CIRF
Circuit Intégré Radio Fréquence

Lecture II

- *Introduction*
- *Negative Resistance Oscillators*
- *Integrated Passive Components*
- *Phase Noise in Local Oscillators*

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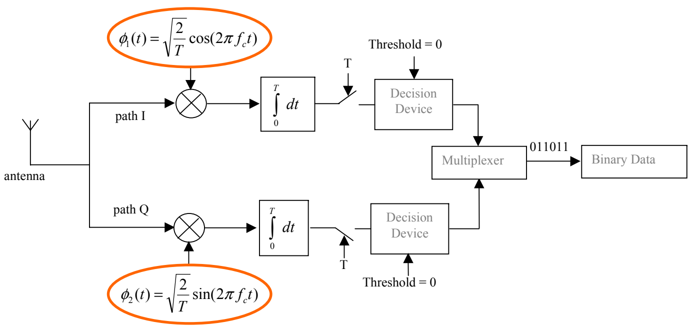
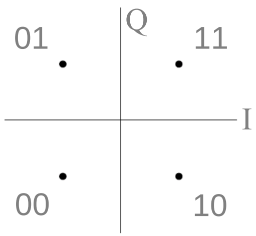
QPSK Receiver



QPSK Constellation Diagram



References

- M. Perrott, "High Speed Communication Circuits and Systems", M.I.T. OpenCourseWare, http://ocw.mit.edu/, Massachusetts Institute of Technology, 2003.
- T. Lee, "The Design of CMOS Radio-Frequency Integrated Circuits", Cambridge University Press, 2004.
- B. Razavi, "Design of Analog CMOS Integrated Circuits", Mc Graw-Hill, 2001.

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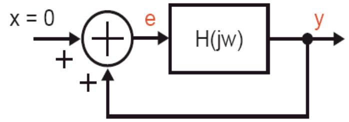
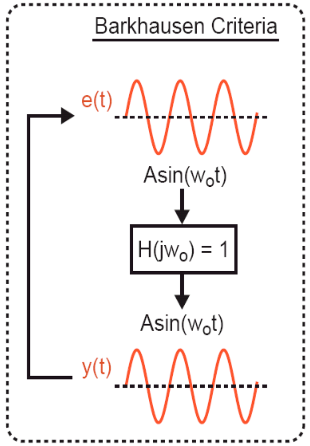
Barkhausen's Criteria for Oscillation





- Closed loop transfer function

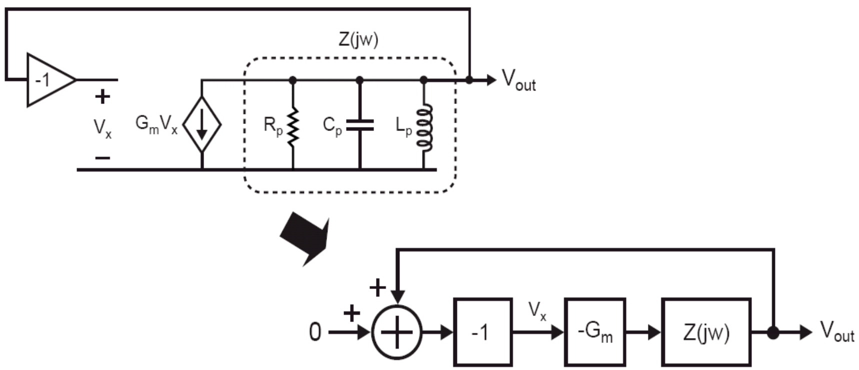
$$G(jw) = \frac{Y(jw)}{X(jw)} = \frac{H(jw)}{1 - H(jw)}$$

- Self-sustaining oscillation at frequency w_o if

$$H(jw_o) = 1$$

  - Amounts to two conditions:
    - Gain = 1 at frequency w_o
    - Phase = n360 degrees (n = 0,1,2,…) at frequency w_o

Resonator-Based Oscillator

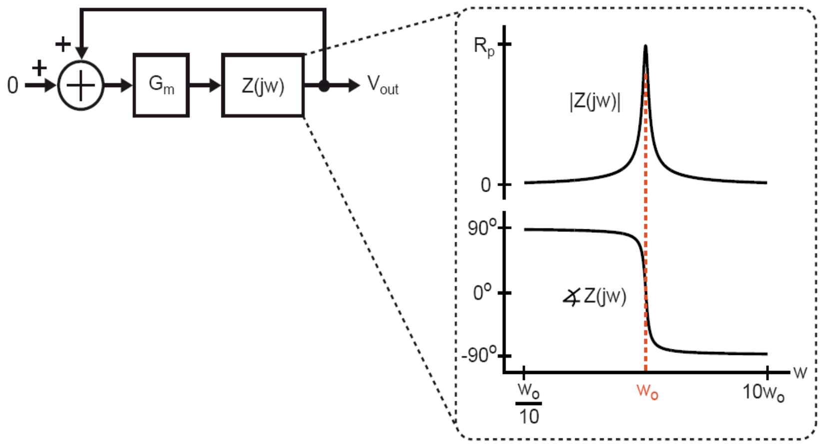


- Barkhausen Criteria for oscillation at frequency w_o:

$$G_m Z(jw_o) = 1$$

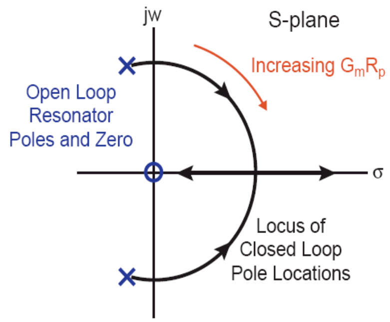
  - Assuming G_m is purely real, $Z(jw_o)$ must also be purely real

A Closer Look At Resonator-Based Oscillator



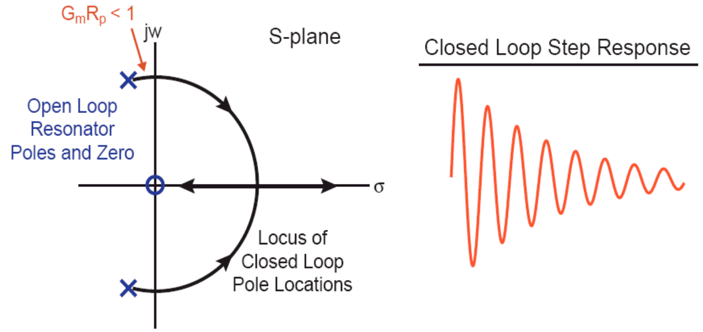
- For parallel resonator at resonance
  - Looks like resistor (i.e., purely real) at resonance
    - Phase condition is satisfied
    - Magnitude condition achieved by setting $G_m R_p = 1$

