Oscillators

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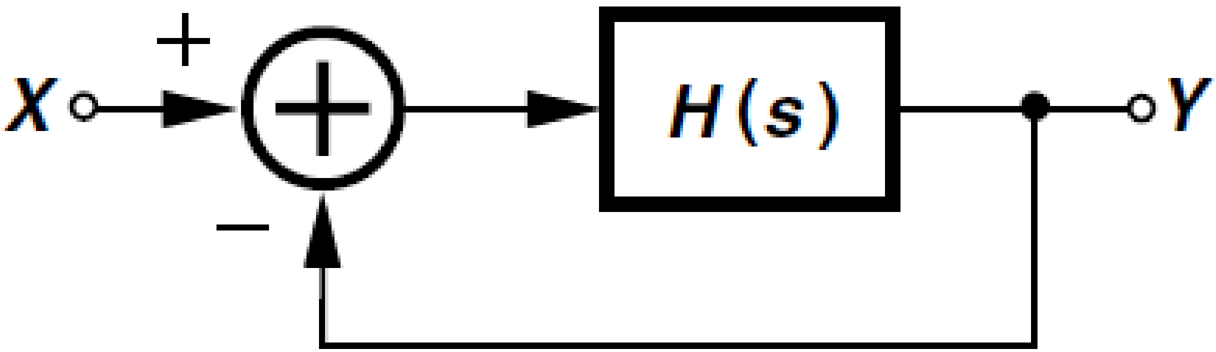
Feedback View



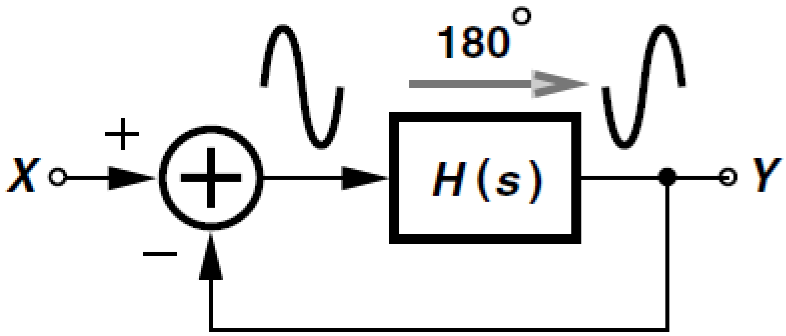
$$\frac{Y}{X}(s) = \frac{H(s)}{1 + H(s)} \qquad |H(s = j\omega_1)| = 1$$

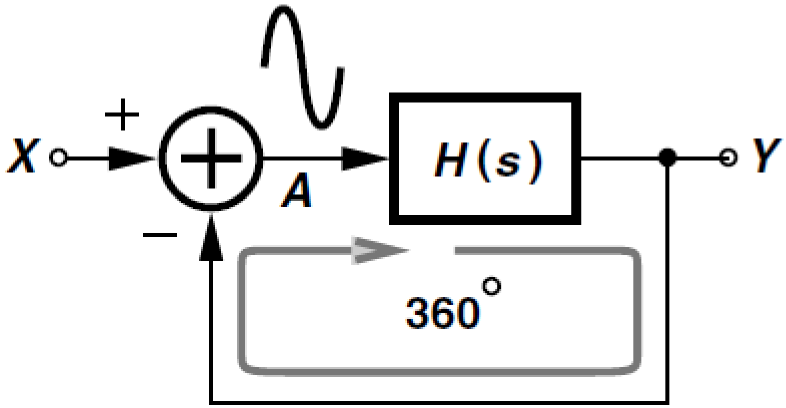
$$\angle H(s = j\omega_1) = 180^{\circ}$$

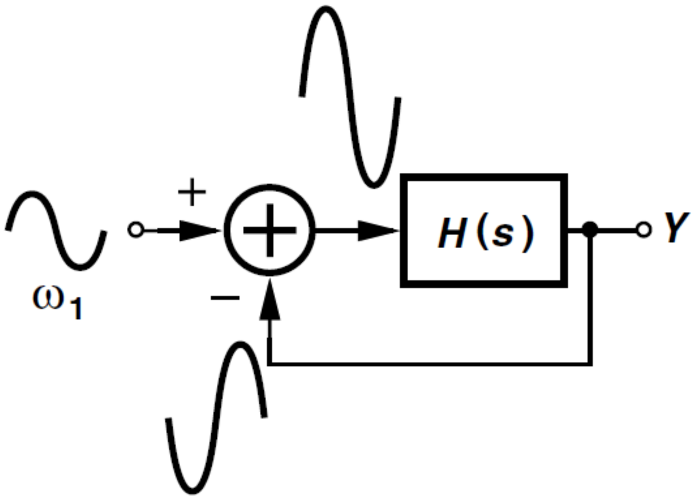
- **If at a sinusoidal frequency, ω_1, $H(s=j\omega_1)$ becomes equal to -1?**
- **The gain from the input to the output goes to infinity, allowing the circuit to amplify a small component at ω_1 indefinitely.**
- **That is, the circuit can sustain an output at ω_1.**

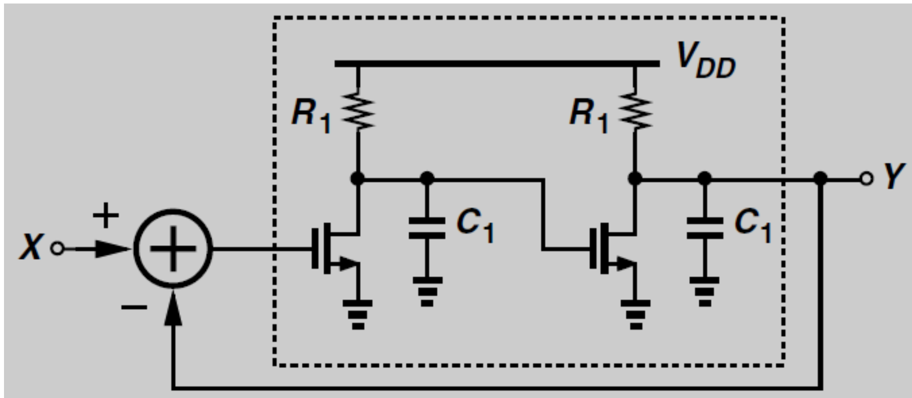
[B. Razavi, RF Microelectronics]

Phase Relationship



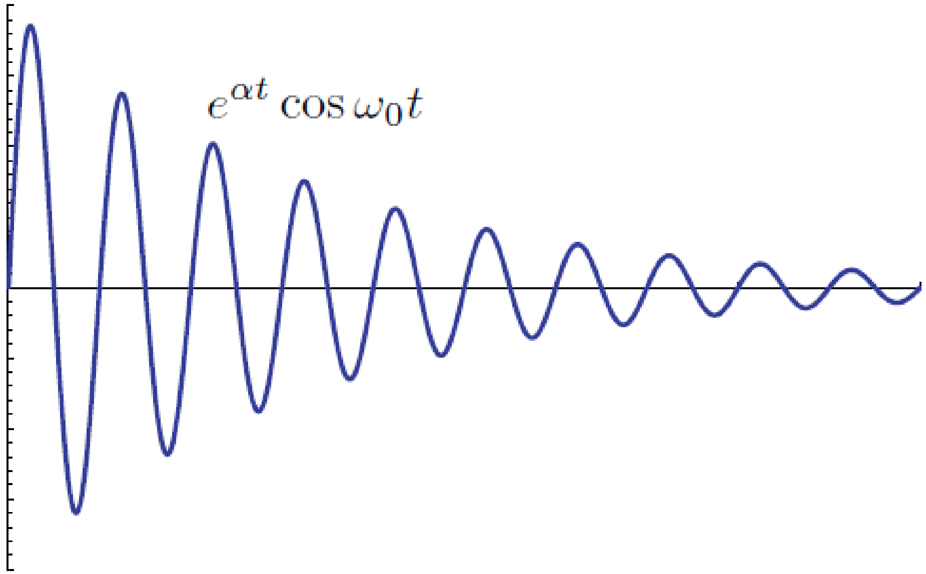






[B. Razavi, RF Microelectronics]

An LC-Tank Oscillator



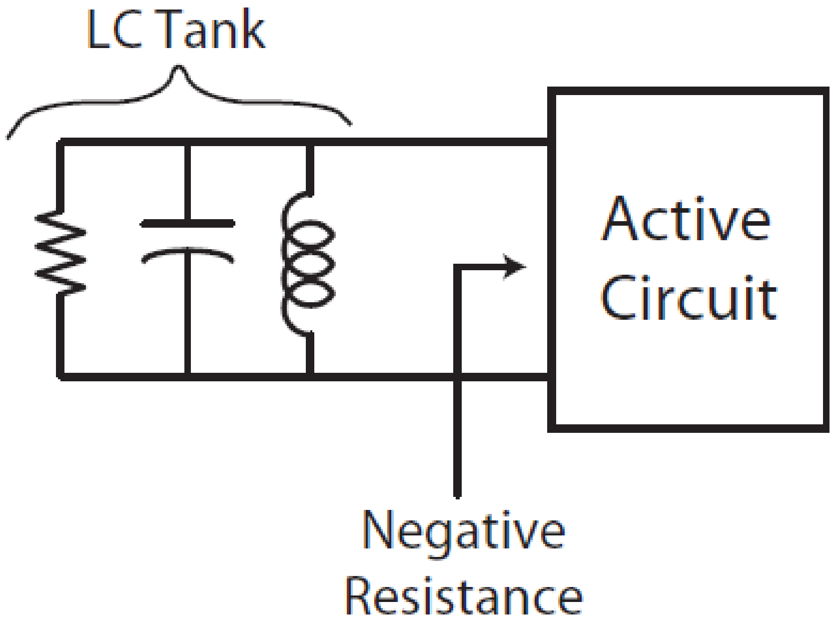
❑**Note that an LC tank alone is not a good oscillator. Due to loss, no matter how small, the amplitude of the oscillator decays.**

❑**Even a very high Q oscillator can only sustain oscillations for about Q cycles. For instance, an LC tank at 1GHz has a Q=20, can only sustain oscillations for about 20ns.**

❑**Even a resonator with high Q=1000, will only sustain oscillations for about 1ms.**

[Ali Niknejad, EE142&EE242 of UC Berkeley]

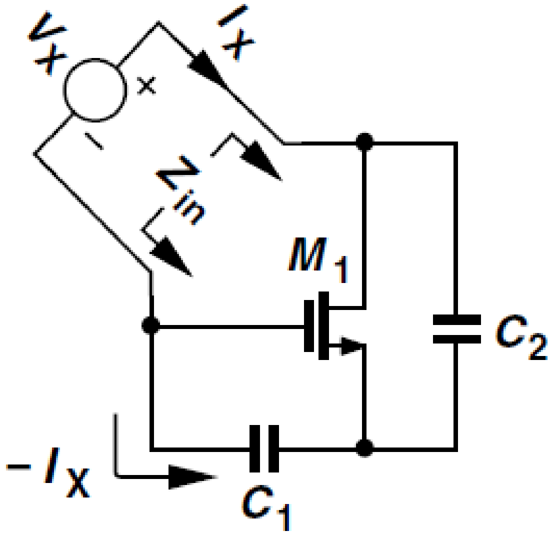
Negative Resistance Perspective

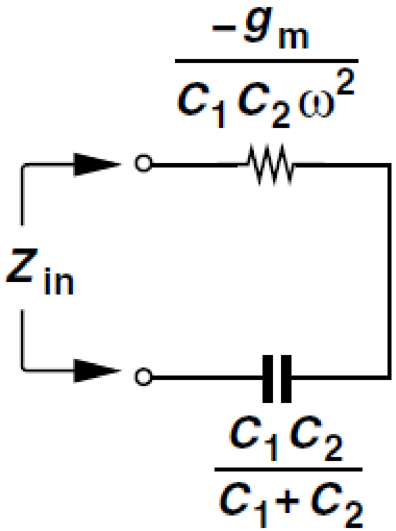


❑**Another perspective is to view the active device as a negative resistance generator. In steady state, the losses in the tank due to resistance R are balanced by the power drawn from the active device through the negative resistance -R.**

[Ali Niknejad, EE142&EE242 of UC Berkeley]

Negative Resistance





$$-\frac{I_X}{C_1 s} + V_X = \left(I_X + I_X \frac{g_m}{C_1 s}\right)\frac{1}{C_2 s}$$

$$\frac{V_X}{I_X}(s) = \frac{1}{C_1 s} + \frac{1}{C_2 s} + \frac{g_m}{C_1 C_2 s^2}$$

$$\frac{V_X}{I_X}(j\omega) = \frac{1}{jC_1\omega} + \frac{1}{jC_2\omega} - \frac{g_m}{C_1 C_2 \omega^2}$$

