump
- Loop Filter
- VCO
- Divider
Voltage-Controlled Oscillator

- Time-domain phase relationship

\[ \phi_{out}(t) = \int \Delta \omega_{out}(t) dt = K_{VCO} \int v_c(t) dt \]

\[ \omega_{out}(t) = \omega_0 + \Delta \omega_{out}(t) = \omega_0 + K_{VCO} v_c(t) \]
Voltage-Controlled Oscillators (VCO)

- Ring Oscillator
  - Easy to integrate
  - Wide tuning range (5x)
  - Higher phase noise

- LC Oscillator
  - Large area
  - Narrow tuning range (20-30%)
  - Lower phase noise
Barkhausen’s Oscillation Criteria

- Sustained oscillation occurs if
  \[ H(j\omega) = 1 \]

- 2 conditions:
  - Gain = 1 at oscillation frequency \( \omega_0 \)
  - Total phase shift around loop is \( n\times360^\circ \) at oscillation frequency \( \omega_0 \)
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Ring Oscillator Example

\[ H(s) = -\frac{A_0^3}{\left(1 + \frac{s}{\omega_0}\right)^3} \]

\[ \omega_{osc} = \sqrt{3}\omega_0 \]

\[ \tan^{-1} \frac{\omega_{osc}}{\omega_0} = 60^\circ \]

\[ \frac{V_{out}(s)}{V_{in}(s)} = \frac{-A_0^3}{\left(1 + \frac{s}{\omega_0}\right)^3} = \frac{-A_0^3}{\left(1 + \frac{s}{\omega_0}\right)^3 + A_0^3} \]

\[ \left[\sqrt{1 + \left(\frac{\omega_{osc}}{\omega_0}\right)^2}\right]^3 = 1 \]

\[ A_0 = 2 \]
Ring Oscillator Example

- 4-stage oscillator – work this one out yourself
  - $A_0 = \sqrt{2}$
  - Phase shift = 45°

- Easier to make a larger-stage oscillator oscillate, as it requires less gain and phase shift per stage, but it will oscillate at a lower frequency
• Oscillation phase shift condition satisfied at the frequency when the LC (and R) tank load displays a purely real impedance, i.e. 0° phase shift

\[
Z_{eq}(s) = \frac{R_s + L_1 s}{1 + L_1 C_1 s^2 + R_s C_1 s}
\]

\[
|Z_{eq}(s = j\omega)|^2 = \frac{R_s^2 + L_1^2 \omega^2}{(1 - L_1 C_1 \omega^2)^2 + R_s^2 C_1^2 \omega^2}
\]
LC Oscillator Example

- Transforming the series loss resistor of the inductor to an equivalent parallel resistance

\[ L_p = L_1 \left(1 + \frac{R_s^2}{L_1 \omega^2}\right), \quad C_p = C_1, \quad R_p \approx \frac{L_1^2 \omega^2}{R_s} \]

\[ \omega_1 = \frac{1}{\sqrt{L_p C_p}} \]

[Razavi]
LC Oscillator Example

- Phase condition satisfied at

- Gain condition satisfied when \((g_m R_P)^2 \geq 1\)

- Can also view this circuit as a parallel combination of a tank with loss resistance 2\(R_P\) and negative resistance of \(2/g_m\)

- Oscillation is satisfied when

\[
\frac{1}{g_m} \leq R_P
\]
• Noise in the system will initiate oscillation, with the signals eventually exhibiting rail-to-rail swings
• While the small-signal transistor parameters ($g_m$, $g_o$, $C_g$, etc...) can be used to predict the initial oscillations during small-signal start-up, these parameters can vary dramatically during large-signal operation
For this large-signal oscillator, the frequency is set by the stage delay, $T_D$

$T_D$ is a function of the nonlinear current drive and capacitances of each stage

As an “edge” has to propagate twice around the loop

$$f_{osc} = \frac{1}{6T_D}, \text{ or } \frac{1}{2NT_D}$$

where $N$ is the oscillator stage number
Supply-Tuned Ring Oscillator

\[ T_{VCO} = 2nT_D \approx \frac{2nC_{stage}}{\beta(V_c - V_{th})} \]

\[ K_{VCO} = \frac{\partial f_{VCO}}{\partial V_c} = \frac{\beta}{2nC_{stage}} \]
Current-Starved Ring Oscillator