Impact of Different G_m Values



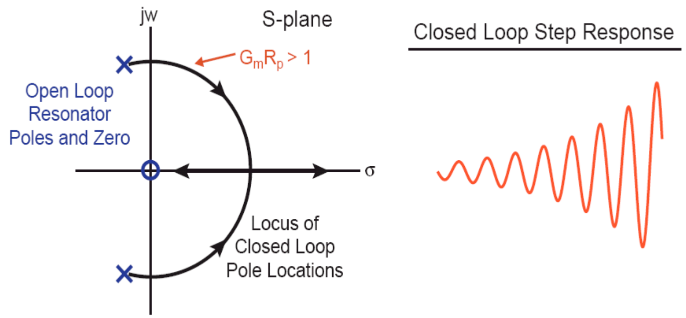
- Root locus plot allows us to view closed loop pole locations as a function of open loop poles/zero and open loop gain (G_mR_p)
  - As gain (G_mR_p) increases, closed loop poles move into right half S-plane

Impact of Setting G_m too low



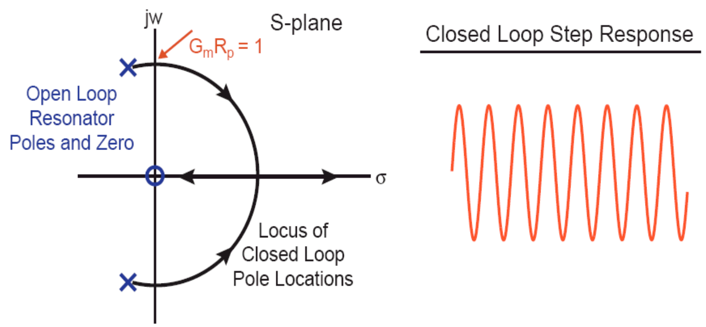
- Closed loop poles end up in the left half S-plane
  - Underdamped response occurs
    - Oscillation dies out

Impact of Setting G_m too High



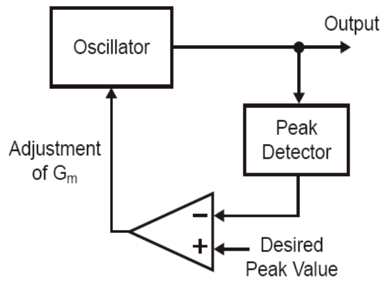
- Closed loop poles end up in the right half S-plane
  - Unstable response occurs
    - Waveform blows up!

Setting G_m To Just the Right Value



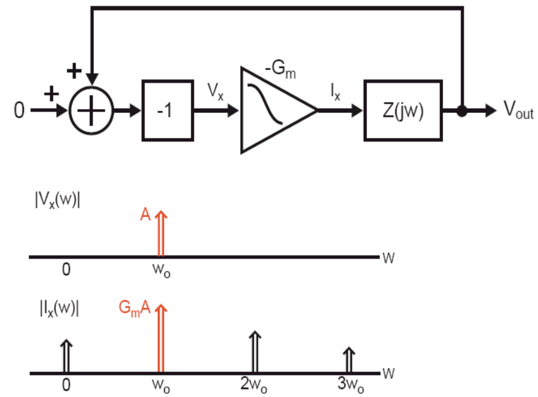
- Closed loop poles end up on jw axis
  - Oscillation maintained
- Issue – G_mR_p needs to *exactly* equal 1
  - How do we achieve this in practice?

Amplitude Feedback Loop



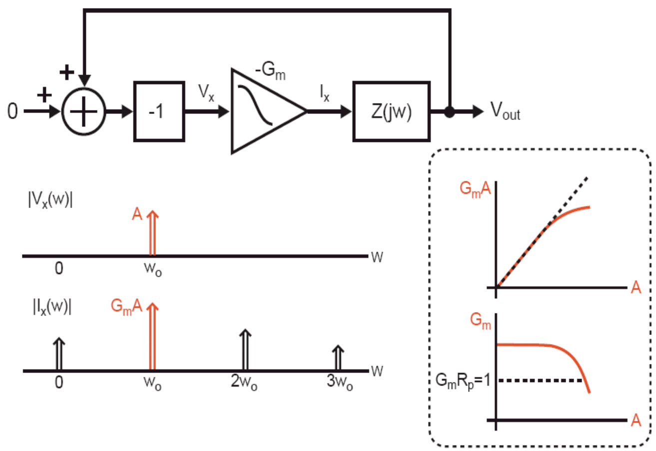
- One thought is to detect oscillator amplitude, and then adjust G_m so that it equals a desired value
  - By using feedback, we can precisely achieve $G_mR_p = 1$
- Issues
  - Complex, requires power, and adds noise

Leveraging Amplifier Nonlinearity as Feedback



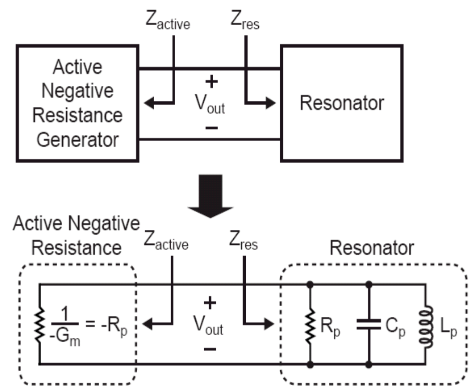
- Practical transconductance amplifiers have saturating characteristics
  - Harmonics created, but filtered out by resonator
  - Our interest is in the relationship between the input and the fundamental of the output

Leveraging Amplifier Nonlinearity as Feedback



- As input amplitude is increased
  - Effective gain from input to fundamental of output drops
  - Amplitude feedback occurs! ($G_mR_p = 1$ in steady-state)

One-Port View of Resonator-Based Oscillators

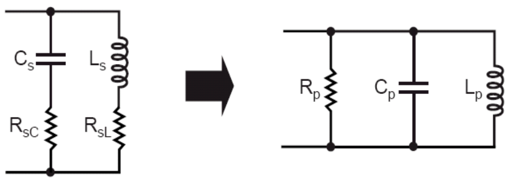


- Convenient for intuitive analysis
- Here we seek to cancel out loss in tank with a negative resistance element
  - To achieve sustained oscillation, we must have

$$\frac{1}{G_m} = R_p \quad \Rightarrow \quad G_m R_p = 1$$

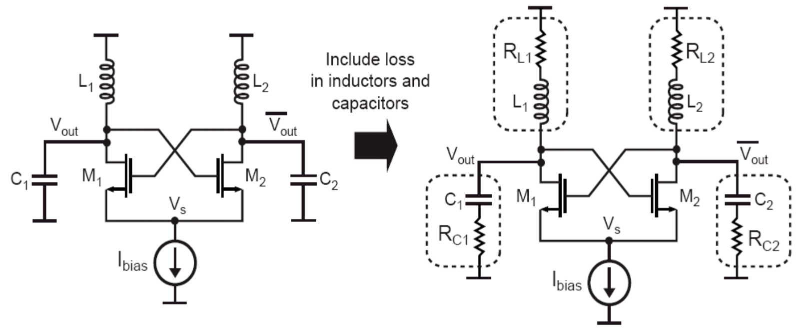
One-Port Modeling Requires Parallel RLC Network

- **Since VCO operates over a very narrow band of frequencies, we can always do series to parallel transformations to achieve a parallel network for analysis**