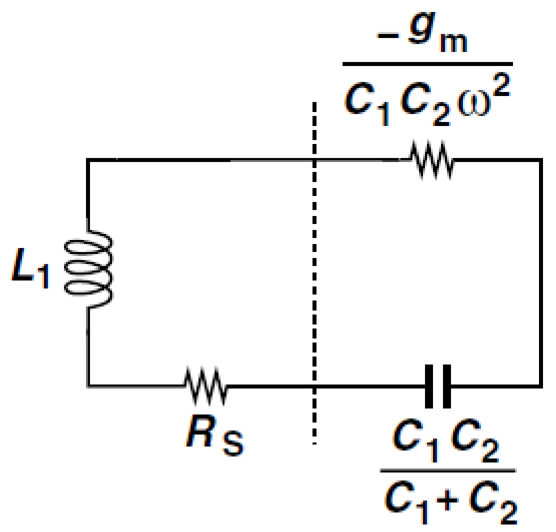
[B. Razavi, RF Microelectronics]

Negative Resistance

- Attach the negative resistance to a lossy tank so as to construct an oscillator.
- Connect an inductor to the negative-resistance port, and model the loss of the inductor by a series resistance, R_S.

$$R_S = \frac{g_m}{C_1 C_2 \omega^2}$$

$$\omega_{osc} = \frac{1}{\sqrt{L_1 \frac{C_1 C_2}{C_1 + C_2}}}$$



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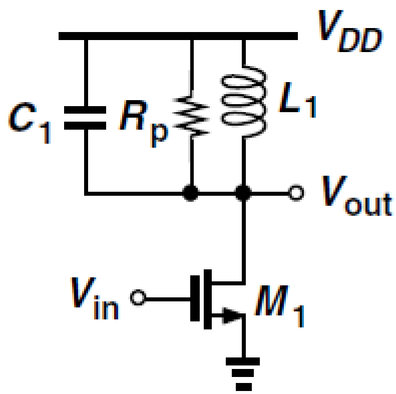
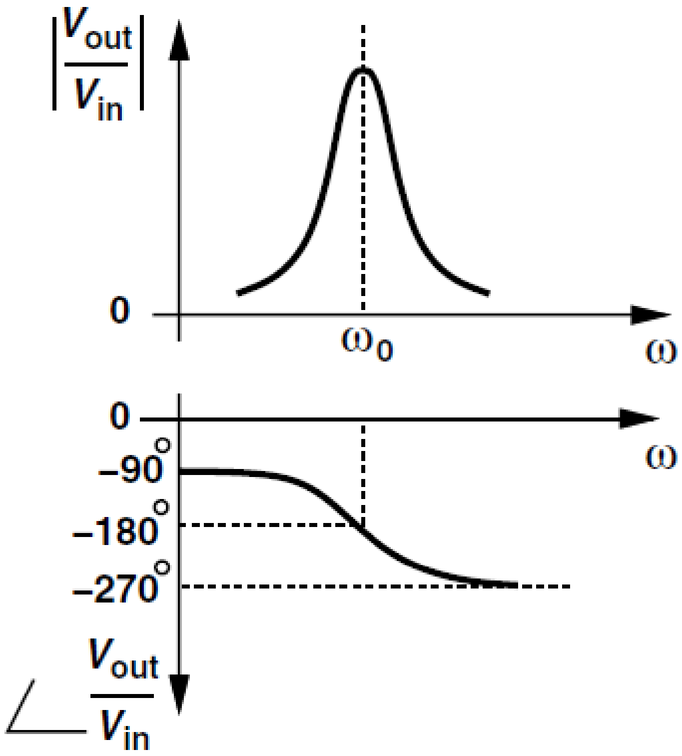
Tuned Amplifier

- ❑A negative-feedback oscillatory system is built using "LC-tuned" amplifier stages.





- ❑At the resonance frequency, ω_0, the tank reduces to R_p and

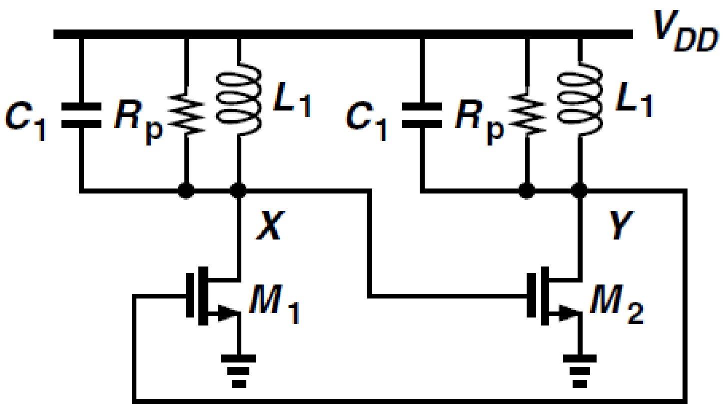
$$\frac{V_{out}}{V_{in}} = -g_m R_p$$

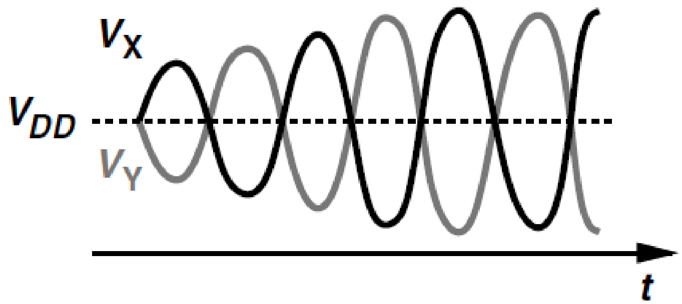
[B. Razavi, RF Microelectronics]

Tuned Amplifier

❑Upon closer examination, we recognize that the circuit provides a phase shift of 180° with possibly adequate gain ($g_m R_p$) at ω_0.

❑We simply need to increase the phase shift to 360°, by inserting another stage in the loop





❑The circuit oscillates if the loop gain: $(g_m R_p)^2 \geq 1$

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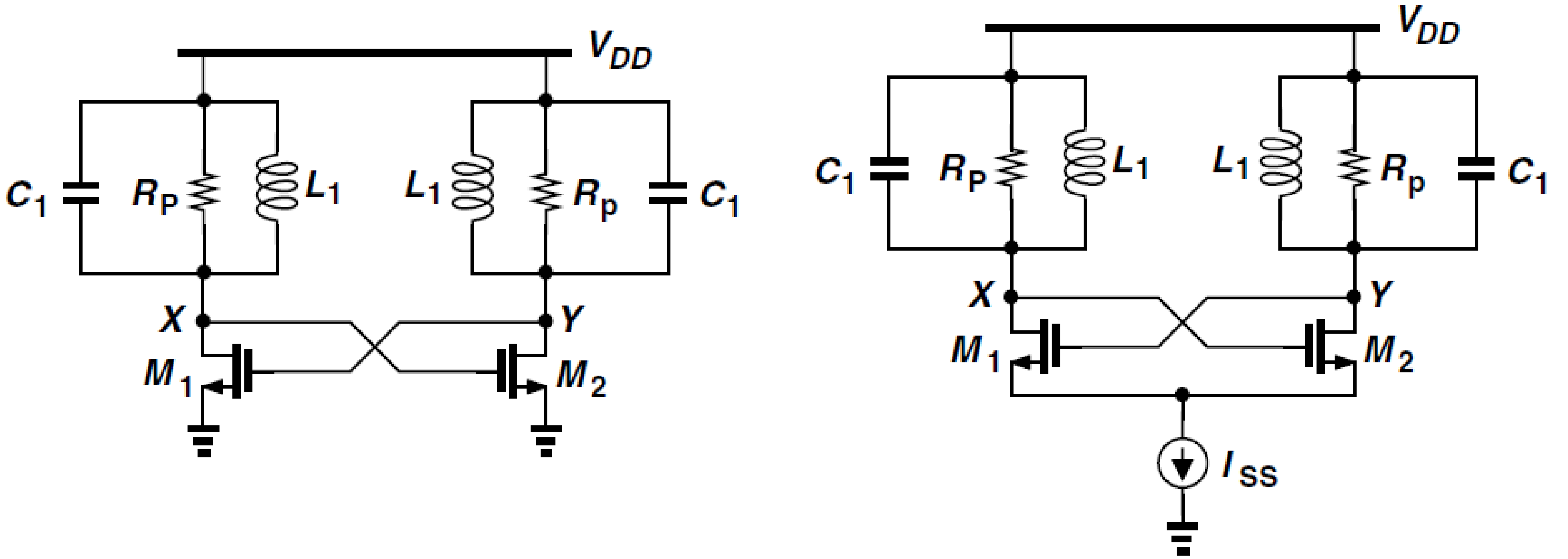
Cross-Coupled Oscillator

- The above circuit can be redrawn and is called a "cross-coupled" oscillator

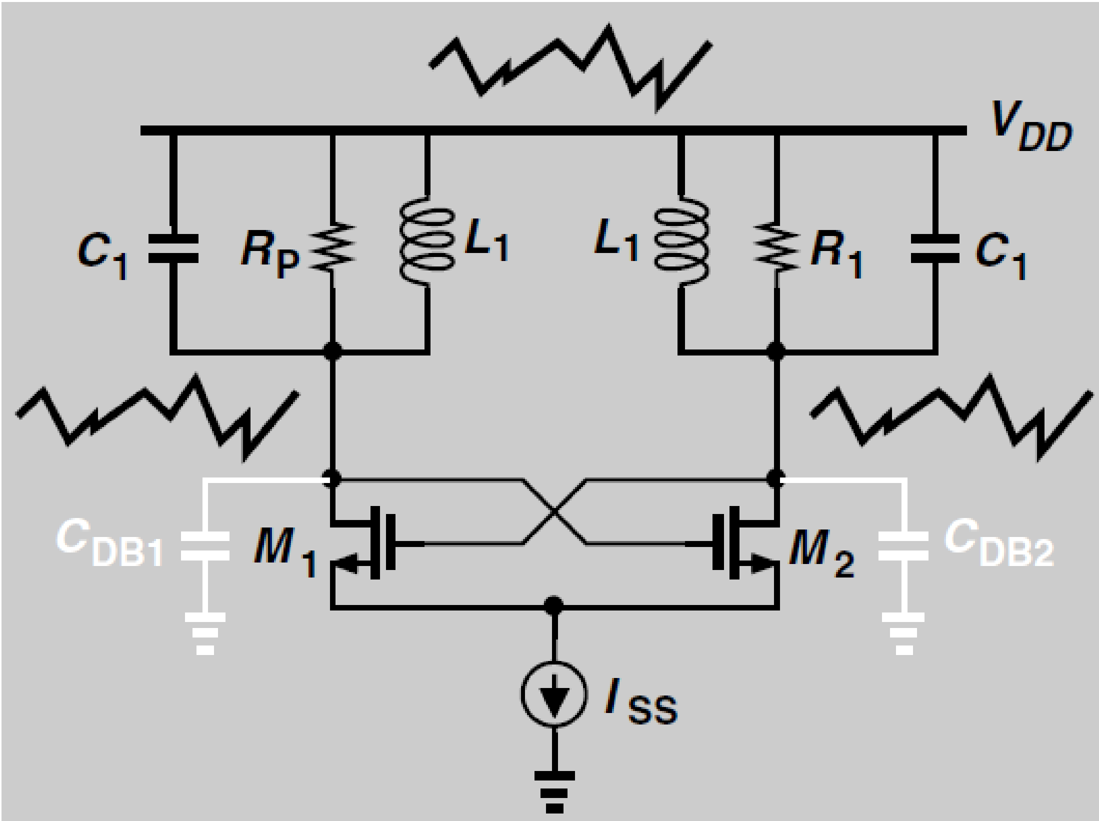


- The oscillating frequency is determined by the L_1, C_1, and parasitic capacitance

[B. Razavi, RF Microelectronics]

Supply Interference

- Supply variations modulate this capacitance and hence the oscillation frequency.



