CIRF
Circuit Intégré Radio Fréquence

Lecture I

- ***Introduction***
- ***Baseband Pulse Transmission***
- ***Digital Passband Transmission***
- ***Circuit Non-idealities Effect***

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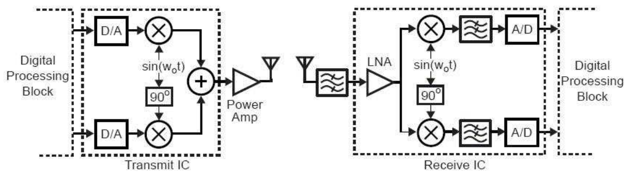
Wireless Systems

- Direct conversion architecture



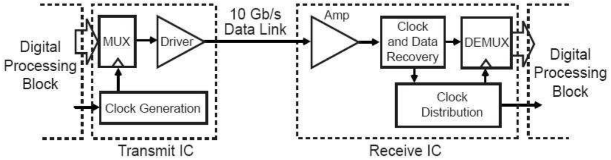
- Transmitter issues
  - Meeting the spectral mask (LO phase noise & feedthrough, quadrature accuracy), D/A accuracy, power amp linearity
- Receiver Issues
  - Meeting SNR (Noise figure, blocking performance, channel selectivity, LO phase noise, A/D nonlinearity and noise), selectivity (filtering), and emission requirements

Future Goals

- Low cost, low power, and small area solutions
  - New architectures and circuits!
- Increased spectral efficiency
  - Example: GSM cellphones (GMSK) to 8-PSK (Edge)
    - Requires a linear power amplifier!
- Increased data rates
  - Example: 802.11b (11 Mb/s) to 802.11a (> 50 Mb/s)
    - GFSK modulation changes to OFDM modulation
- Higher carrier frequencies
  - 802.11b (2.5 GHz) to 802.11a (5 GHz) to ? (60 GHz)
- New modulation formats
  - GMSK, CDMA, OFDM, pulse position modulation
- New application areas

High Speed Data Links

- A common architecture



- Transmitter Issues
  - Intersymbol interference (limited bandwidth of IC amplifiers, packaging), clock jitter, power, area
- Receiver Issue
  - Intersymbol interference (same as above), jitter from clock and data recovery, power, area

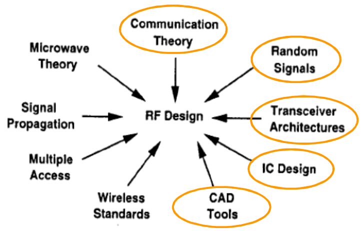
Future Goals

- Low cost, low power, small area solutions
  - New architectures and circuits!
- Increased data rates
  - 40 Gb/s for optical (moving to 120 Gb/s!)
    - Electronics is a limitation (optical issues getting significant)
  - > 5 Gb/s for backplane applications
    - The channel (i.e., the PC board trace) is the limitation
- High frequency compensation/equalization
  - Higher data rates, lower bit error rates (BER), improved robustness in the face of varying conditions
  - How do you do this at GHz speeds?
- Multi-level modulation
  - Better spectral efficiency (more bits in given bandwidth)

What are the Issues with Wireless Systems?

- Noise
  - Need to extract the radio signal with sufficient SNR
- Selectivity (filtering, processing gain)
  - Need to remove interferers (which are often much larger!)
- Nonlinearity
  - Degrades transmit spectral mask
  - Degrades selectivity for receiver

Multidisciplinarity of radio design



B. Razavi,
RF Microelectronics, Prentice Hall, 1998

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- S. Haykin, "Communication Systems", Wiley 1994.
- B. Razavi, "RF Microelectronics", Prentice Hall, 1997.
- M. Perrott, "High Speed Communication Circuits and Systems", M.I.T.OpenCourseWare, http://ocw.mit.edu/, Massachusetts Institute of Technology, 2003.
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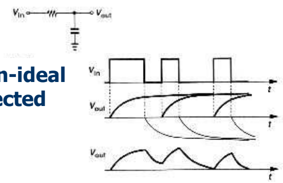
Digital Baseband Transmission

Major sources of errors in the detection of transmitted digital data:

ISI : InterSymbol Interference



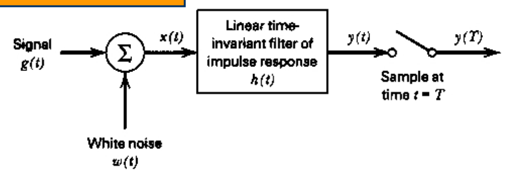
The result of data transmission over a non-ideal channel is that each received pulse is affected by adjacent pulses.