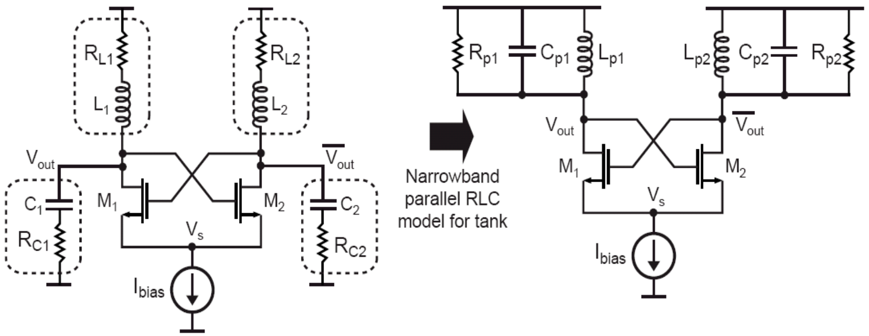
  - **Warning – in practice, RLC networks can have secondary (or more) resonant frequencies, which cause undesirable behavior**
    - Equivalent parallel network masks this problem in hand analysis
    - Simulation will reveal the problem

Example – Negative Resistance Oscillator



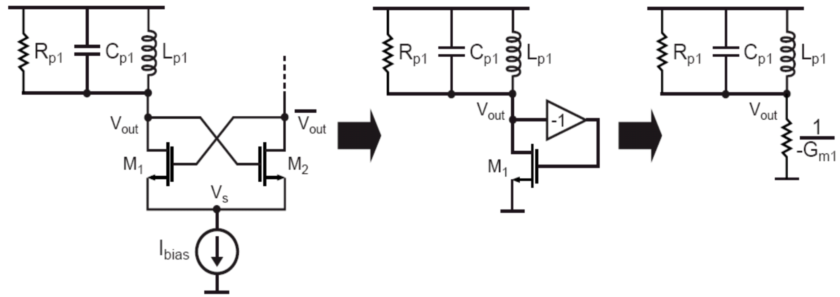
- **This type of oscillator structure is quite popular in current CMOS implementations**
  - **Advantages**
    - Simple topology
    - Differential implementation (good for feeding differential circuits)
    - Good phase noise performance can be achieved

Analysis of Negative Resistance Oscillator (Step 1)



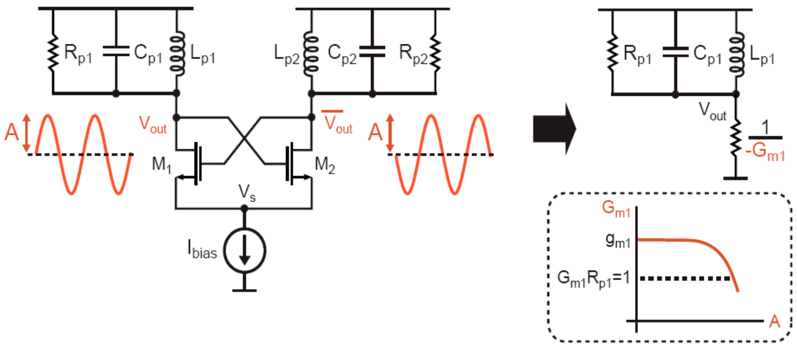
- **Derive a parallel RLC network that includes the loss of the tank inductor and capacitor**
  - **Typically, such loss is dominated by series resistance in the inductor**

Analysis of Negative Resistance Oscillator (Step 2)



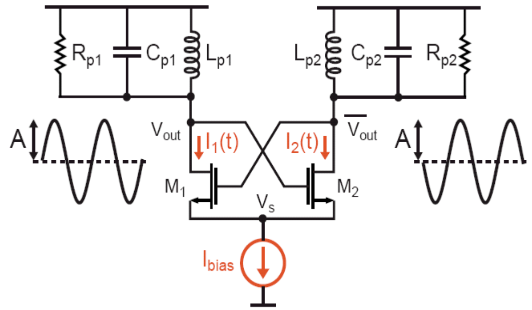
- **Split oscillator circuit into half circuits to simplify analysis**
  - **Leverages the fact that we can approximate V_s as being incremental ground (this is not quite true, but close enough)**
- **Recognize that we have a diode connected device with a negative transconductance value**
  - **Replace with negative resistor**
    - Note: G_m is ***large signal*** transconductance value

Design of Negative Resistance Oscillator



- Design tank components to achieve high Q
  - Resulting R_p value is as large as possible
- Choose bias current (I_{bias}) for large swing (without going far into saturation)
  - We'll estimate swing as a function of I_{bias} shortly
- Choose transistor size to achieve adequately large g_{m1}
  - Usually twice as large as $1/R_{p1}$ to guarantee startup

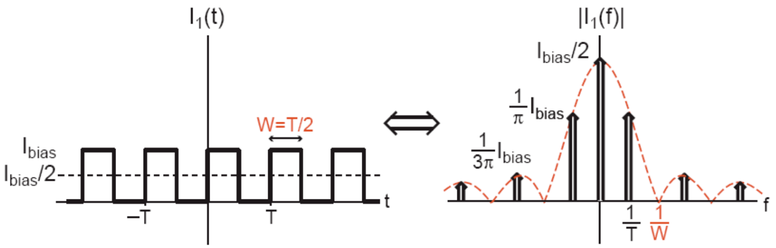
Calculation of Oscillator Swing



- Design tank components to achieve high Q
  - Resulting R_p value is as large as possible
- Choose bias current (I_{bias}) for large swing (without going far into saturation)
  - We'll estimate swing as a function of I_{bias} in next slide
- Choose transistor size to achieve adequately large g_{m1}
  - Usually twice as large as $1/R_{p1}$ to guarantee startup

Calculation of Oscillator Swing as a Function of I_{bias}

- By symmetry, assume $I_1(t)$ is a square wave
  - We are interested in determining fundamental component
    - (DC and harmonics filtered by tank)



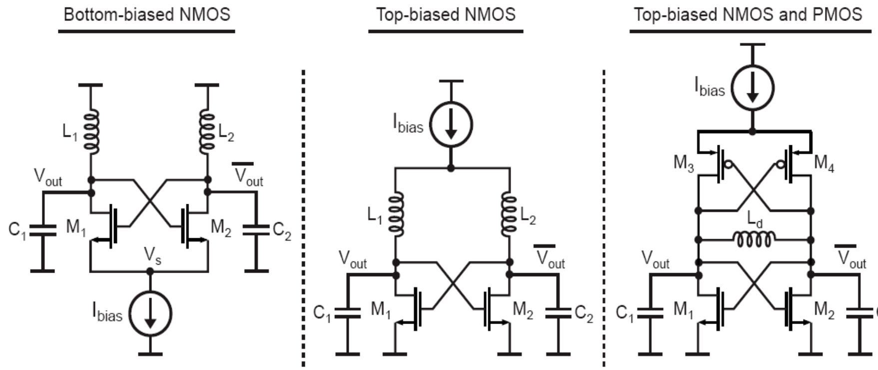
  - Fundamental component is

$$I_1(t)\Big|_{fundamental} = \frac{2}{\pi} I_{bias} \sin(w_o t), \quad \text{where} \quad w_o = \frac{2\pi}{T}$$