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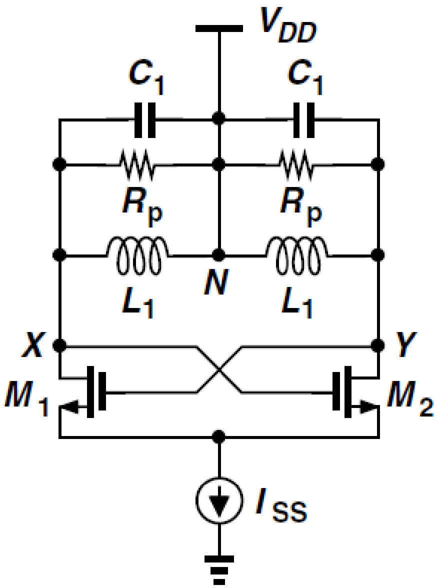
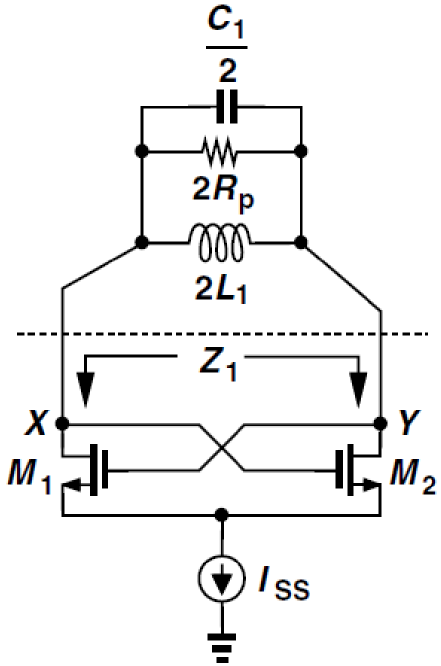
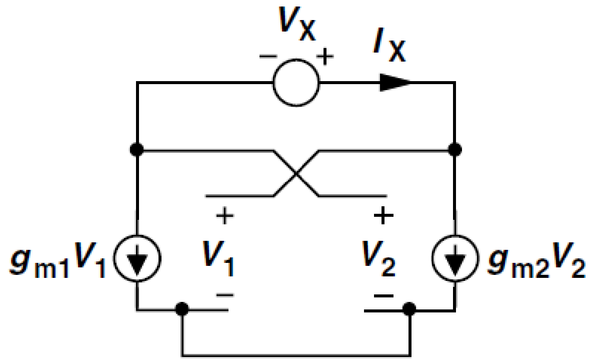
Oscillation Startup







$$I_X = -g_{m1}V_1 = g_{m2}V_2$$

$$\frac{V_X}{I_X} = -\left(\frac{1}{g_{m1}} + \frac{1}{g_{m2}}\right)$$

$$\frac{V_X}{I_X} = -\frac{2}{g_m} \qquad \frac{2}{g_m} \le 2R_p \qquad g_m R_p \ge 1$$

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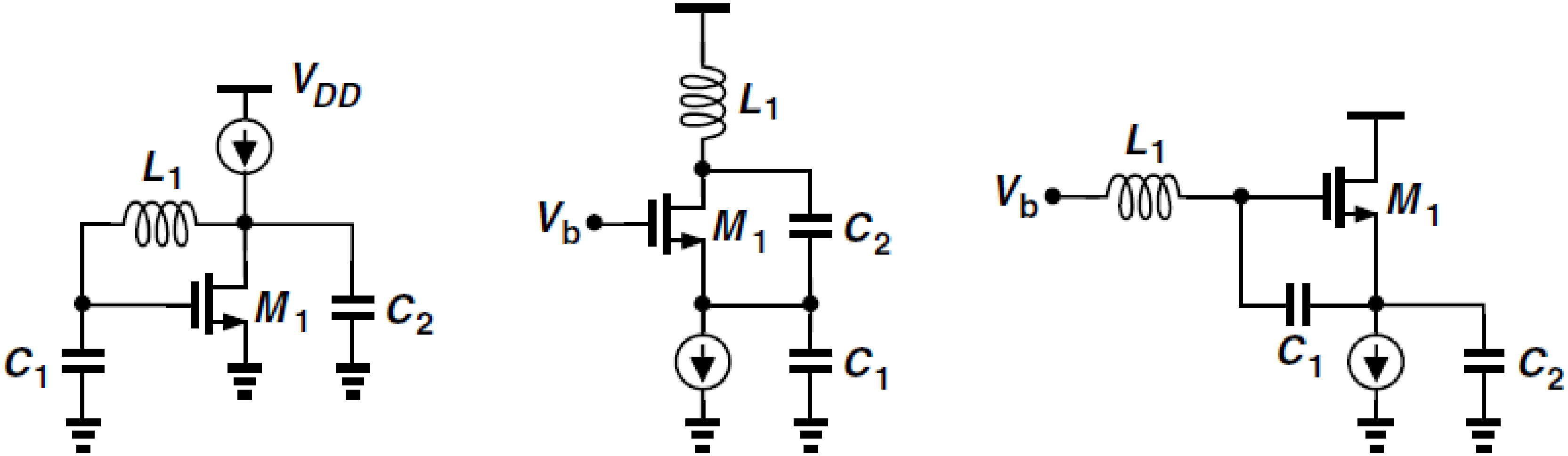
Three-Point Oscillator

- Three different oscillator topologies can be obtained by grounding each of the transistor terminals

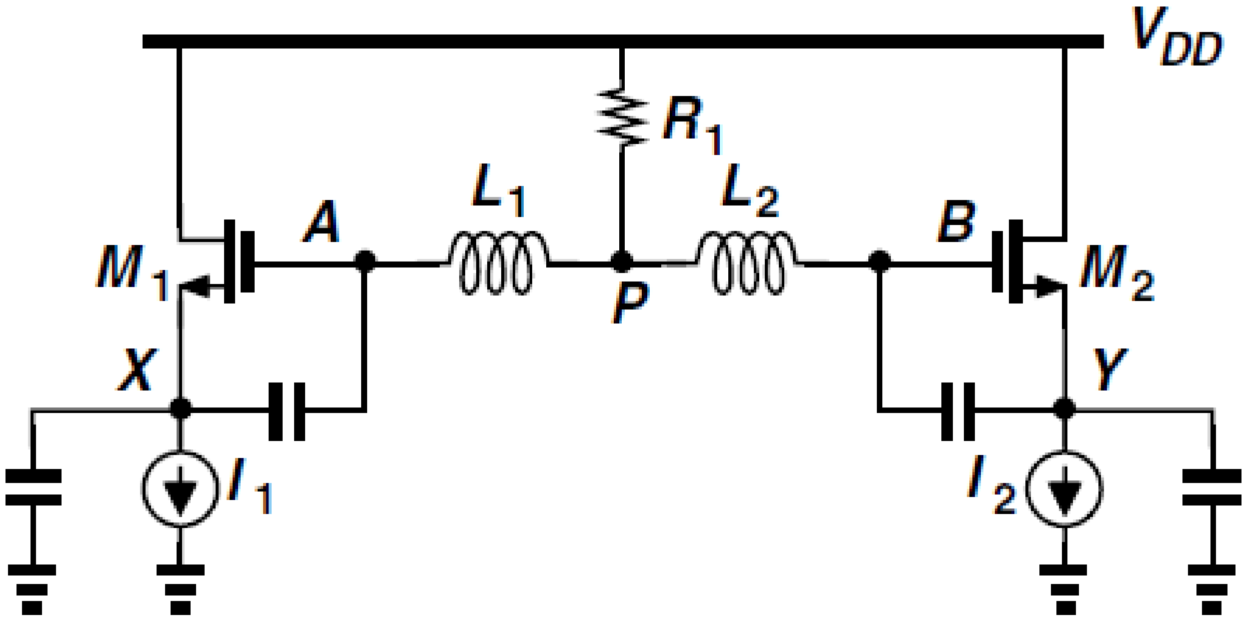


- There is only one MOSFET for negative resistance, so the start-up condition is more stringent than cross-coupled one
- The circuits produce only single-ended output

[B. Razavi, RF Microelectronics]

Three-Point Oscillator

- It is possible to couple two copies of one oscillator so that they operate differentially.



- The noise of I_1 and I_2 directly corrupts the oscillation.

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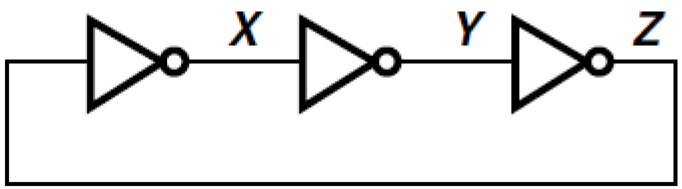
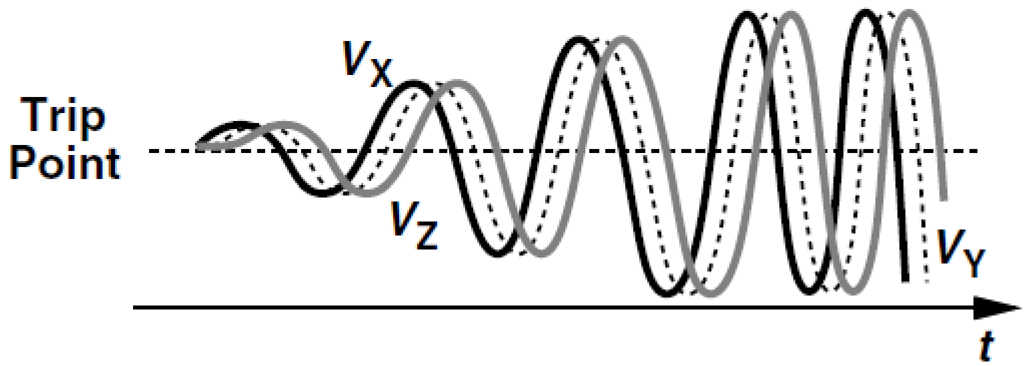
Ring Oscillator





❑Each stage operates as an amplifier, leading to an oscillation frequency at which each inverter contributes a frequency-dependent phase shift of 60^{O}.

❑In the steady state, the output of each inverter swings from nearly zero to nearly V_{DD}.

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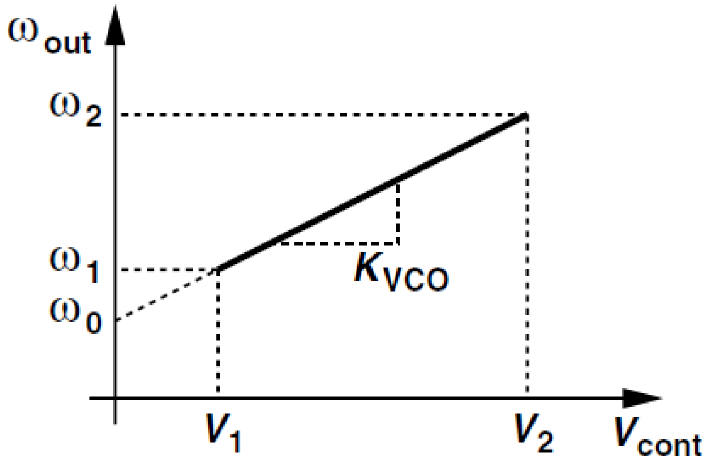
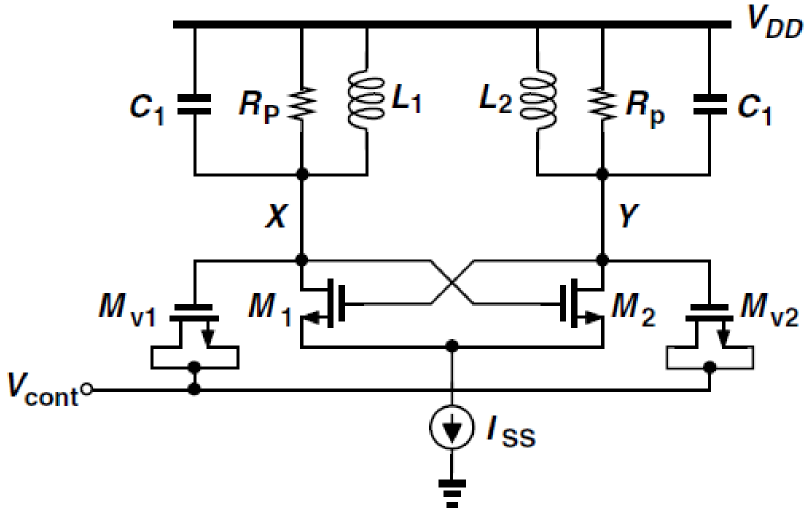
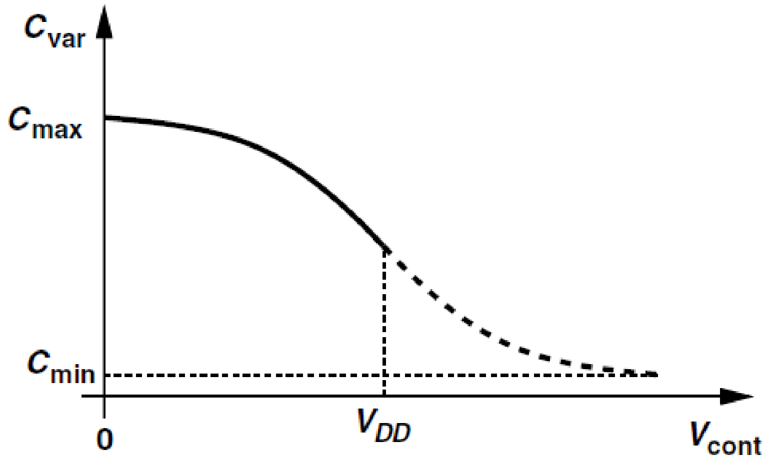
Voltage-Controlled Oscillator (VCO)