Channel Noise

Detecting a pulse transmitted over a channel that is corrupted by additive noise.

Matched Filter

Linear Receiver Model



- $g(t)$: transmitted pulse signal, binary symbol '1' or '0'.
- $w(t)$: channel noise, sample function of a white noise process of zero mean and power spectral density $N_0/2$.

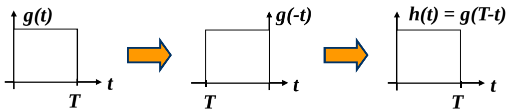
$$x(t) = g(t) + w(t) \quad , \quad 0 \le t \le T \longrightarrow h(t) \longrightarrow y(t) = g_0(t) + n(t)$$

- Filter Requirements, $h(t)$:
  - Make the instantaneous power in the output signal $g_0(t)$, measured at time $t=T$, as large as possible compared with the average power of the output noise, $n(t)$.

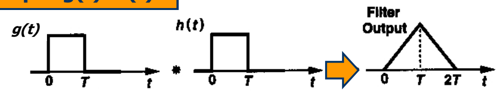
Matched Filter for Rectangular Pulse

$h(t)$ for a rectangular Pulse:

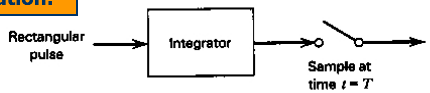
$$h_{opt}(t) = k\, g(T-t)$$



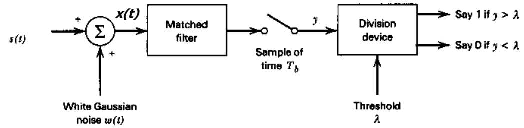
Filter Output g(t)*h(t):



Implementation:



Error Rate due to Noise



In the interval $0 \le t \le T_b$, the received signal:

$$x(t) = \begin{cases} +A + w(t) & , \text{ symbol '1' was sent} \\ -A + w(t) & , \text{ symbol '0' was sent} \end{cases}$$

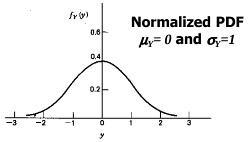
T_b is the bit duration, A is the transmitted pulse amplitude

- **The receiver has prior knowledge of the pulse shape but not its polarity.**
- **There are two possible kinds of error to be considered:**
  (1) Symbol '1' is chosen when a '0' was transmitted.
  (2) Symbol '0' is chosen when a '1' was transmitted.

PDF: Probability Density Function

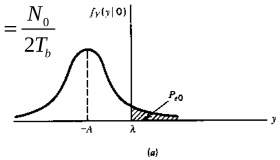
- **Gaussian Distribution:**

$$\boxed{f_Y(y) = \frac{1}{\sigma_Y \sqrt{2\pi}} \exp\left[-\frac{(y-\mu_Y)^2}{2\sigma_Y^2}\right]}$$



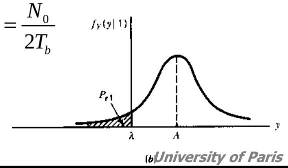
- **Symbol '0' was sent:** $\mu_Y = -A, \quad \sigma_Y^2 = \frac{N_0}{2T_b}$

$$P_{e0} = P(y > \lambda | \text{symbol '0' was sent})$$
$$= \int_{\lambda}^{\infty} f_Y(y|0)\, dy$$



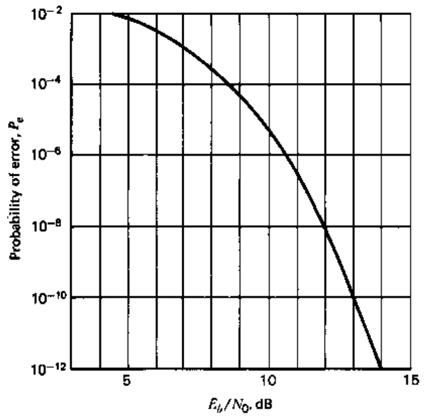
- **Symbol '1' was sent:** $\mu_Y = +A, \quad \sigma_Y^2 = \frac{N_0}{2T_b}$

$$P_{e0} = P(y < \lambda | \text{symbol '1' was sent})$$
$$= \int_{-\infty}^{\lambda} f_Y(y|1)\, dy$$



BER in a PCM receiver

$$P_e = p_0 P_{e0} + p_1 P_{e1}$$
$$P_{e0} = P_{e1}$$
$$p_0 = p_1 = \frac{1}{2}$$
$$P_e = P_{e0} = P_{e1}$$
$$\boxed{P_e = \frac{1}{2} erfc\left(\sqrt{\frac{E_b}{N_0}}\right)}$$



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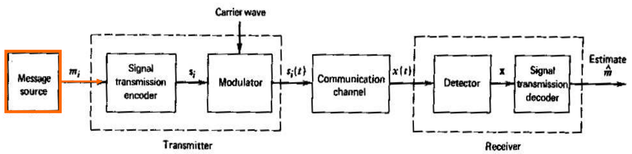
Why Modulation?