  - Resulting oscillator amplitude

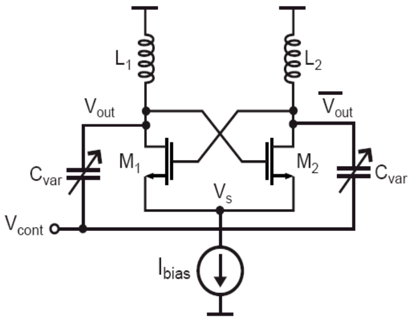
$$A = \frac{2}{\pi} I_{bias} R_p$$

Variations on a Theme



- Biasing can come from top or bottom
- Can use either NMOS, PMOS, or both for transconductor
  - Use of both NMOS and PMOS for coupled pair would appear to achieve better phase noise at a given power dissipation
    - See Hajimiri et. al, "Design Issues in CMOS Differential LC Oscillators", JSSC, May 1999 and Feb, 2000 (pp 286-287)

Voltage Controlled Oscillators (VCO's)



- Include a tuning element to adjust oscillation frequency
  - Typically use a variable capacitor (varactor)
- Varactor incorporated by replacing fixed capacitance
  - Note that much fixed capacitance cannot be removed (transistor junctions, interconnect, etc.)
    - Fixed cap lowers frequency tuning range

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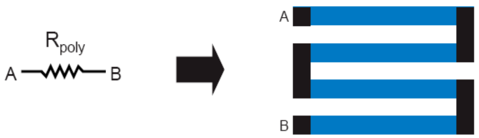
Polysilicon Resistors

- Use unsilicided polysilicon to create resistor



- Key parameters
  - Resistance (usually 100- 200 Ohms per square)
  - Parasitic capacitance (usually small)
    - Appropriate for high speed amplifiers
  - Linearity (quite linear compared to other options)
  - Accuracy (usually can be set within ± 15%)

MOS Resistors

- Bias a MOS device in its triode region

$$R_{ds} \approx \frac{1}{\mu C_{ox} W/L((V_{gs} - V_T) - V_{DS})}$$

- High resistance values can be achieved in a small area (MegaOhms within tens of square microns)
- Resistance is quite nonlinear
  - Appropriate for small swing circuits

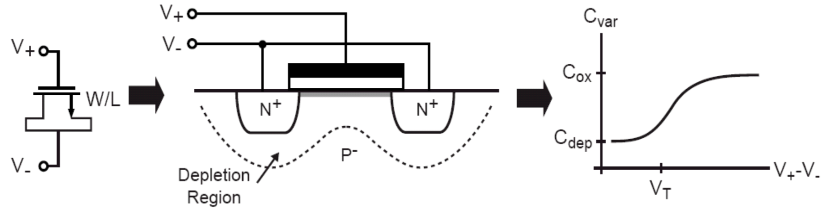
High Density Capacitors (Biasing, Decoupling)

- MOS devices offer the highest capacitance per unit area
  - Limited to a one terminal device
  - Voltage must be high enough to invert the channel

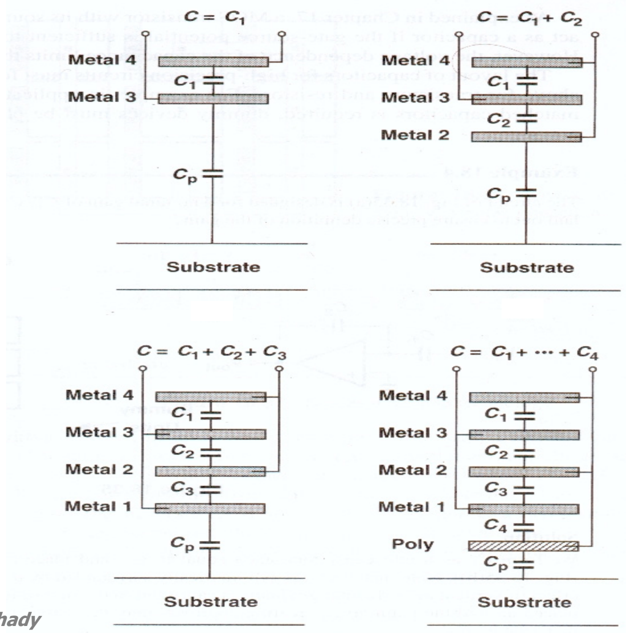
- Key parameters
  - Capacitance value
    - Raw cap value from MOS device is 6.1 fF/μ m² for 0.24u CMOS
  - Q (i.e., amount of series resistance)
    - Maximized with minimum L (tradeoff with area efficiency)

A Recently Popular Approach – The MOS Varactor

- Consists of a MOS transistor (NMOS or PMOS) with drain and source connected together
  - Abrupt shift in capacitance as inversion channel forms
- Advantage – easily integrated in CMOS
- Disadvantage – Q is relatively low in the transition region
  - Note that large signal is applied to varactor – transition region will be swept across each VCO cycle

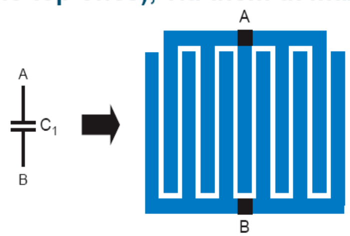


Vertical Metal Capacitors



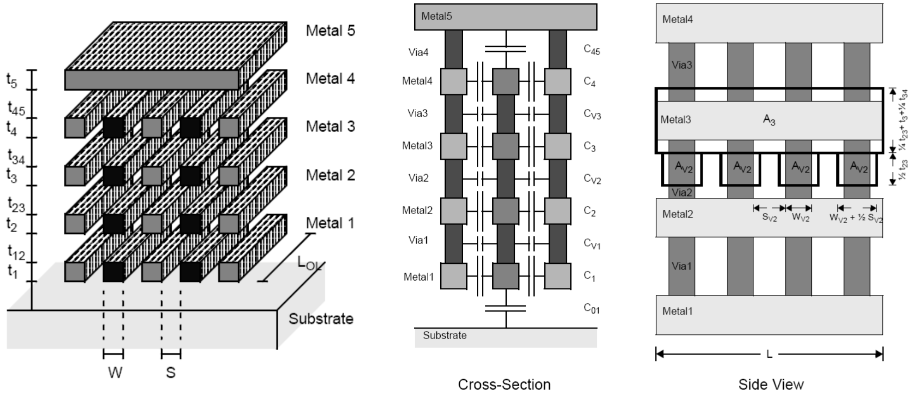
Lateral Metal Capacitors

- Lateral metal capacitors offer high Q and reasonably large capacitance per unit area
  - Stack many levels of metal on top of each other (best layers are the top ones), via them at maximum density