[Sanchez]

Current - starved VCO.
Capacitive-Tuned Ring Oscillator

\[ C_{eff} = \frac{C}{1 + sCR} \]
Symmetric Load Ring Oscillator

- Symmetric load provides frequency tuning at excellent supply noise rejection
- See Maneatis papers for self-biased techniques to obtain constant damping factor and loop bandwidth (% of ref clk)
LC Oscillator

- A variable capacitor (varactor) is often used to adjust oscillation frequency.

- Total capacitance includes both tuning capacitance and fixed capacitances which reduce the tuning range.

\[
\omega_{osc} = \frac{1}{\sqrt{L_p C_p}} = \frac{1}{\sqrt{L_p (C_{tune} + C_{fixed})}}
\]
Varactors

- **pn junction varactor**
  - Avoid forward bias region to prevent oscillator nonlinearity

- **MOS varactor**
  - Accumulation-mode devices have better Q than inversion-mode

[Perrott]

[Razavi]
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Oscillator Noise

\[ V(t) \]

**Jitter**

\[ \Phi(t) \]

\[ 4\pi \]
\[ 3\pi \]
\[ 2\pi \]
\[ \pi \]
\[ 0 \]

\[ \omega(t) \]

PHASE NOISE

[McNeill]
Phase Noise Definition

- An ideal oscillator has an impulse shape in the frequency domain.
- A real oscillator has phase noise “skirts” centered at the carrier frequency.
- Phase noise is quantified as the normalized noise power in a 1Hz bandwidth at a frequency offset $\Delta\omega$ from the carrier.

$$L(\Delta\omega) = 10\log\left(\frac{P_{\text{sideband}}(\omega_c + \Delta\omega, 1\text{Hz})}{P_{\text{carrier}}}\right) \text{ (dBc/Hz)}$$
Phase Noise Impact in RF Communication

- At the RX, a large interferer can degrade the SNR of the wanted signal due to “reciprocal mixing” caused by the LO phase noise.
- Having large phase noise at the TX can degrade the performance of a nearby RX.
Jitter Impact in HS Links

- RX sample clock jitter reduces the timing margin of the system for a given bit-error-rate
- TX jitter also reduces timing margin, and can be amplified by low-pass channels
Ideal Oscillator Phase Noise

The tank resistance will introduce thermal noise

$$\frac{\bar{i}_n^2}{\Delta f} = \frac{4kT}{R}$$

The spectral density of the mean-squared noise voltage is

$$\frac{\bar{v}_n^2}{\Delta f} = \frac{\bar{i}_n^2}{\Delta f} \left| Z_{\text{tank}} \right|^2$$
Tank Impedance Near Resonance

\[
Z_{\text{tank}}(\omega) = \frac{1}{j\omega C} \left| j\omega L = \frac{j\omega L}{1 - \omega^2 LC} \right.
\]

Resonance Frequency: \( \omega_o = \frac{1}{\sqrt{LC}} \)

Consider frequencies close to resonance \( \omega = \omega_o + \Delta \omega \)

\[
Z_{\text{tank}}(\Delta \omega) = \frac{j(\omega_o + \Delta \omega)L}{1 - \omega_o^2 LC - 2\omega_o \Delta \omega LC - \Delta \omega^2 LC} \approx -\frac{j\omega_o L}{-2\omega_o \Delta \omega LC} = -\frac{j}{2\omega_o C} \left( \frac{\omega_o}{\Delta \omega} \right)
\]