- In wired systems, coaxial lines exhibit superior shielding at higher frequencies
- In wireless systems, the antenna size should be a significant fraction of the wavelength to achieve a reasonable gain.
- Communication must occur in a certain part of the spectrum because of FCC regulations.
- Modulation allows simpler detection at the receive end in the presence of non-idealities in the communication channel.

Message Source

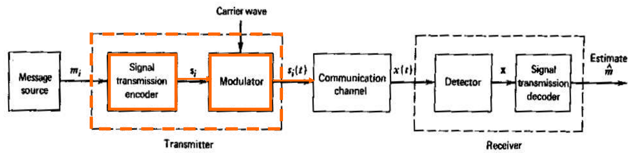


- m_i : one symbol every T seconds
- Symbols belong to an alphabet of M symbols: $m_1, m_2, \ldots, m_M$
- Message output probability:

$$P(m_1) = P(m_2) = \ldots = P(m_M)$$
$$p_i = P(m_i) = \frac{1}{M}$$

- Example: Quaternary PCM, 4 symbols: $00, 01, 10, 11$

Transmitter



- Signal Transmission Encoder: produces a vector s_i made up of N real elements, where $N \leq M$.
- Modulator: constructs a distinct signal $s_i(t)$ representing m_i of duration T .
- Energy of $s_i(t)$:

$$E_i = \int_0^T s_i^2(t) \quad dt, \quad i = 1, 2, \ldots, M$$

- $s_i(t)$ is real valued and transmitted every T seconds.

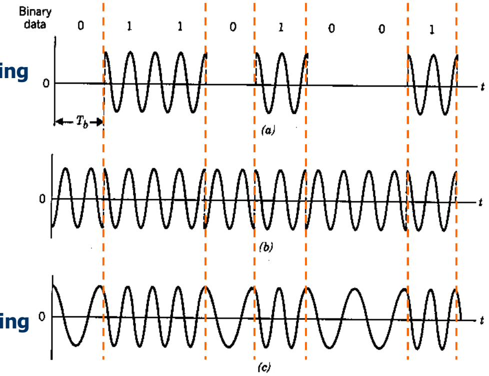
Examples of Transmitted signals: $s_i(t)$

• The modulator performs a step change in the amplitude, phase or frequency of the sinusoidal carrier

• ASK:
Amplitude Shift Keying

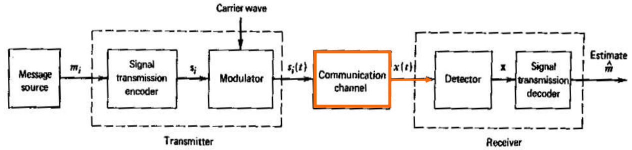
• PSK:
Phase Shift Keying

• FSK:
Frequency Shift Keying



Special case: Symbol Duration T = Bit Duration, T_b

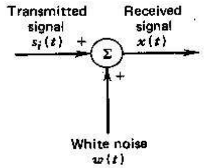
Communication Channel



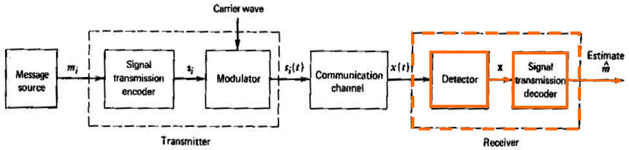
• Two Assumptions:
  •The channel is linear (no distortion).
  • $s_i(t)$ is perturbed by an Additive, zero-mean, stationnary, White, Gaussian Noise process (AWGN).