  - Accuracy often better than ±10%
  - Parasitic side cap is symmetric, less than 10% of cap value
- Example: C_T = 1.5 fF/μm² for 0.24μm process with 7 metals, L_{min} = W_{min} = 0.24μm, t_{metal} = 0.53μm
  - See "Capacity Limits and Matching Properties of Integrated Capacitors", Aparicio et. al., JSSC, Mar 2002

Vertical Mesh Metal Capacitors



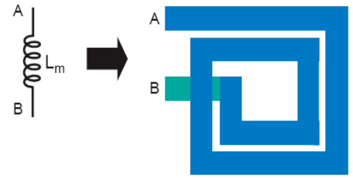
$$C_{mesh} = 2(C_1 + C_{V1} + C_2 + C_{V2} + C_3 + C_{V3} + C_4) + C_{45}$$

$$C_{par} = C_{01}$$

K.T. Christensen,"Low Power RF Filtering for CMOS Transceivers", Ph.D. Denmark Technical University, 2001, http://phd.dtv.dk/2001/oersted/k_t_christensen.pdf

Spiral Inductors

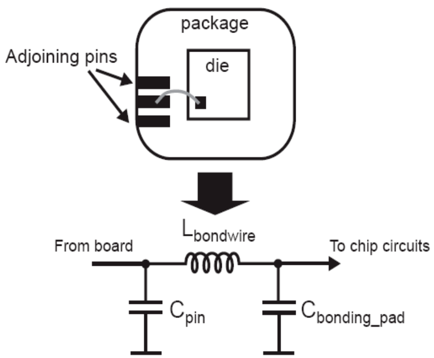
- Create integrated inductor using spiral shape on top level metals (may also want a patterned ground shield)



  - Key parameters are Q (< 10), L (1-10 nH), self resonant freq.
  - Usually implemented in top metal layers to minimize series resistance, coupling to substrate
  - Design using Mohan et. al, “Simple, Accurate Expressions for Planar Spiral Inductances, JSSC, Oct, 1999, pp 1419-1424
  - Verify inductor parameters (L, Q, etc.) using ASITIC http://formosa.eecs.berkeley.edu/~niknejad/asitic.html

Bondwire Inductors

- Used to bond from the package to die
  - Can be used to advantage



- Key parameters
  - Inductance ($\approx$ 1 nH/mm – usually achieve 1-5 nH)
  - Q (much higher than spiral inductors – typically > 40)

Other Types of Resonators

- Quartz crystal
  - Very high Q, and very accurate and stable resonant frequency
    - Confined to low frequencies (< 200 MHz)
    - Non-integrated
  - Used to create low noise, accurate, “reference” oscillators
- SAW devices
  - High frequency, but poor accuracy (for resonant frequency)
- MEMS devices
  - Cantilever beams – promise high Q, but non-tunable and haven’t made it to the GHz range, yet, for resonant frequency
  - FBAR – Q > 1000, but non-tunable and poor accuracy
  - Other devices are on the way!

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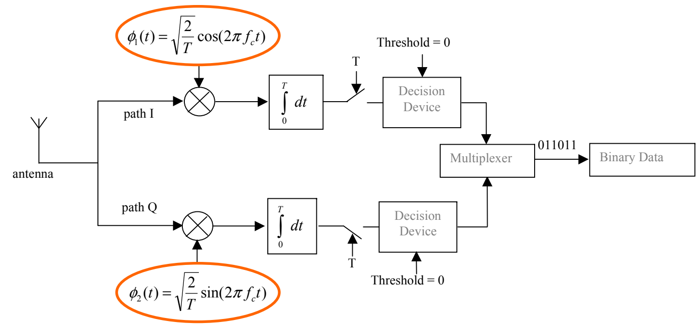
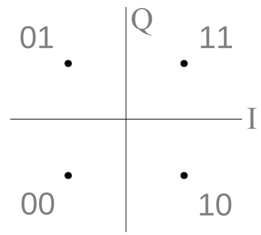
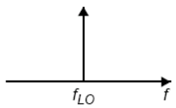
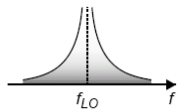