Tank \( Q = R\omega_o C \Rightarrow \frac{1}{\omega_o C} = \frac{R}{Q} \)

\[
Z_{\text{tank}}(\Delta \omega) \approx -\frac{j}{2Q} \left( \frac{\omega_o}{\Delta \omega} \right)
\]

\[
|Z_{\text{tank}}(\Delta \omega)|^2 = \left( \frac{R\omega_o}{2Q \Delta \omega} \right)^2
\]
Ideal Oscillator Phase Noise

\[ \frac{v_n^2}{\Delta f} = \frac{i_n^2}{\Delta f} |Z_{\text{tank}}|^2 = \left( \frac{4kT}{R} \right) \left( \frac{R \omega_o}{2Q\Delta \omega} \right)^2 = 4kTR \left( \frac{\omega_o}{2Q\Delta \omega} \right)^2 \]

The Equipartition Theorem [Lee JSSC 2000] states that, in equilibrium, amplitude and phase - noise power are equal. Therefore, this noise power is split evenly \( \left( \frac{1}{2} \right) \) between amplitude and phase.

\[
L\{\Delta \omega\} = 10 \log \left[ \frac{\left( \frac{1}{2} \right) v_n^2}{\Delta f} \right] = 10 \log \left[ \frac{\left( \frac{1}{2} \right) 4kTR \left( \frac{\omega_o}{2Q\Delta \omega} \right)^2}{v_{\text{sig}}^2} \right] = 10 \log \left[ \frac{2kT}{P_{\text{sig}}} \left( \frac{\omega_o}{2Q\Delta \omega} \right)^2 \right] \text{ (dBc/Hz)}
\]

Phase noise due to thermal noise will display a \(-20\text{dB/dec}\) slope away from the carrier

- Phase noise improves as both the carrier power and Q increase
Other Phase Noise Sources

- Tank thermal noise is only one piece of the phase noise puzzle
- Oscillator transistors introduce their own thermal noise and also flicker (1/f) noise
Leeson Phase Noise Model

Leeson’s model modifies the previously derived expression to account for the high frequency noise floor and 1/f noise upconversion.

A empirical fitting parameter F is introduced to account for increased thermal noise.

Model predicts that the \((1/\Delta \omega)^3\) region boundary is equal to the 1/f corner of device noise and the oscillator noise flattens at half the resonator bandwidth.

\[
L(\Delta \omega) = 10 \log \left[ \frac{2 F k T}{P_{\text{sig}}} \left( 1 + \left( \frac{\omega_o}{2 Q \Delta \omega} \right)^2 \right) \left( 1 + \frac{\Delta \omega_{1/f^3}}{|\Delta \omega|} \right) \right] \quad \text{(dBc/Hz)}
\]

Noise Floor

1/f Noise
A 3.5GHz LC tank VCO Phase Noise

Measure Phase noise

-30dB/decade

-20dB/decade

-105dBc
VCO Output Spectrum Example

Make sure to account for the spectrum analyzer resolution bandwidth

RBW=10K
PN=-85dBm-(-20dBm)-10\log_{10}(10e3) = -105dBc

dBc---in dB with respect to carrier

-85dBm
Leeson Model Issues

- The empirical fitting parameter $F$ is not known in advance and can vary with different process technologies and oscillator topologies.

- The actual transition frequencies predicted by the Leeson model does not always match measured data.
Harjimiri’s Model (T. H. Lee)

- Injection at Peak (amplitude noise only)

- Injection at Zero Crossing (maximum phase noise)
A time-Varying Phase Noise model: Hajimiri-Lee model

Impulse applied to the tank to measure its sensitivity function

The impulse response for the phase variation can be represented as

\[ h_\phi(t, \tau) = \frac{\Gamma(\omega_0 \tau)}{q_{\text{max}}} u(t - \tau), \]

\( \Gamma \) is the impulse sensitivity function ISF

\( q_{\text{max}} \), the maximum charge displacement across the capacitor, is a normalizing factor
Impulse Sensitivity Function (ISF) Model

- The phase variation due to injected noise can be modeled as

\[ \Delta \phi = \Gamma(\omega_0 \tau) \frac{\Delta V}{V_{\text{max}}} = \Gamma(\omega_0 \tau) \frac{\Delta q}{q_{\text{max}}} \quad \Delta q \ll q_{\text{max}} \]