• Received signal $x(t)$:

$$x(t) = s_i(t) + w(t), \quad \begin{cases} 0 \le t \le T \\ i = 1, 2, \ldots, M \end{cases}$$



Receiver

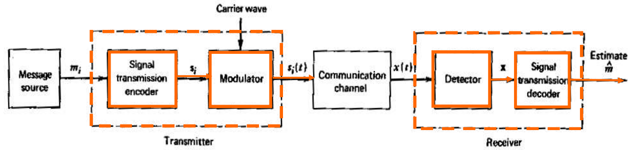


•TASK: observe received signal, $x(t)$, for a duration T and make a best estimate of transmitted symbol, m_i .
  •Detector: produces observation vector x .
  •Signal Transmission Decoder: estimates $\hat{m}$ using x, the modulation format and $P(m_i)$.

• The requirement is to design a receiver so as to minimize the average probability of symbol error:

$$P_e = \sum_{i=1}^{M} P(\hat{m} \neq m_i) P(m_i)$$

Coherent and Non-Coherent Detection



• Coherent Detection:
  - The receiver is time synchronized with the transmitter.
  - The receiver knows the instants of time when the modulator changes state.
  - The receiver is phase-locked to the transmitter.

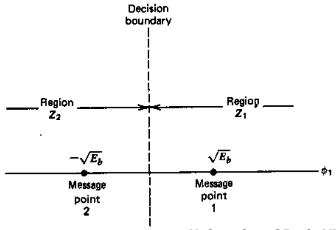
• Non-Coherent Detection:
  - No phase synchronism between transmitter and receiver.

Coherent Binary PSK:

- *M=2, N=1* $0 \le t \le T_b$ $s_1(t) = \sqrt{\frac{2E_b}{T_b}}\cos(2\pi f_c t)$ $s_2(t) = \sqrt{\frac{2E_b}{T_b}}\cos(2\pi f_c t + \pi)$
- **To ensure that each transmitted bit contains an integral number of cycles of the carrier wave, fc =nc/Tb, for some fixed integer nc.**
- **One basis function:** $\phi_1(t) = \sqrt{\frac{2}{T_b}}\cos(2\pi f_c t) \quad , \quad 0 \le t \le T_b$
- **Signal constellation consists of two message points:**

$$s_{11} = \int_0^{T_b} s_1(t)\,\phi_1(t)dt = \sqrt{E_b}$$

$$s_{21} = \int_0^{T_b} s_2(t)\,\phi_1(t)dt = -\sqrt{E_b}$$

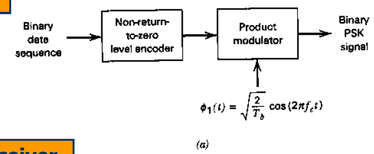


Generation and Detection of Coherent Binary PSK

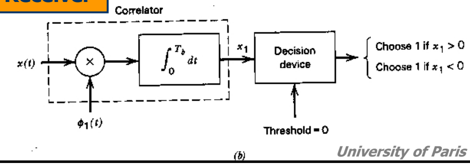
- **Assuming white Gaussian Noise with** *PSD = $N_0/2$* **, The Bit Error Rate for coherent binary PSK is:**

$$P_e = \frac{1}{2} erfc\left(\sqrt{\frac{E_b}{N_0}}\right)$$

Binary PSK Transmitter



Coherent Binary PSK Receiver



Coherent QPSK:

- *M=4, N=2:*

$$s_i(t) = \begin{cases} \sqrt{\frac{2E}{T}}\cos\left[2\pi f_c t + (2i-1)\frac{\pi}{4}\right] & , \quad 0 \le t \le T \\ 0 & , \quad \text{elsewhere} \end{cases}$$

$$s_1(t) = \sqrt{\frac{2E}{T}}\cos(2\pi\, f_c t + \frac{\pi}{4})$$

$$s_2(t) = \sqrt{\frac{2E}{T}}\cos(2\pi\, f_c t + 3\frac{\pi}{4})$$

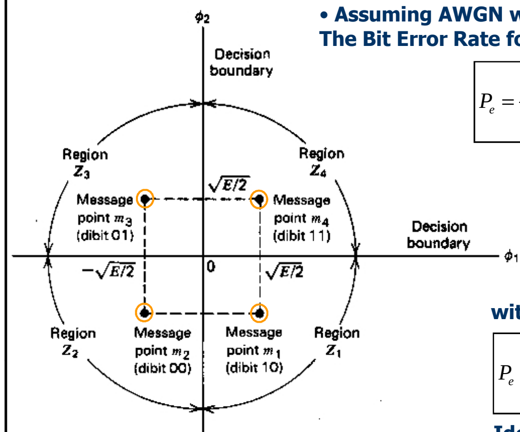
$$s_3(t) = \sqrt{\frac{2E}{T}}\cos(2\pi\, f_c t + 5\frac{\pi}{4})$$

$$s_4(t) = \sqrt{\frac{2E}{T}}\cos(2\pi\, f_c t + 7\frac{\pi}{4})$$

- **Two basis function:** $\phi_1(t) = \sqrt{\frac{2}{T}}\cos(2\pi\, f_c t) \quad , \quad 0 \le t \le T$

$$\phi_2(t) = \sqrt{\frac{2}{T}}\sin(2\pi\, f_c t) \quad , \quad 0 \le t \le T$$

Constellation Diagram of Coherent QPSK System



- **Assuming AWGN with** *PSD = $N_0/2$* **, The Bit Error Rate for coherent QPSK is:**