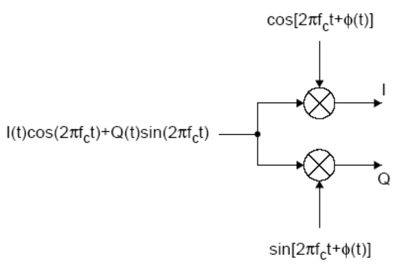
QPSK Constellation Diagram

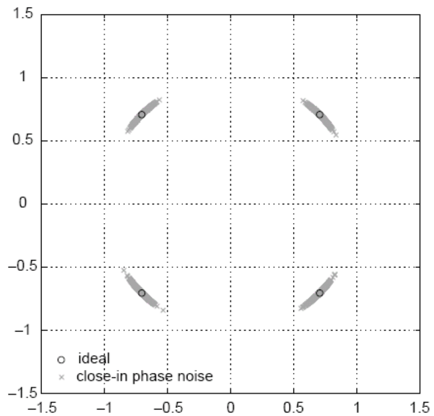


Local Oscillator Phase Noise

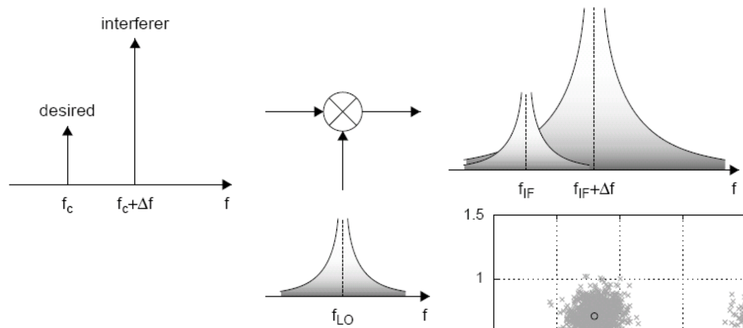


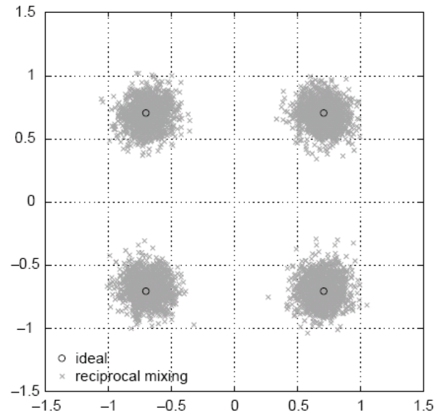




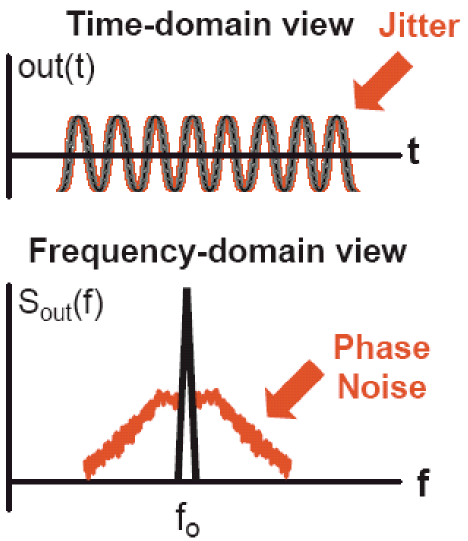


Reciprocal Mixing

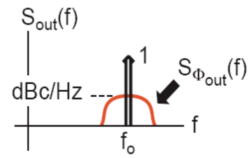




Phase Noise in Oscillators



Measurement of Phase Noise in dBc/Hz



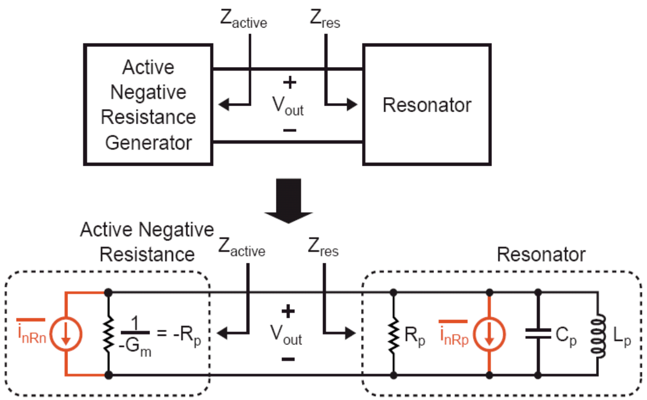
- **Definition of L(f)**

$$L(f) = 10\log\left(\frac{\text{Spectral density of noise}}{\text{Power of carrier}}\right)$$

  - Units are dBc/Hz
- **For this case**

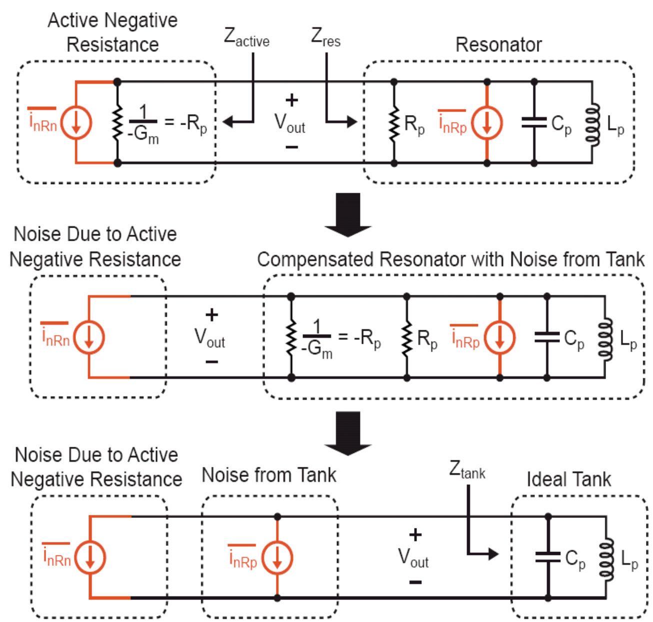
$$L(f) = 10\log\left(\frac{S_{\Phi_{out}}(f)}{1}\right) = 10\log(S_{\Phi_{out}}(f))$$

Calculation of Intrinsic Phase Noise in Oscillators

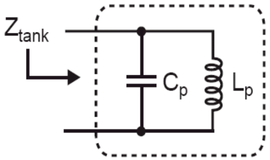


- Noise sources in oscillators are put in two categories
  - Noise due to tank loss
  - Noise due to active negative resistance
- We want to determine how these noise sources influence the phase noise of the oscillator

Equivalent Model for Noise Calculations



Calculate Impedance Across Ideal LC Tank Circuit



$$Z_{tank}(w) = \frac{1}{jwC_p}||jwL_p = \frac{jwL_p}{1 - w^2L_pC_p}$$

- Calculate input impedance about resonance

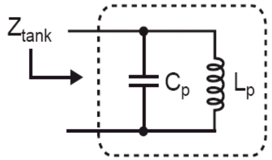
$$\text{Consider } w = w_o + \Delta w, \text{ where } w_o = \frac{1}{\sqrt{L_pC_p}}$$

$$Z_{tank}(\Delta w) = \frac{j(w_o + \Delta w)L_p}{1 - (w_o + \Delta w)^2 L_pC_p}$$

$$= \frac{j(w_o + \Delta w)L_p}{\underbrace{1 - w_o^2L_pC_p}_{=0} - 2\Delta w(w_oL_pC_p) - \underbrace{\Delta w^2L_pC_p}_{\text{negligible}}} \approx \frac{j(w_o + \Delta w)L_p}{-2\Delta w(w_oL_pC_p)}$$

$$\Rightarrow Z_{tank}(\Delta w) \approx \frac{jw_oL_p}{-2\Delta w(w_oL_pC_p)} = \boxed{-\frac{j}{2}\frac{1}{w_oC_p}\left(\frac{w_o}{\Delta w}\right)}$$

A Convenient Parameterization of LC Tank Impedance



$$Z_{tank}(\Delta w) \approx -\frac{j}{2}\frac{1}{w_oC_p}\left(\frac{w_o}{\Delta w}\right)$$

- Actual tank has loss that is modeled with R_p
  - Define Q according to actual tank

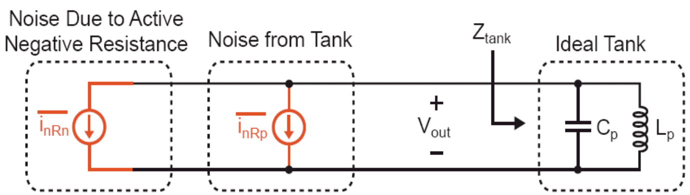
$$Q = R_pw_oC_p \Rightarrow \frac{1}{w_oC_p} = \frac{R_p}{Q}$$

- Parameterize ideal tank impedance in terms of Q of actual tank

$$Z_{tank}(\Delta w) \approx -\frac{j}{2}\frac{R_p}{Q}\left(\frac{w_o}{\Delta w}\right)$$

$$\Rightarrow |Z_{tank}(\Delta f)|^2 \approx \boxed{\left(\frac{R_p}{2Q}\frac{f_o}{\Delta f}\right)^2}$$

Overall Noise Output Spectral Density



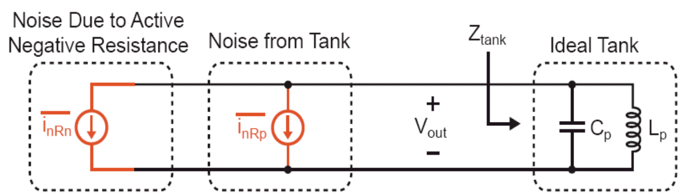
- Assume noise from active negative resistance element and tank are uncorrelated

$$\frac{\overline{v_{out}^2}}{\Delta f} = \left(\frac{\overline{i_{nRp}^2}}{\Delta f} + \frac{\overline{i_{nRn}^2}}{\Delta f}\right)|Z_{tank}(\Delta f)|^2$$

$$= \frac{\overline{i_{nRp}^2}}{\Delta f}\left(1 + \frac{\overline{i_{nRn}^2}}{\Delta f} \Big/ \frac{\overline{i_{nRp}^2}}{\Delta f}\right)|Z_{tank}(\Delta f)|^2$$