- The function \( \Gamma(\omega_0 \tau) \) is a time-varying proportionality factor called the “impulse sensitivity function”
  - Encodes information about the sensitivity of the oscillator to an impulse injected at phase \( \omega_0 \tau \)
  - Phase shift is assumed linear to charge injection
  - ISF has the same oscillation period as the oscillator

- The phase impulse response can be written as

\[ h_\phi(t, \tau) = \frac{\Gamma(\omega_0 \tau)}{q_{\text{max}}} u(t - \tau) \]
How to obtain the Impulse sensitivity function for a LC oscillator

\[ \Gamma(\omega \tau) \] can be obtained using Cadence

Consider the effect on phase noise of each noise source
Typical ISF Example

- The ISF can be estimated analytically or calculated from simulation.
- The ISF reaches peak during zero crossing and zero at peak for typical LC and ring oscillators.
Phase Noise Computation

The impulse sensitivity function is used to obtain the phase noise impulse function

\[ h_\phi(t, \tau) = \frac{\Gamma(\omega_0 \tau)}{q_{\text{max}}} u(t - \tau) \]

The phase noise can then be computed by the superposition (convolution) integral of the any arbitrary noise current with the phase noise impulse function

\[ \phi(t) = \int_{-\infty}^{\infty} h_\phi(t, \tau) i(\tau) d\tau = \frac{1}{q_{\text{max}}} \int_{-\infty}^{t} \Gamma(\omega_0 \tau) i(\tau) d\tau \]
In order to gain further insight, and because the ISF is periodic, it may be expressed as a Fourier series

\[
\Gamma(\omega_o \tau) = \frac{c_0}{2} + \sum_{n=1}^{\infty} c_n \cos(n\omega_o \tau + \theta_n)
\]

where the coefficients \(c_n\) are real and \(\theta_n\) is the phase of the \(n\)th ISF harmonic. Note, \(\theta_n\) is typically ignored, as it is assumed that the noise components are uncorrelated, and their relative phase is irrelevant.

The phase noise can then be computed by

\[
\phi(t) = \frac{1}{q_{\text{max}}} \left[ \frac{c_0}{2} \int_{-\infty}^{t} i(\tau) d\tau + \sum_{n=1}^{\infty} c_n \int_{-\infty}^{t} i(\tau) \cos(n\omega_o \tau) d\tau \right]
\]

This allows the excess phase from an arbitrary noise source to be computed once the ISF Fourier coefficients are determined. Essentially, the current noise is mixed down from different frequency bands and scaled according to the ISF coefficients.
Phase Noise Frequency Conversion

First consider a simple case where we have a sinusoidal noise current whose frequency is near an integer multiple $m$ of the oscillation frequency

$$i(t) = I_m \cos[(m\omega_0 + \Delta\omega)t]$$

When performing the phase noise computation integral, there will be a negligible contribution from all terms other than $n = m$

$$\varphi(t) \approx \frac{I_m c_m \sin(\Delta\omega t)}{2q_{\max} \Delta\omega}$$

The resulting frequency spectrum will show two equal sidebands at $\pm \Delta\omega$. Assuming a sinusoidal waveform $v_{out}(t) = \cos[\omega_0 t + \varphi(t)]$, there will be two equally weighted sidebands symmetric about the carrier with power

$$P_{SBC}(\Delta\omega) \approx 10\log\left(\frac{I_m c_m}{4q_{\max} \Delta\omega}\right)^2$$

Note that this power is proportional to $\left(\frac{1}{\Delta\omega}\right)^2$.

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Figure 4.18: The conversion of tones in the vicinity of integer multiples of $\omega_0$. 

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Phase Noise Due to White & 1/f Sources

Extending the previous analysis to the general case of a white noise source results in

\[
P_{SBC}(\Delta \omega) \approx 10 \log \left( \frac{i_n^2}{\Delta f} \sum_{m=0}^{\infty} \frac{c_m^2}{4q_{\text{max}}^2 \Delta \omega^2} \right)
\]

Here noise components near integer multiples of the carrier frequency all fold near the carrier itself and are weighted by \( \left( \frac{1}{\Delta \omega} \right)^2 \).