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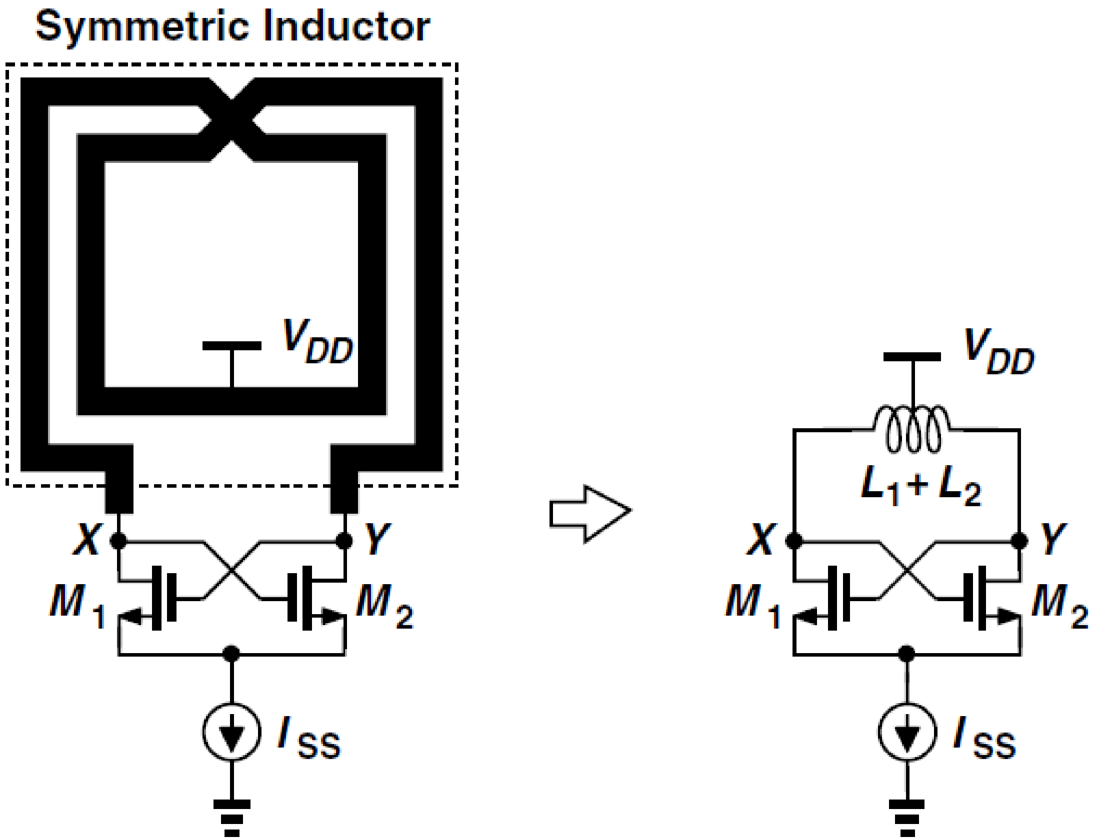
Symmetric Inductor

❑Symmetric spiral inductors excited by differential waveforms exhibit a higher Q than their single-ended counterparts



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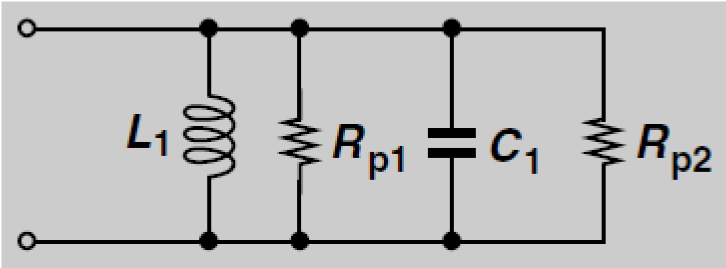
Overall Q

- The loss of an inductor or a capacitor can be modeled by a parallel resistance



$$Q_L = \frac{R_{p1}}{L_1\omega}$$

$$Q_C = R_{p2}C_1\omega$$

- In the vicinity of resonance $L_1\omega = (C_1\omega)^{-1}$

- Merging R_{p1} and R_{p2} yields the overall Q:

$$Q_{tot} = \frac{R_{p1}R_{p2}}{R_{p1}+R_{p2}} \cdot \frac{1}{L_1\omega} = \frac{1}{\dfrac{L_1\omega}{R_{p1}} + \dfrac{1}{R_{p2}C_1\omega}}$$

$$\boxed{\frac{1}{Q_{tot}} = \frac{1}{Q_L} + \frac{1}{Q_C}}$$

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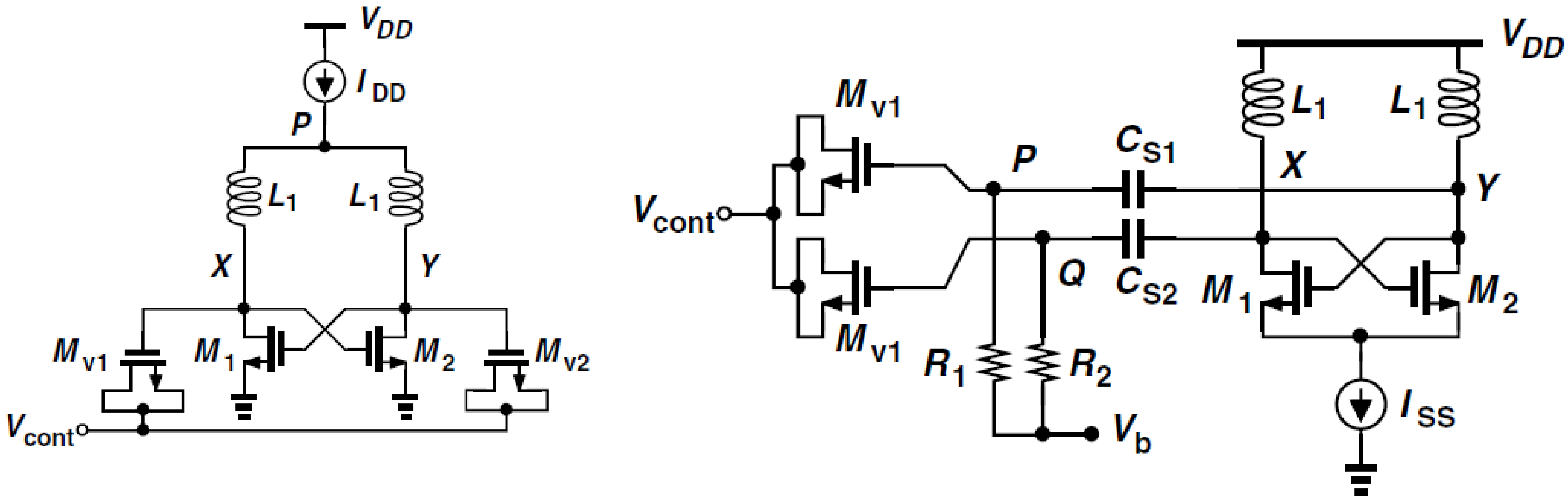
VCO Bias

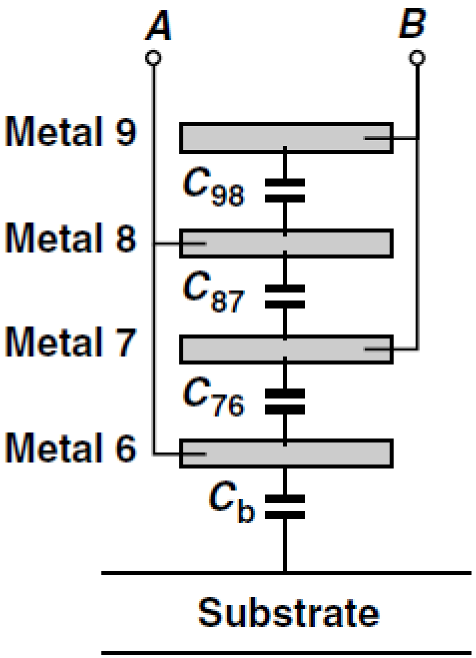


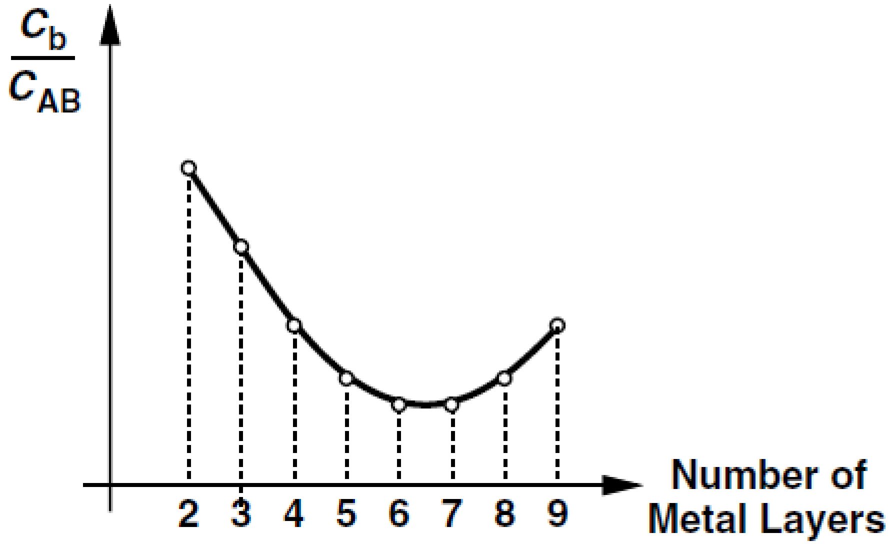
❑**The VCO can be tail biased or top biased**

❑**AC coupling can be used to set a free bias voltage, but the tuning range will be reduced**

[B. Razavi, RF Microelectronics]

Impact of MIM Cap





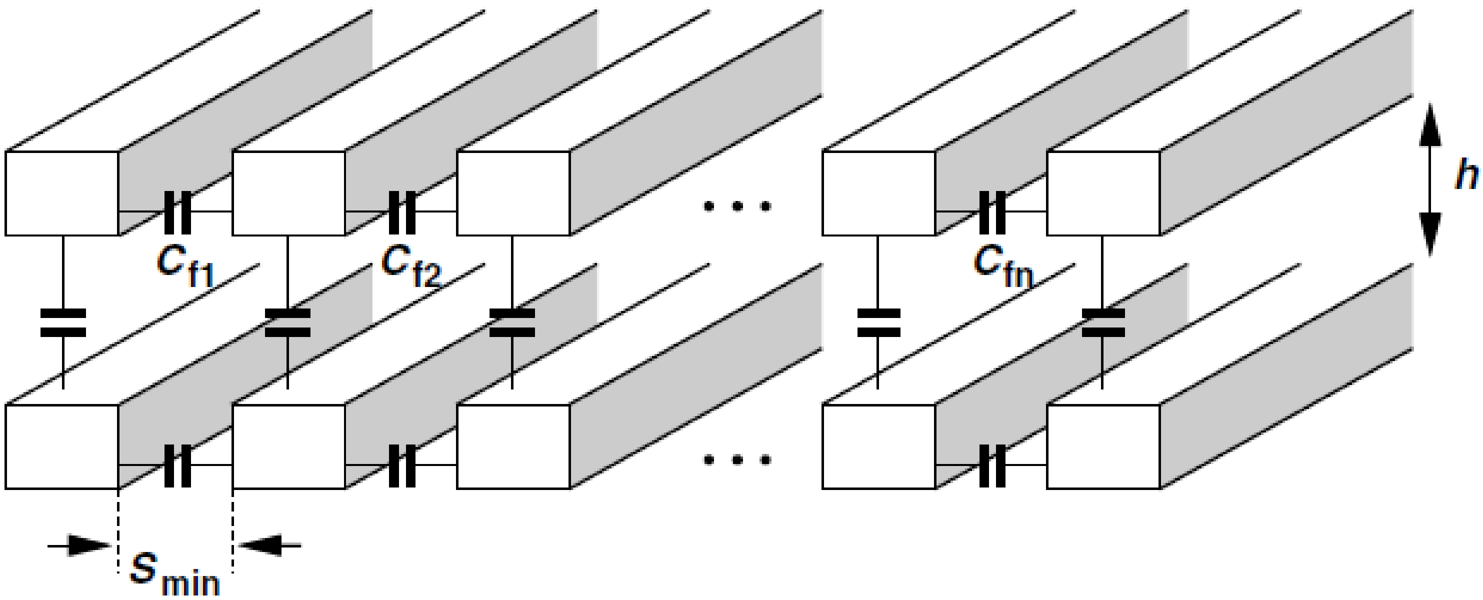
❑More metal layers increase the overall capacitance C_{AB}

❑As more layers are stacked, the capacitance between the bottom layer and the substrate increase

[B. Razavi, RF Microelectronics]

MOM Cap

- A MOM capacitor can exhibits lower parasitics than the MIM structure
- The capacitance per unit volume is larger than that of the metal sandwich, leading to a smaller parasitic.