  - Note that the above expression represents total noise that impacts both amplitude and phase of oscillator output

Parameterize Noise Output Spectral Density



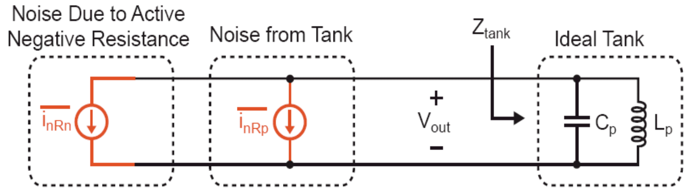
- From previous slide

$$\frac{\overline{v_{out}^2}}{\Delta f} = \frac{\overline{i_{nRp}^2}}{\Delta f}\underbrace{\left(1 + \frac{\overline{i_{nRn}^2}}{\Delta f} \Big/ \frac{\overline{i_{nRp}^2}}{\Delta f}\right)}_{F(\Delta f)}|Z_{tank}(\Delta f)|^2$$

- $F(\Delta f)$ is defined as

$$F(\Delta f) = \frac{\text{total noise in tank at frequency } \Delta f}{\text{noise in tank due to tank loss at frequency } \Delta f}$$

Fill in Expressions



- **Noise from tank is due to resistor R_p**

$$\frac{\overline{i^2_{nRp}}}{\Delta f} = 4kT\frac{1}{R_p} \quad \text{(single-sided spectrum)}$$

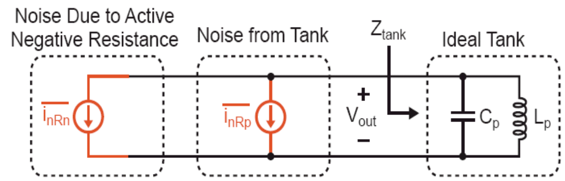
- **$Z_{tank}(\Delta f)$ found previously**

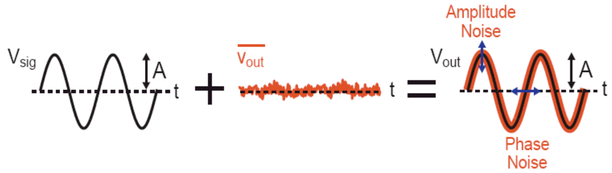
$$|Z_{tank}(\Delta f)|^2 \approx \left(\frac{R_p}{2Q}\frac{f_o}{\Delta f}\right)^2$$

- **Output noise spectral density expression (single-sided)**

$$\frac{\overline{v^2_{out}}}{\Delta f} = 4kT\frac{1}{R_p}F(\Delta f)\left(\frac{R_p}{2Q}\frac{f_o}{\Delta f}\right)^2 = \boxed{4kTF(\Delta f)R_p\left(\frac{1}{2Q}\frac{f_o}{\Delta f}\right)^2}$$

Separation into Amplitude and Phase Noise

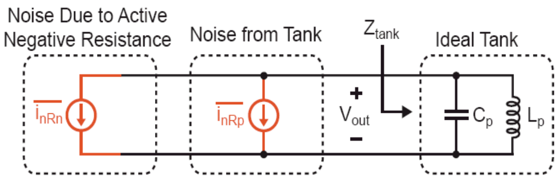


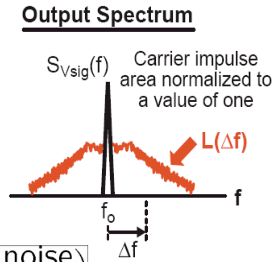


- **Equipartition theorem → states that noise impact splits evenly between amplitude and phase for V_{sig} being a sine wave**
  - **Amplitude variations suppressed by feedback in oscillator**

$$\Rightarrow \left.\frac{\overline{v^2_{out}}}{\Delta f}\right|_{\text{phase}} = \boxed{2kTF(\Delta f)R_p\left(\frac{1}{2Q}\frac{f_o}{\Delta f}\right)^2} \quad \text{(single-sided)}$$

Output Phase Noise Spectrum (Leeson's Formula)





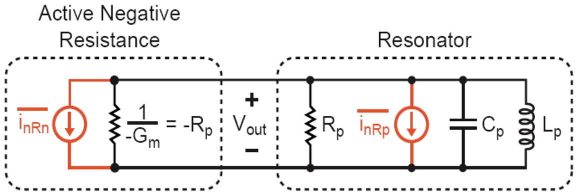
$$L(\Delta f) = 10\log\left(\frac{\text{Spectral density of noise}}{\text{Power of carrier}}\right)$$

- **All power calculations are referenced to the tank loss resistance, R_p**

$$P_{sig} = \frac{V^2_{sig,rms}}{R_p} = \frac{(A/\sqrt{2})^2}{R_p}, \quad S_{noise}(\Delta f) = \frac{1}{R_p}\frac{\overline{v^2_{out}}}{\Delta f}$$

$$L(\Delta f) = 10\log\left(\frac{S_{noise}(\Delta f)}{P_{sig}}\right) = \boxed{10\log\left(\frac{2kTF(\Delta f)}{P_{sig}}\left(\frac{1}{2Q}\frac{f_o}{\Delta f}\right)^2\right)}$$

Example: Active Noise Same as Tank Noise

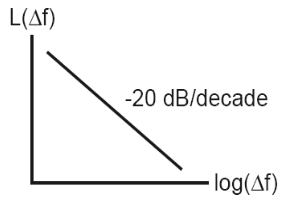


- **Noise factor for oscillator in this case is**

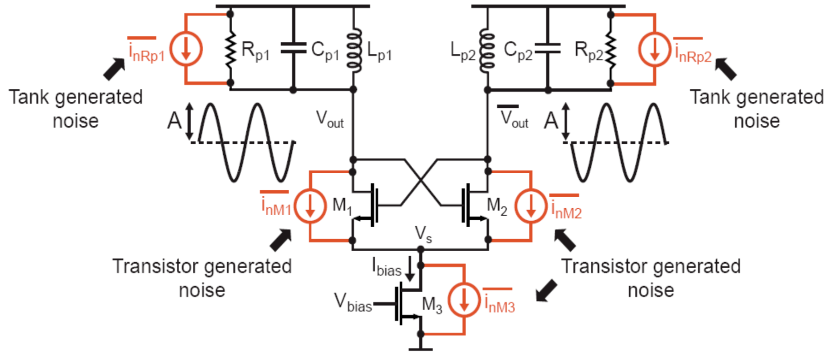
$$F(\Delta f) = 1 + \frac{\overline{i^2_{nRn}}}{\Delta f} \Big/ \frac{\overline{i^2_{nRp}}}{\Delta f} = 2$$