Noise near dc gets upconverted, weighted by coefficient \( c_0 \), so \( 1/f \) noise becomes \( 1/f^3 \) noise near the carrier.

White noise near the carrier is weighted by \( c_1 \) and \( 1/f^2 \) and stays near the carrier.

White noise near higher integer multiples of the carrier gets downconverted and weighted by \( c_m \) and \( 1/f^2 \).
How to Minimize Phase Noise?

In order to minimize phase noise, the ISF coefficients \( c_n \) should be minimized. Using Parseval's theorem

\[
\sum_{m=0}^{\infty} c_m^2 = \frac{1}{\pi} \int_{0}^{2\pi} \left| \Gamma(x) \right|^2 dx = 2\Gamma_{\text{rms}}^2
\]

The spectrum in the \( 1/f^2 \) region can be expressed as

\[
L(\Delta \omega) = 10 \log \left( \frac{\frac{\Delta f}{\Delta f_{\text{rms}}} \Gamma_{\text{rms}}^2}{2 q_{\text{max}}^2 \Delta \omega^2} \right)
\]

Thus, reducing \( \Gamma_{\text{rms}} \) will reduce the phase noise at all frequencies.
1/f Corner Frequency

Consider current noise which includes 1/f content

\[ \overline{i_n^2_{1/f}} = i_n^2 \frac{\omega_{1/f}}{\Delta \omega} \]

where \( \omega_{1/f} \) is the 1/f corner frequency

From the previous slide

\[ L(\Delta \omega) = 10 \log \left( \frac{i_n^2}{\Delta f c_0^2} \frac{\omega_{1/f}}{8 q_{\max}^2 \Delta \omega^2 \Delta \omega} \right) \]

Thus, the 1/f³ corner frequency is

\[ \Delta \omega_{1/f^3} = \omega_{1/f} \frac{c_0^2}{4 \Gamma_{rms}^2} = \omega_{1/f} \left( \frac{\Gamma_{dc}}{\Gamma_{rms}} \right)^2 \]

This is generally lower than the 1/f device/circuit noise corner. If \( \Gamma_{dc} \) is minimized through rise - and fall - time symmetry, then there is the potential for dramatic reductions in 1/f noise.
Cyclostationary Noise Treatment

Transistor drain current, and thus noise, can change dramatically over an oscillator cycle. The LTV model can easily handle this by treating it as the product of stationary white noise and a periodic function.

\[ i_n(t) = i_{n0}(t)\alpha(\omega_0 t) \]

Here \( i_{n0} \) is a stationary white noise source whose peak value is equal to that of the cyclostationary noise source, and \( \alpha(x) \) is a periodic unitless function with a peak value of unity. Using this, we can formulate an effective ISF

\[ \Gamma_{NMF}(x) = \Gamma(x)\alpha(x) \]
Key Oscillator Design Points from Hajimiri Model

- As the LTI model predicts, oscillator signal power and Q should be maximized.
- Ideally, the energy returned to the tank should be delivered all at once when the ISF is minimum.
- Oscillators with symmetry properties that have small $\Gamma_{dc}$ will provide minimum 1/f noise upconversion.
Phasor-Based Phase Noise Analysis

Physical Processes of Phase Noise in Differential LC Oscillators

J. J. Rael and A. A. Abidi

\[ V_{\text{out}} = V_1 \cos(\omega_0 t) + \phi_1 [\cos(\omega_1 t) - \cos(\omega_2 t)] + \alpha_1 [\cos(\omega_1 t) + \cos(\omega_2 t)] + \phi_2 [\sin(\omega_1 t) + \sin(\omega_2 t)] + \alpha_2 [\sin(\omega_1 t) - \sin(\omega_2 t)] \]

- Models noise at 2 sideband frequencies with modulation terms
- The \( \alpha_1 \) and \( \alpha_2 \) terms sum co-linear with the carrier phasor and produce amplitude modulation (AM)
- The \( \phi_1 \) and \( \phi_2 \) terms sum orthogonal with the carrier phasor and produce phase modulation (PM)
This phasor-based approach can be used to find closed-form expressions for LC oscillator phase noise that provide design insight.