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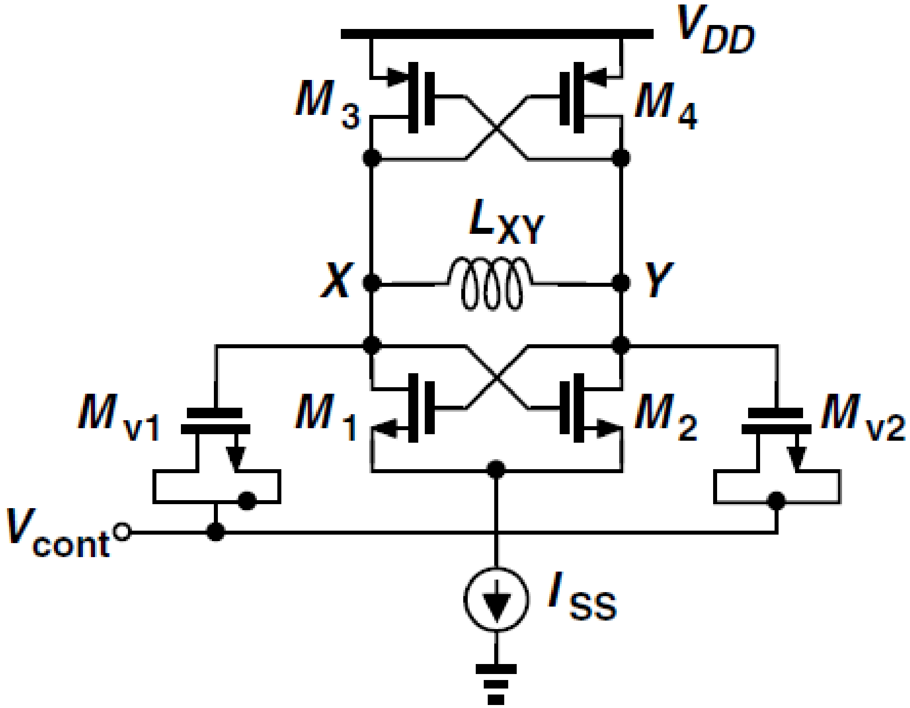
PMOS & NMOS Cross-Coupled Pairs

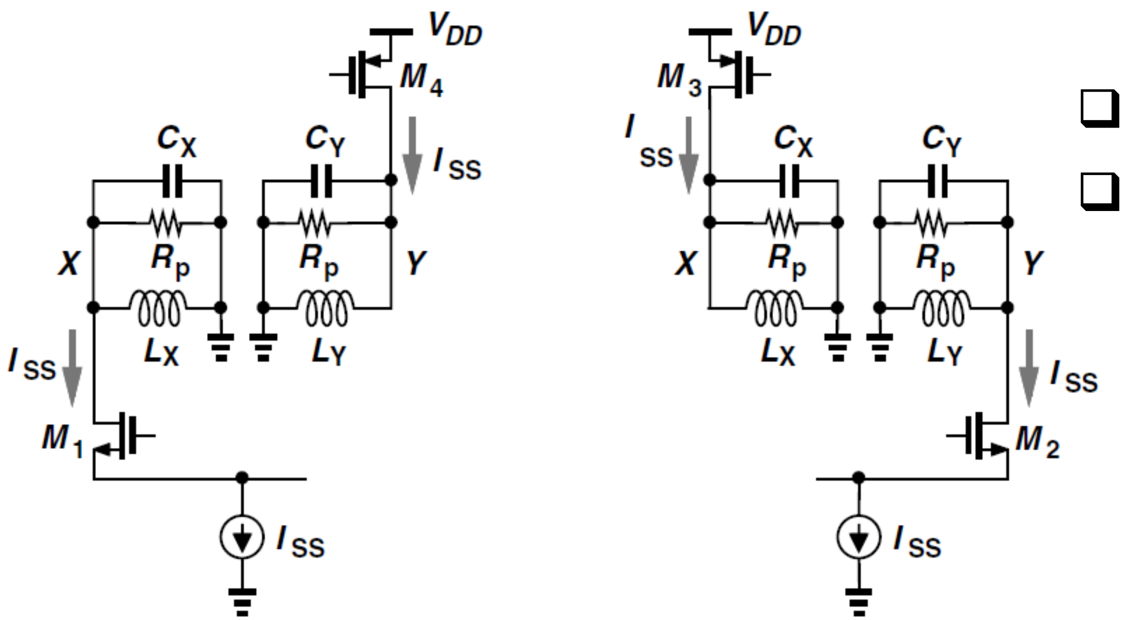


- ❑The VCO topology naturally provides an output CM level approximately equal to $V_{DD}/2$
- ❑The bias current is "reused" by the PMOS devices, providing a higher transconductance

[B. Razavi, RF Microelectronics]

PMOS & NMOS Cross-Coupled Pairs

- The typology produces twice the voltage swing for a given bias current and inductor design
- The current in each tank swings between I_{SS} and $-I_{SS}$, whereas in previous topologies it swings between I_{SS} and zero. The output voltage swing is therefore doubled

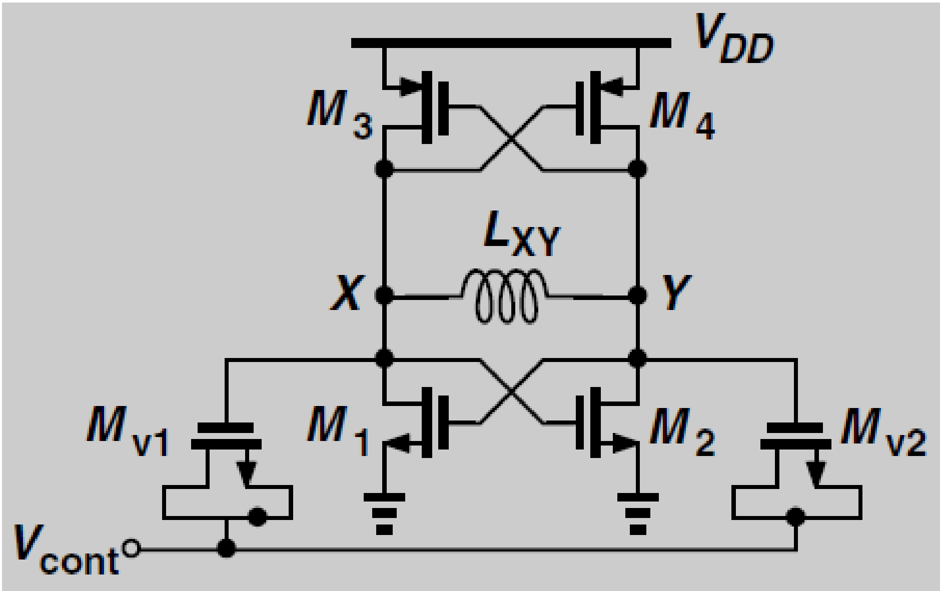


- Large voltage headroom
- Large parasitics

[B. Razavi, RF Microelectronics]

PMOS & NMOS Cross-Coupled Pairs

- If we remove the noise of the tail current source:

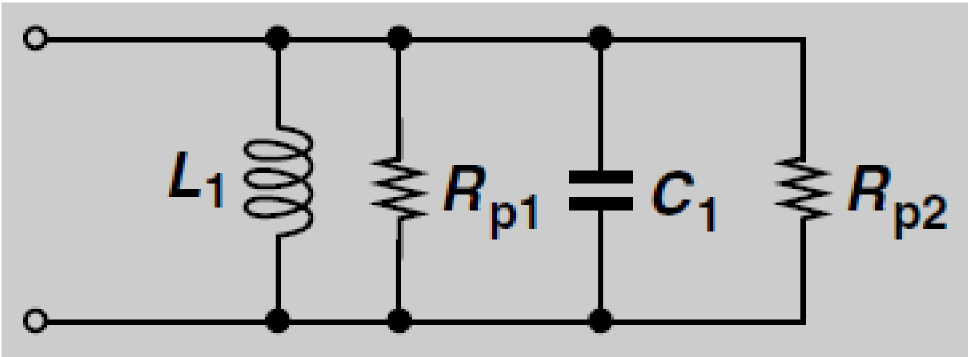


- The circuit indeed avoids frequency modulation due to the tail current noise.
- It saves the voltage headroom associated with the tail current
- The circuit is now very sensitive to the supply voltage.

[B. Razavi, RF Microelectronics]

Amplitude Variation

❑Frequency tuning induces variation of oscillation amplitude:



$$Q = \frac{L_1\omega}{R_S} = \frac{R_p}{L_1\omega}$$

$$R_p = \frac{L_1^2\omega^2}{R_S}$$

❑R_p falls in proportion to ω^2 as more capacitance is presented to the tank

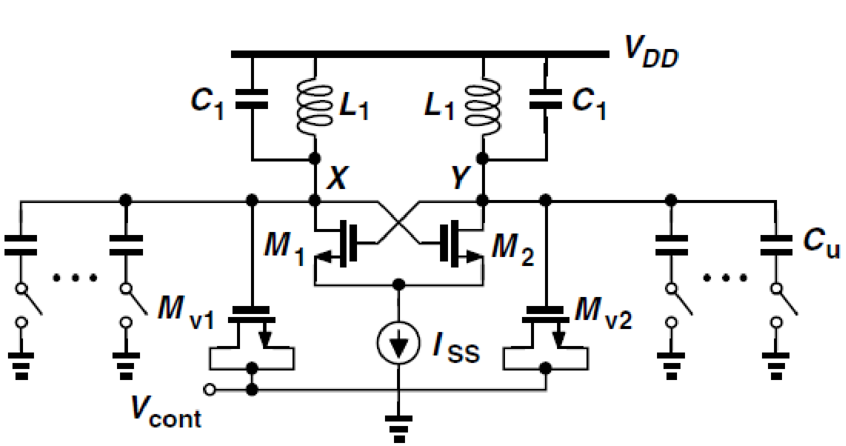
❑The oscillation amplitude is proportional to R_p

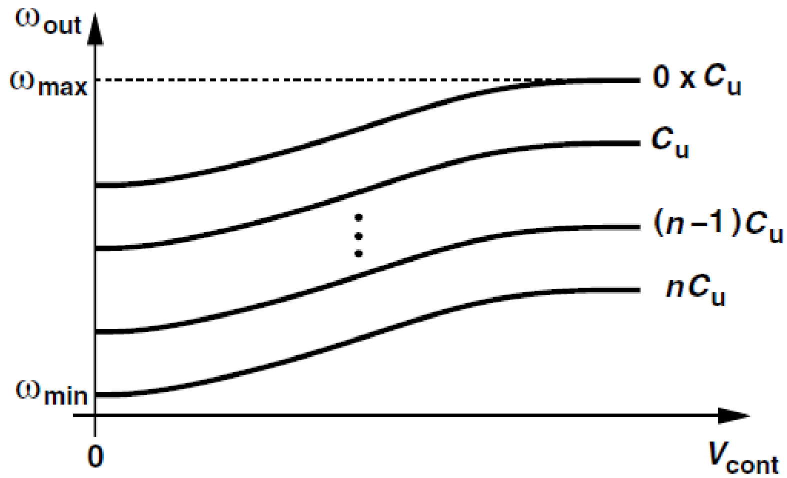
❑For example, a 10% change in ω yields a 20% change in the amplitude

[B. Razavi, RF Microelectronics]

Discrete Tuning

- A small varactor leads to a relatively narrow tuning range
- A large varactor leads to a high VCO gain (K_{VCO}), making the circuit sensitive to noise on the control voltage
- Discrete tuning works as a "coarse control" to enlarge the tuning range, and the varactor works as a "fine control" to decrease the K_{VCO}





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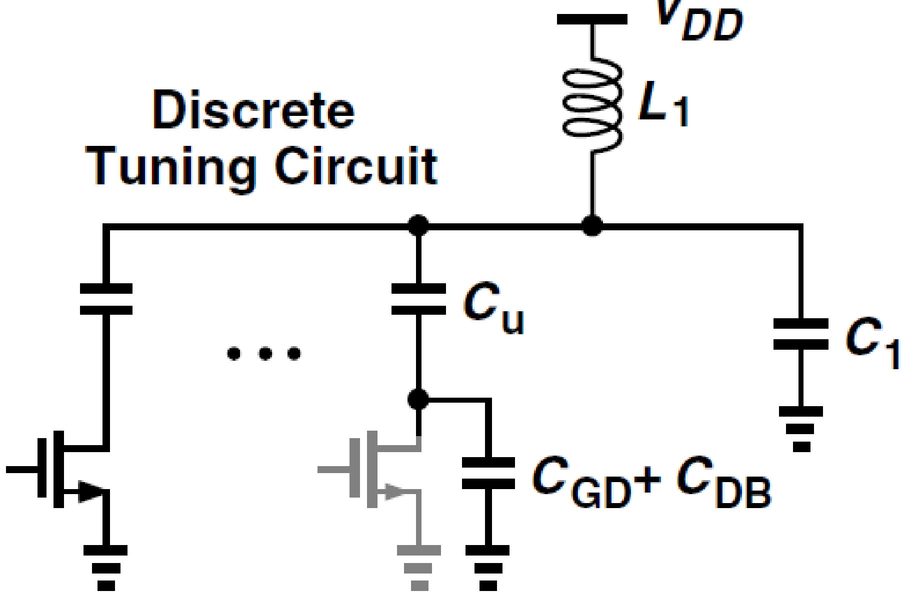
Switch Design Trade-Off