- **Resulting phase noise**

$$\boxed{L(\Delta f) = 10\log\left(\frac{4kT}{P_{sig}}\left(\frac{1}{2Q}\frac{f_o}{\Delta f}\right)^2\right)}$$

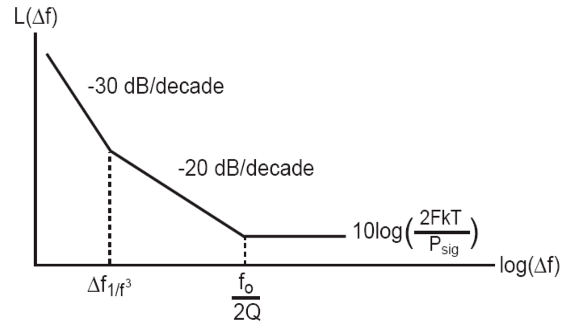


The Actual Situation is Much More Complicated



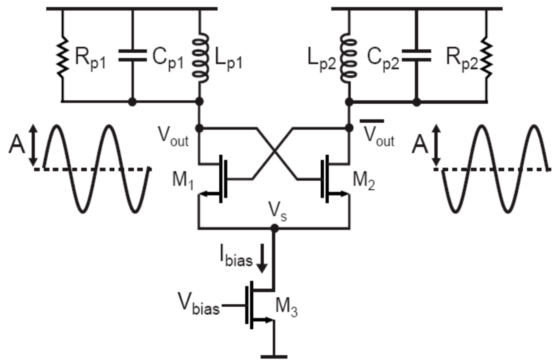
- Impact of tank generated noise easy to assess
- Impact of transistor generated noise is complicated
  - Noise from M_1 and M_2 is modulated on and off
  - Noise from M_3 is modulated before influencing V_{out}
  - Transistors have 1/f noise
- Also, transistors can degrade Q of tank

Phase Noise of A Practical Oscillator



- Phase noise drops at -20 dB/decade over a wide frequency range, but deviates from this at:
  - Low frequencies – slope increases (often -30 dB/decade)
  - High frequencies – slope flattens out (oscillator tank does not filter all noise sources)
- Frequency breakpoints and magnitude scaling are not readily predicted by the analysis approach taken so far

Approach for Negative Resistance Oscillator

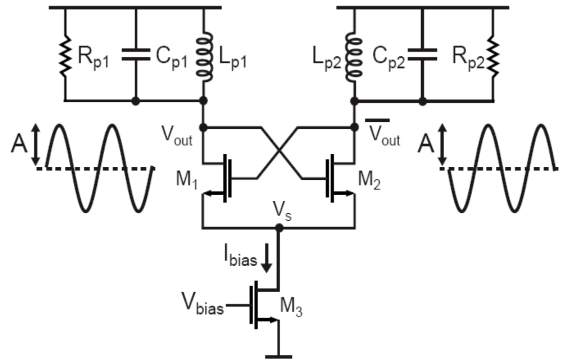


- Recall Leeson's formula

$$L(\Delta f) = 10\log\left(\frac{2kTF(\Delta f)}{P_{sig}}\left(\frac{1}{2Q}\frac{f_o}{\Delta f}\right)^2\right)$$

  - Key question: how do you determine $F(\Delta f)$?

$F(\Delta f)$ Has Been Determined for This Topology



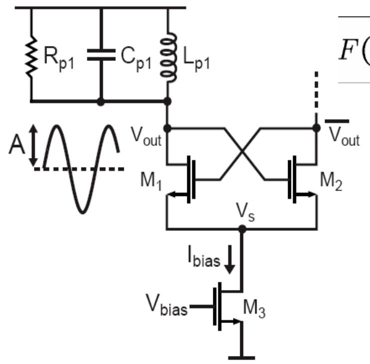
- Rael et. al. have come up with a closed form expression for $F(\Delta f)$ for the above topology
- In the region where phase noise falls at -20 dB/dec:

$$F(\Delta f) = 1 + \frac{2\gamma I_{bias}R_p}{\pi A} + \gamma\frac{4}{9}g_{do,M3}R_p \quad (R_p = R_{p1} = R_{p2})$$

References to Rael Work

- Phase noise analysis
  - J.J. Rael and A.A. Abidi, "Physical Processes of Phase Noise in Differential LC Oscillators", Custom Integrated Circuits Conference, 2000, pp 569-572
- Implementation
  - Emad Hegazi et. al., "A Filtering Technique to Lower LC Oscillator Phase Noise", JSSC, Dec 2001, pp 1921-1930

Designing for Minimum Phase Noise



$$F(\Delta f) = 1 + \underbrace{\frac{2\gamma I_{bias} R_p}{\pi A}}_{(B)} + \underbrace{\gamma \frac{4}{9} g_{do,M3} R_p}_{(C)}$$

(A) (B) (C)

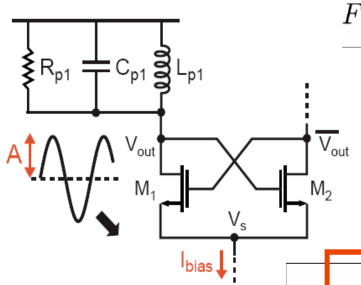
(A) Noise from tank resistance

(B) Noise from M_1 and M_2

(C) Noise from M_3

- To achieve minimum phase noise, we'd like to minimize F(Δf)
- The above formulation provides insight of how to do this
  - Key observation: (C) is often quite significant

Minimization of Component (B) in F(Δf)



$$F(\Delta f) = 1 + \frac{2\gamma I_{bias} R_p}{\pi A}$$

(B)

- Recall

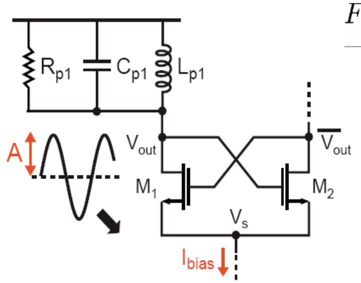
$$A = \frac{2}{\pi} I_{bias} R_p$$

$$\Rightarrow F(\Delta f) = 1 + \frac{2\gamma I_{bias} R_p}{\pi (2/\pi) I_{bias} R_p} = 1 + \gamma$$

- So, it would seem that I_{bias} has no effect!
  - Not true – want to maximize A (i.e. P_{sig}) to get best phase noise, as seen by:

$$L(\Delta f) = 10 \log \left(\frac{2kTF(\Delta f)}{P_{sig}} \left(\frac{1}{2Q} \frac{f_o}{\Delta f} \right)^2 \right)$$

Current-Limited Versus Voltage-Limited Regimes



$$F(\Delta f) = 1 + \frac{2\gamma I_{bias} R_p}{\pi A}$$

(B)

- Oscillation amplitude, A, cannot be increased above supply imposed limits
- If I_{bias} is increased above the point that A saturates, then (B) increases

- Current-limited regime: amplitude given by $A = \frac{2}{\pi} I_{bias} R_p$
- Voltage-limited regime: amplitude saturated

Best phase noise achieved at boundary between these regimes!