In particular, an accurate expression for the Leeson model $F$ parameter is obtained:

$$L\{\Delta \omega\} = 10\log \left[ \frac{2FkT}{P_{sig}} \left( \frac{\omega_o}{2Q\Delta \omega} \right)^2 \right] \text{ (dBc/Hz)}$$
LC Oscillator F Parameter

\[ L\{\Delta \omega\} = 10 \log \left[ \frac{2FkT}{P_{\text{sig}}} \left( \frac{\omega_o}{2Q\Delta \omega} \right)^2 \right] \text{ (dBc/Hz)} \]

\[ F = 1 + \frac{4\gamma I R}{\pi V_o} + \gamma \frac{4}{9} g_{m\text{bias}} R \]

- 1st Term = Tank Resistance Noise
- 2nd Term = Cross-Coupled Pair Noise
- 3rd Term = Tail Current Source Noise

- The above expression gives us insight on how to optimize the oscillator to reduce phase noise
- The tail current source is often a significant contributor to total noise
Loading in Current-Biased Oscillator

- The current source plays 2 roles
  - It sets the oscillator bias current
  - Provides a high impedance in series with the switching transistors to prevent resonator loading
Tail Current Noise

- The switching differential pair can be modeled as a mixer for noise in the current source.
- Low frequency noise only produces amplitude noise, not phase noise.
- Only the noise located at even harmonics will produce phase noise.

![Current Noise PSD](image)
Noise Filtering in Oscillator

- Only thermal noise in the current source transistor around 2\textsuperscript{nd} harmonic of the oscillation causes phase noise.
- In balanced circuits, odd harmonics circulate in a differential path, while even harmonics flow in a common-mode path.
- A high impedance at the tail is only required at the 2\textsuperscript{nd} harmonic to stop the differential pair FETs in triode from loading the resonator.

How can we present a low-impedance for the 2\textsuperscript{nd} harmonic noise current to filter it and a high-impedance to the tank at the 2\textsuperscript{nd} harmonic to avoid loading the tank?
Noise Filtering in Oscillator

- Tail-biased VCO with noise filtering.
Phase Noise w/ Tail Current Filtering

- Tail current noise filtering provides near 7dB improvement
Noise Filtering in Oscillator

- A top-biased VCO often provides improved substrate noise rejection and reduced flicker noise.
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Open-Loop VCO Jitter

- Measure distribution of clock threshold crossings
- Plot $\sigma$ as a function of delay $\Delta T$
Open-Loop VCO Jitter

- Jitter $\sigma$ is proportional to $\sqrt{\Delta T}$
- $\kappa$ is VCO time domain figure of merit

\[ \sigma_{\Delta T(OL)}(\Delta T) \approx \kappa \sqrt{\Delta T} \]
VCO in Closed-Loop PLL Jitter

- PLL limits $\sigma$ for delays longer than loop bandwidth $\tau_L$

$$\tau_L = \frac{1}{2\pi f_L}$$
Ref Clk-Referenced vs Self-Referenced

• Generally, we care about the jitter w.r.t. the ref. clock ($\sigma_x$)
• However, may be easier to measure w.r.t. delayed version of output clk
  • Due to noise on both edges, this will be increased by a sqrt(2) factor relative to the reference clock-referred jitter
Converting Phase Noise to Jitter

- RMS jitter for $\Delta T$ accumulation
  $$\sigma_{\Delta T}^2 = \frac{8}{\omega_0^2} \int_0^\infty S_\phi(f) \sin^2(\pi f \Delta T) df$$

- As $\Delta T$ goes to $\infty$
  $$\sigma_T^2 = \frac{2}{\omega_0^2} R_\phi(0) = \frac{4}{\omega_0^2} \int_0^\infty S_\phi(f) df$$

- Integration range depends on application bandwidth
  - $f_{\text{min}}$ set by standard
    - Ex. Assumed CDR tracking bandwidth
  - Usually stop integration at $f_o/2$ or $f_o$ due to measurement limitations and aliasing components

[Mansuri]
Next Time

• Divider Circuits