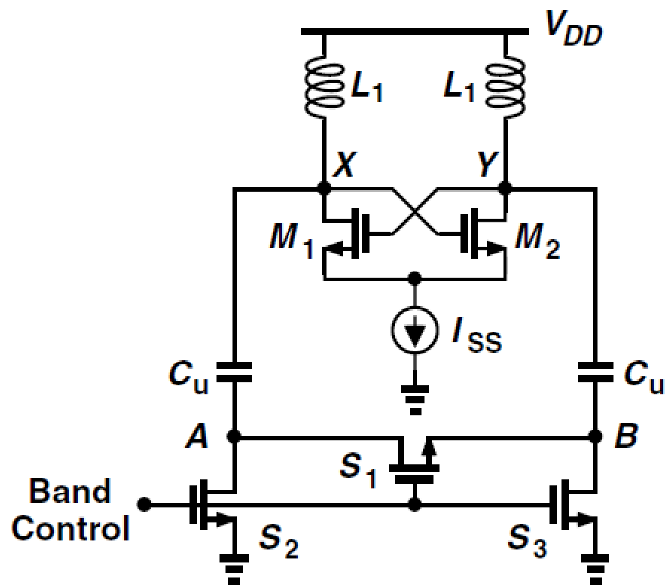
- ❑Increasing the width of the switch transistors help to minimize the tuning-on resistance, increasing the Q
- ❑Unfortunately, wider switches introduce a larger capacitance from the bottom plate of the unit capacitors to ground, thereby presenting a substantial capacitance to the tanks

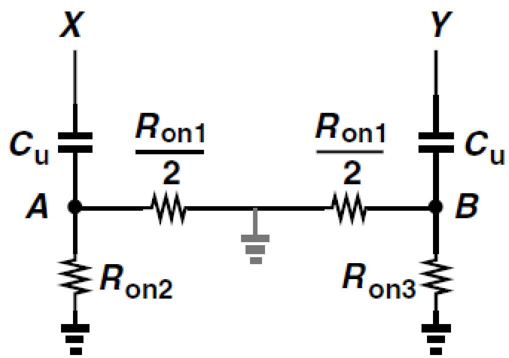


[B. Razavi, RF Microelectronics]

Floating Switch

- ❑The main switch (S_1) is placed between nodes A and B so that, with differential swings at these nodes, only half of R_{on} appears in series with each unit capacitor
- ❑Switches S_2 and S_3 are minimum-size devices, merely defining the CM level of A and B.





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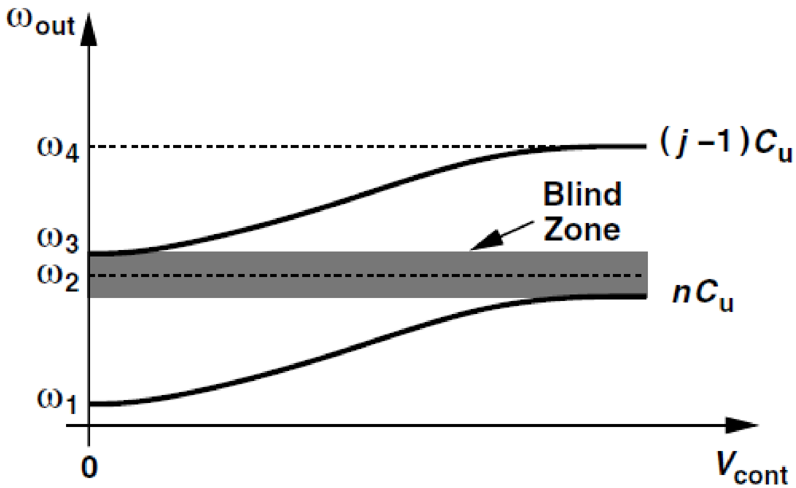
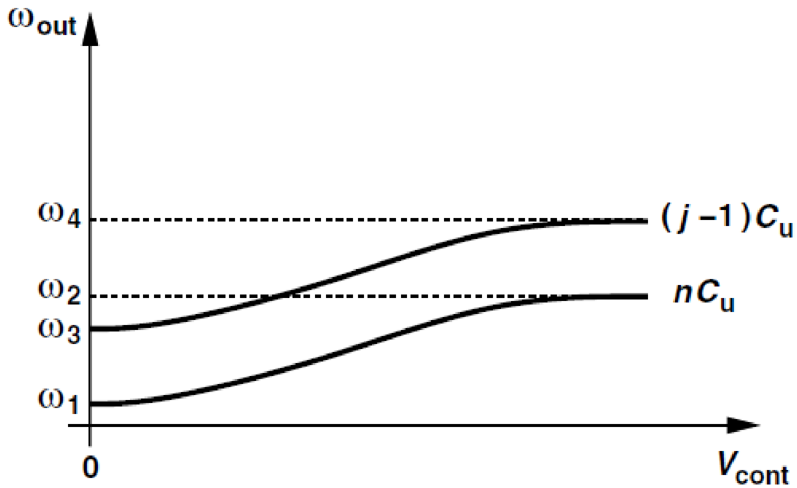
Blind Zone

- Sometimes the oscillator fails to cover the range for any combination of fine and coarse controls
- One can design overlap between consecutive characteristics to avoid blind zone.





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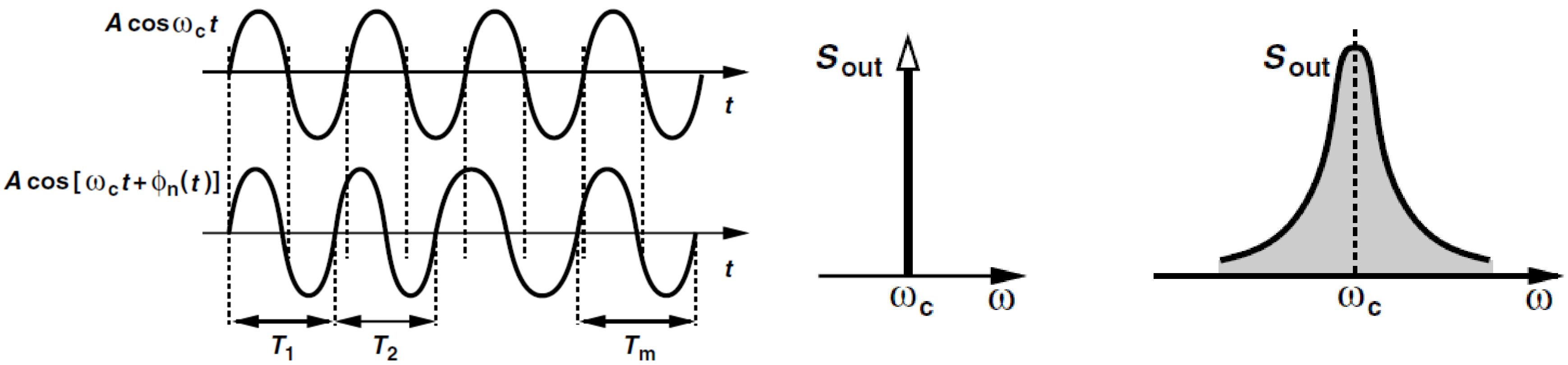
Phase Noise

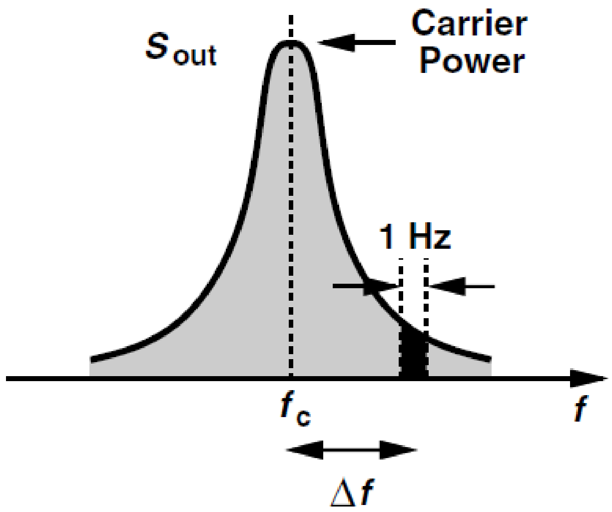
- An oscillator output $x(t)=A\cos[\omega_c t+\varphi_n(t)]$, where $\varphi_n(t)$ is a small random phase quantity that deviates the zero crossings from integer multiples of Tc
- The term $\varphi_n(t)$ is called the "phase noise"



[B. Razavi, RF Microelectronics]

Phase Noise Specification

- Since the phase noise falls at frequencies farther from ω_c, it must be specified at a certain "frequency offset," i.e., a certain difference with respect to ω_c.
- We consider a 1-Hz bandwidth of the spectrum at an offset of f, measure the power in this bandwidth, and normalize the result to the "carrier power."

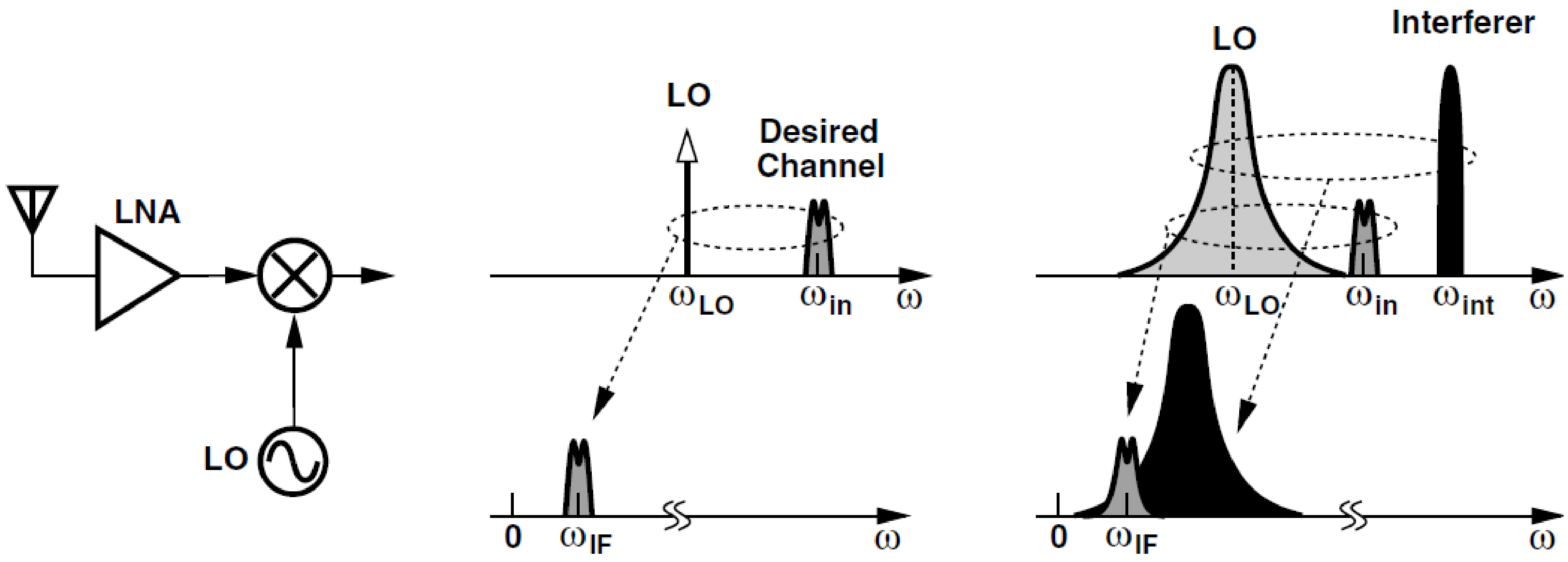


- "dB with respect to the carrier," the unit dBc signifies normalization of the noise power to the carrier power.

[B. Razavi, RF Microelectronics]

Reciprocal Mixing

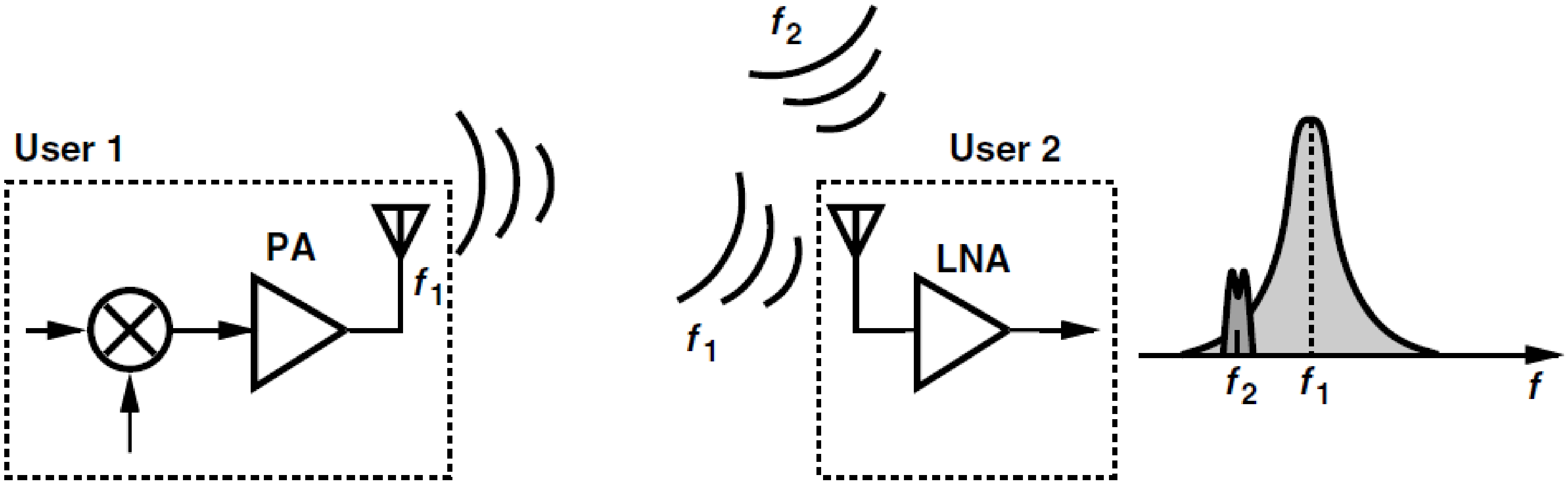
- Suppose the LO suffers from phase noise and the desired signal is accompanied by a large interferer.
- The convolution of the desired signal and the interferer with the noisy LO spectrum results in a broadened downconverted interferer whose noise skirt corrupts the desired IF signal.



[B. Razavi, RF Microelectronics]

Phase Noise in TX

- ❑Two users are located in close proximity, with user#1 transmitting a high-power signal at f_1 and user#2 receiving this signal and a weak signal at f_2
- ❑The phase noise skirt of f_1 received by user #2 greatly corrupts it even before downconversion.

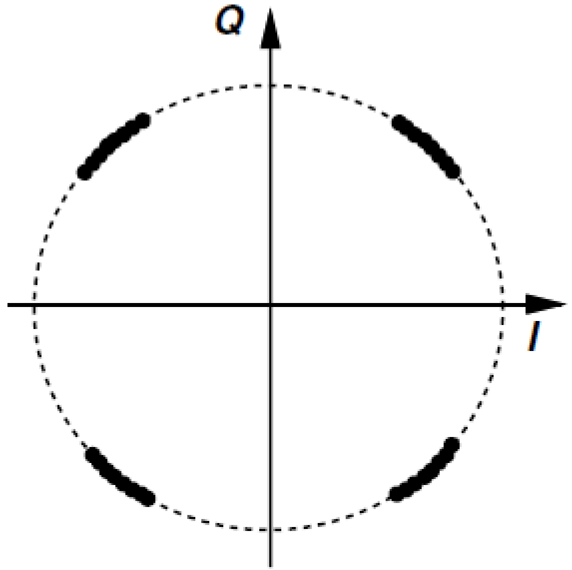


[B. Razavi, RF Microelectronics]

Phase Noise in TX

- The LO phase noise also corrupts phase-modulated signals in the process of upconversion or downconversion
- Since the phase noise is indistinguishable from phase (or frequency) modulation, the mixing of the signal with a noisy LO in the TX or RX path corrupts the signal constellation

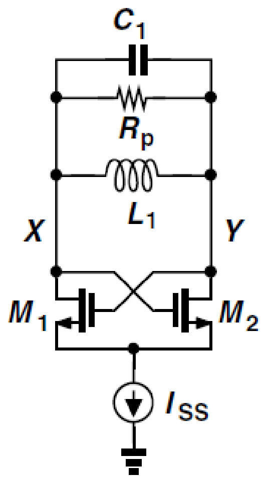
- Corruption of a QPSK signal due to phase noise:

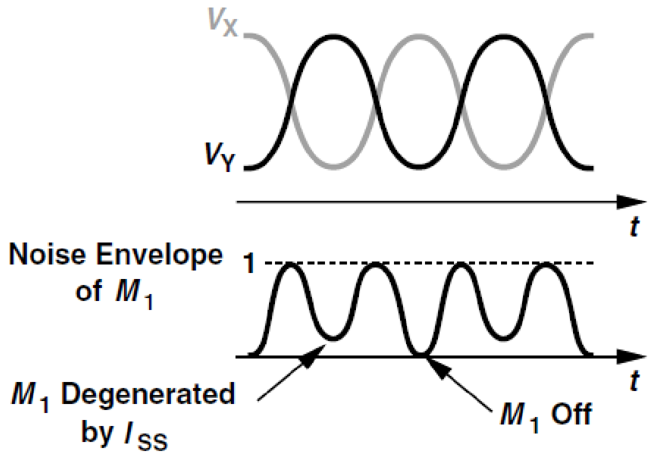


[B. Razavi, RF Microelectronics]

Cyclostationary Noise

- When M_1 turns off, injecting no noise
- When M_1 and M_2 are near equilibrium, they inject maximum noise current
- When M_1 is on but degenerated by the tail current, injecting little noise to the output.
- So, the total noise current experiences an envelope having twice the oscillation frequency

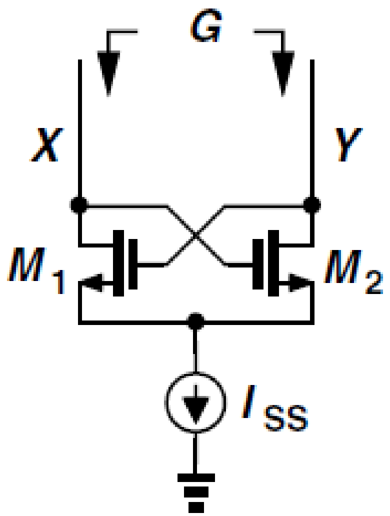


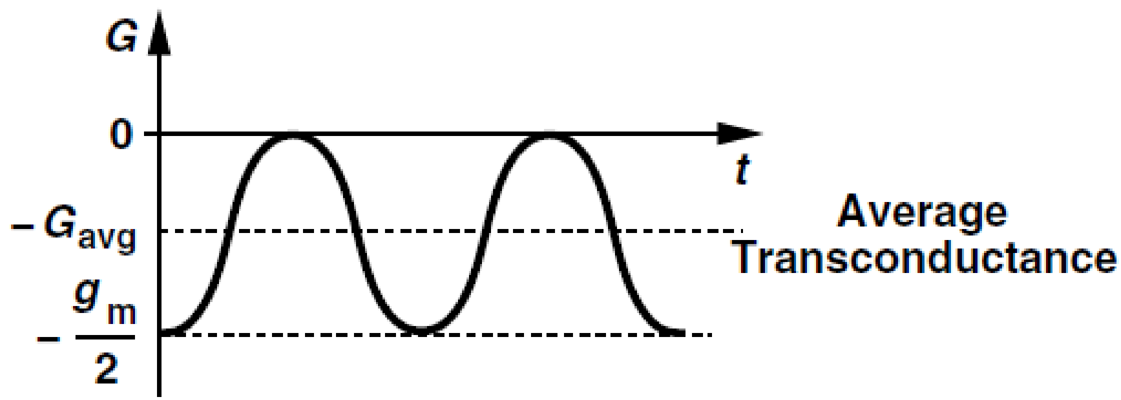


[B. Razavi, RF Microelectronics]

Time-Variation Resistance

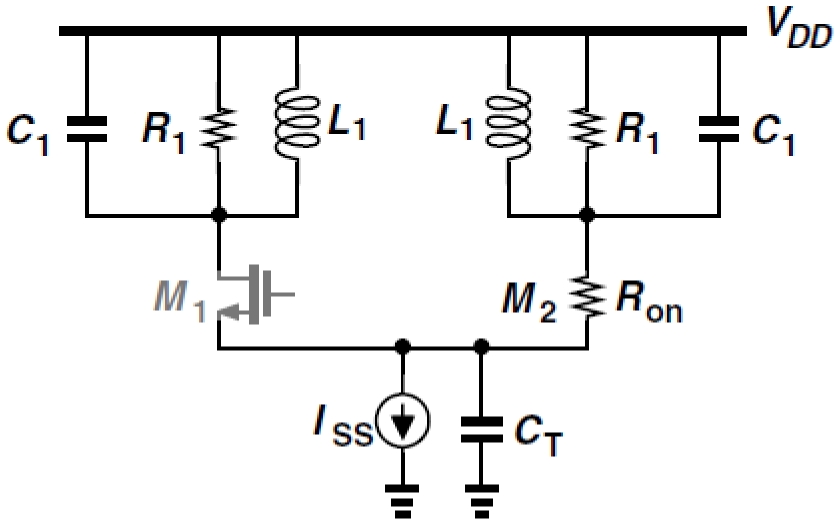
❑For the time variation of the resistance presented by the cross-coupled pair, we may consider a time average of the resistance



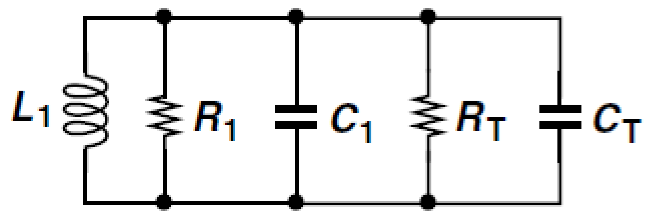


[B. Razavi, RF Microelectronics]

Tail Capacitance



$$R_T = (R_{on} C_T^2 \omega_0^2)^{-1}$$

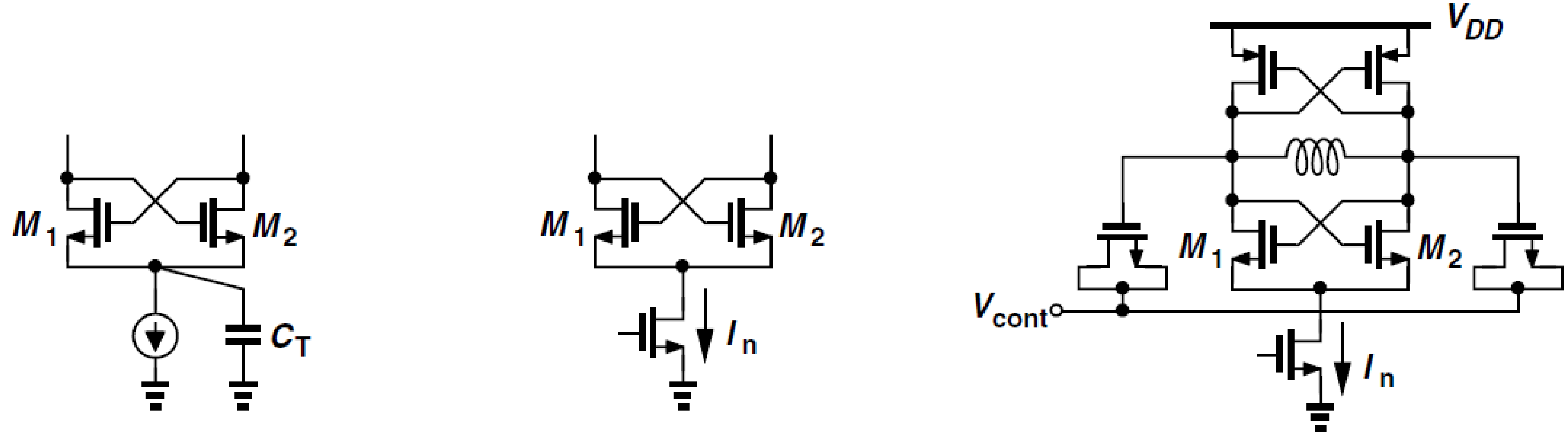


- This capacitance may be large due to the parasitics of I_{SS}
- The Q degrades significantly. Equivalently, the noise injected by M_2 rises considerably
- It may be added deliberately to shunt the noise of I_{SS} to ground

[B. Razavi, RF Microelectronics]

Noise of Bias Current

- ❑Oscillators typically employ a bias current source so as to minimize sensitivity to the supply voltage and noise therein

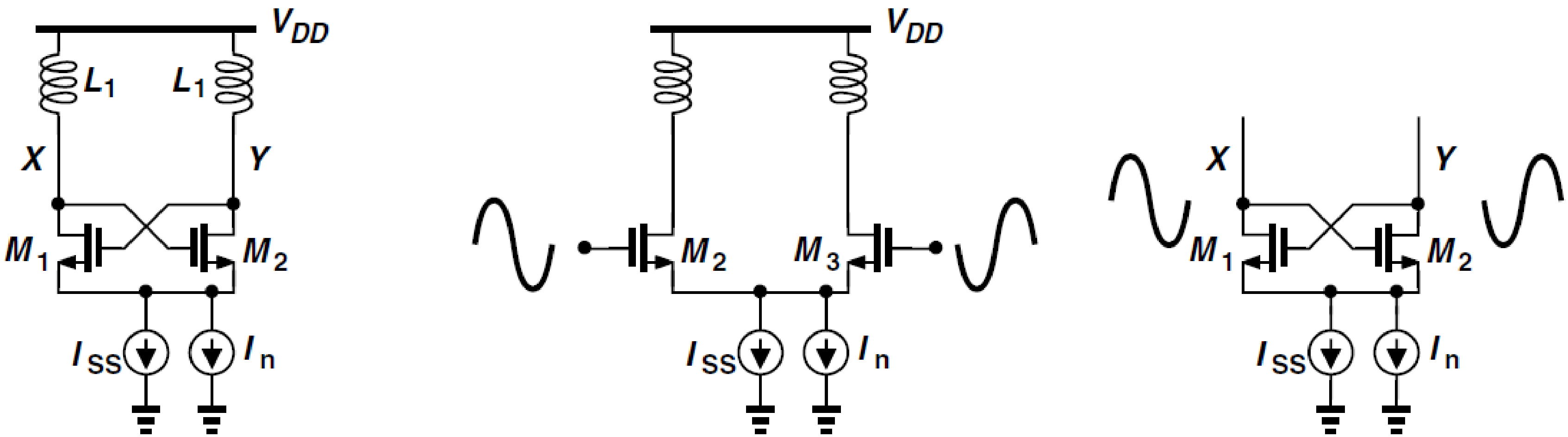


Cross–coupled transistors enter triode region and	Tail current source has noise at $2\omega_0$.	Tail current source has flicker noise and
Tail has large capacitance.		Output CM level is modulated by flicker noise or
		Varactors have even–order voltage dependence.

[B. Razavi, RF Microelectronics]

Noise of Bias Current

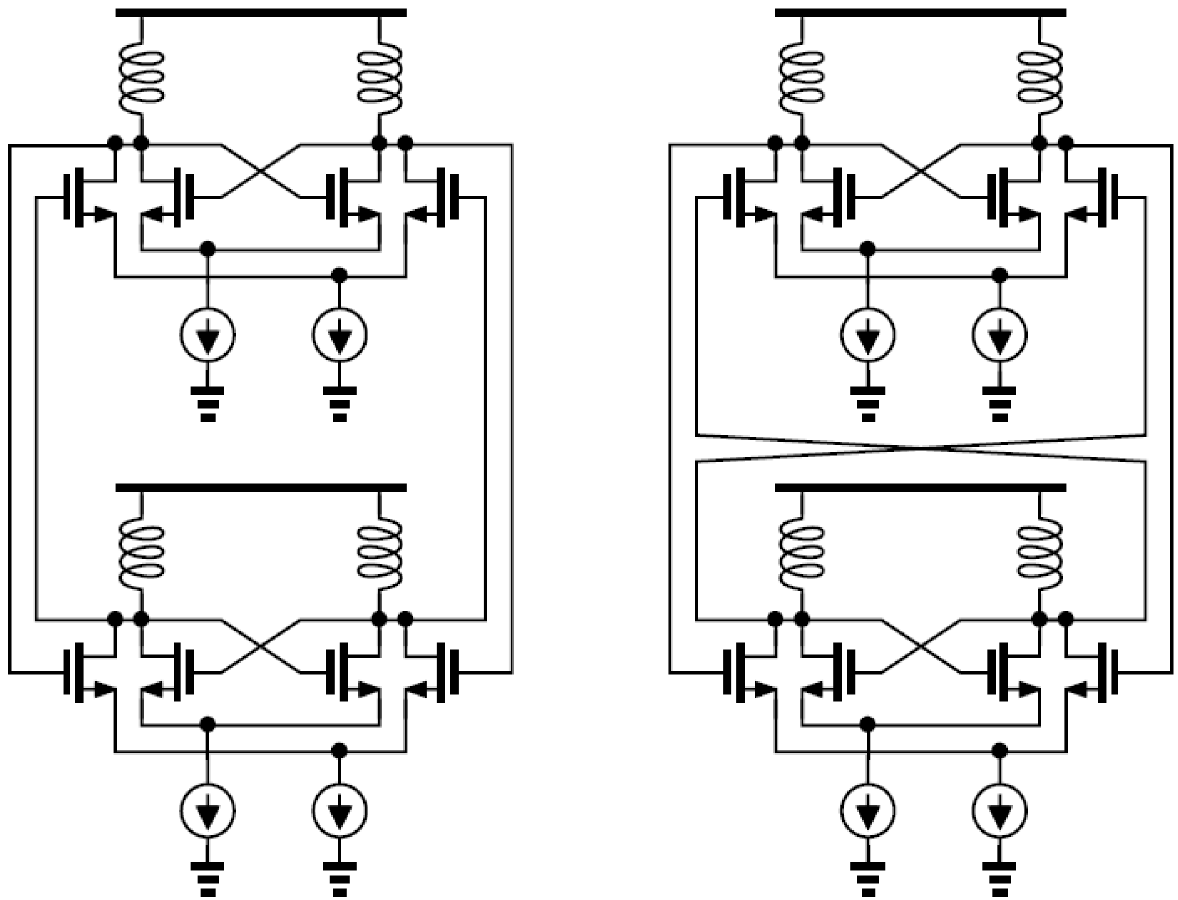
- In models the noise of I_{SS}, including flicker noise near zero frequency, thermal noise around the oscillation frequency, ω_0, and thermal noise around $2\omega_0$
- M_1 and M_2 are periodically turned on and off, thus steering I_{SS} I_n and hence operating as a mixer



[B. Razavi, RF Microelectronics]

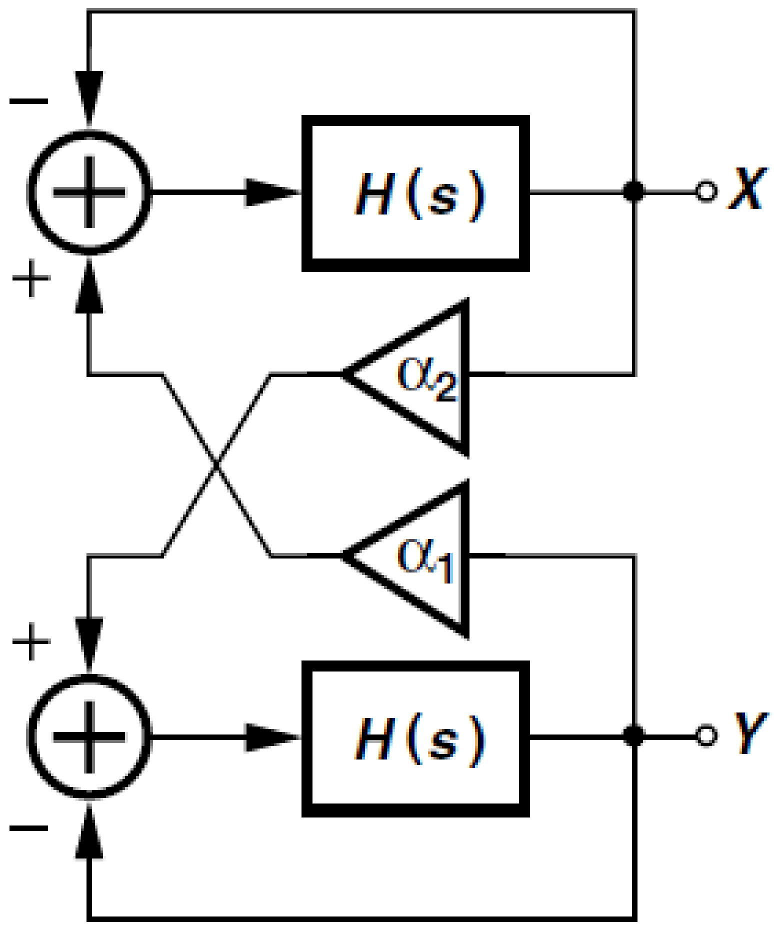
Coupling Oscillators

❑**Two oscillators can be coupled in phase or anti-phase**



[B. Razavi, RF Microelectronics]

Coupling Oscillators



$$X = (\alpha_1 Y - X)H(s)$$

$$Y = (\alpha_2 X - Y)H(s)$$

$$\left(\alpha_2 X^2 - \alpha_1 Y^2\right)[1 + H(s)] = 0$$

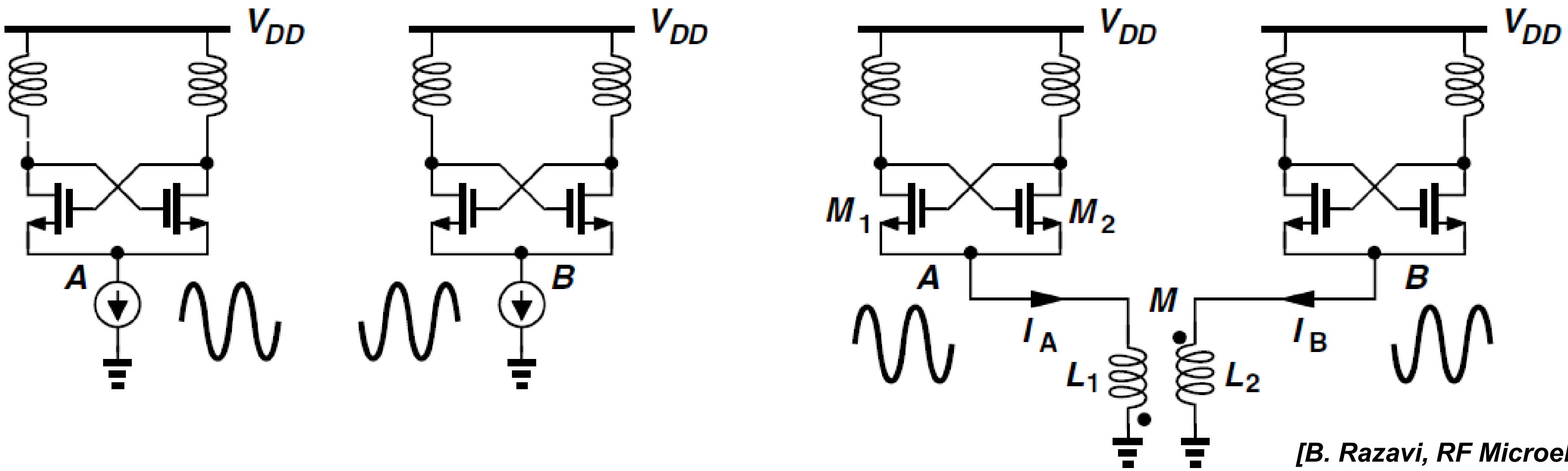
$$\alpha_2 X^2 = \alpha_1 Y^2$$

- If α_1=α_2 (in-phase coupling), then X=±Y, i.e., the two oscillators operate with a zero or 180° phase difference.

[B. Razavi, RF Microelectronics]

Quadrature Oscillators

- Two oscillators are somehow forced to operate in quadrature at a frequency of ω_{osc}
- The tail nodes, A and B, thus exhibit periodic waveforms at $2\omega_{osc}$ and 180^o out of phase.
- Conversely, if additional circuitry forces A and B to sustain a phase difference of 180^o, then the two oscillators operate in quadrature

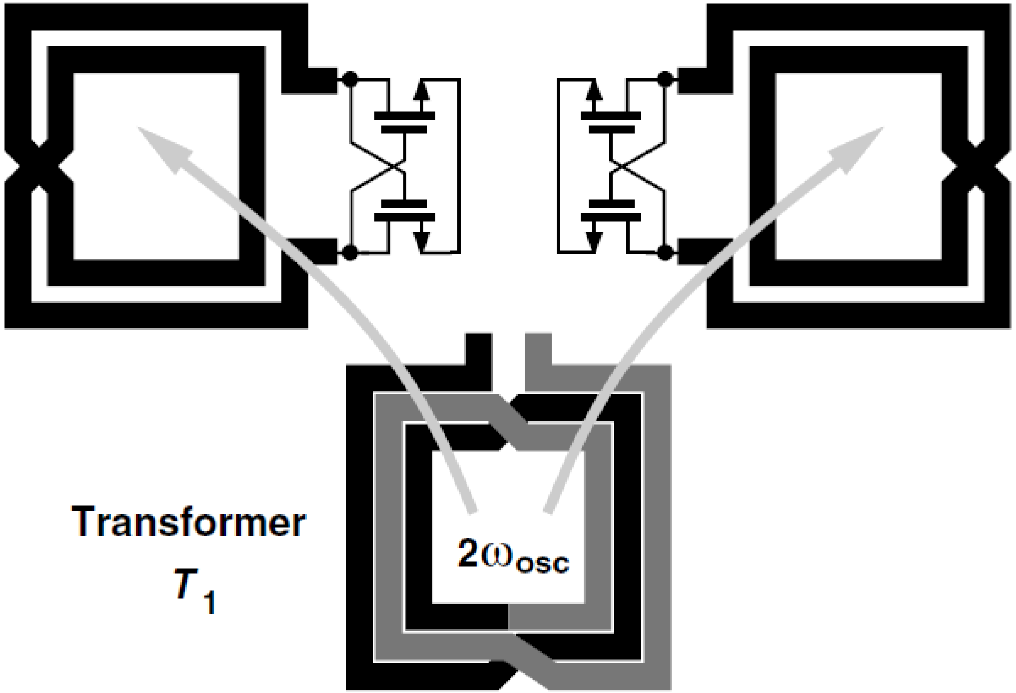


[B. Razavi, RF Microelectronics]

Quadrature Oscillators

❑The tail transformer T_1 must remain relatively far from the main inductors so as to minimize the leakage of $2\omega_{osc}$ to the two core oscillators.

❑Such leakage distorts the duty cycle of the outputs



[B. Razavi, RF Microelectronics]