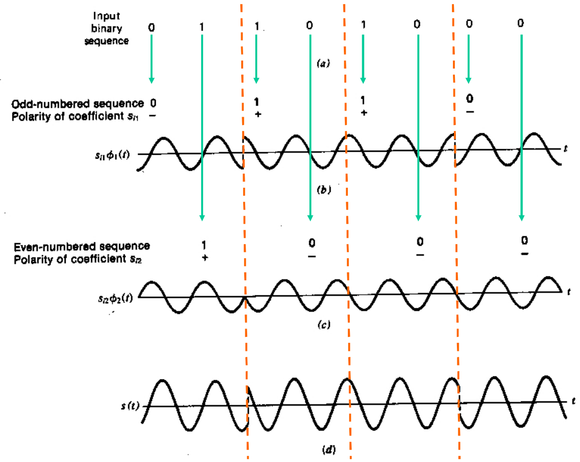
$$P_e = \frac{1}{2} erfc\left(\sqrt{\frac{E/2}{N_0}}\right)$$

with *E = 2 E_b*

$$P_e = \frac{1}{2} erfc\left(\sqrt{\frac{E_b}{N_0}}\right)$$

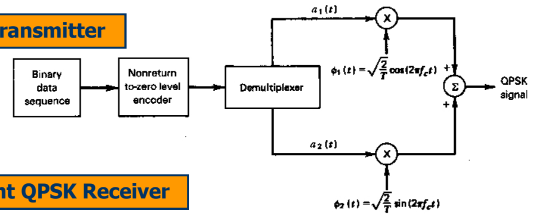
Identical to BPSK

QPSK waveform: 01101000

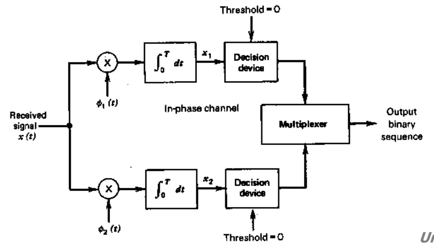


Generation and Detection of Coherent QPSK Signals

QPSK Transmitter



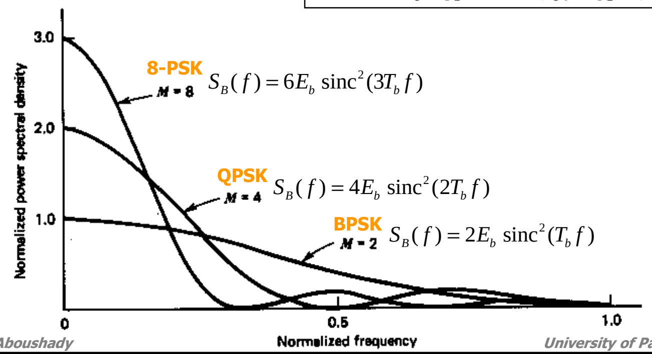
Coherent QPSK Receiver



Power Spectra of BPSK ,QPSK and M-ary PSK

- **Symbol Duration:** $\boxed{T = T_b \log_2 M}$
- **Power Spectral Density of an M-ary PSK signal:**

$$S_B(f) = 2E\,\mathrm{sinc}^2(Tf) = 2E_b \log_2 M\,\mathrm{sinc}^2(T_b f \log_2 M)$$



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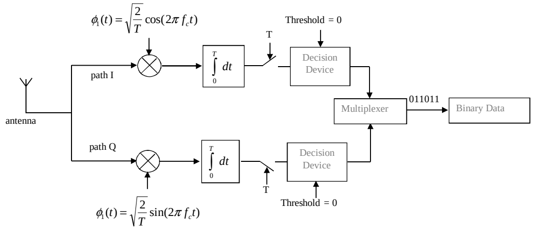
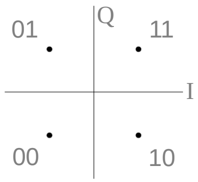
QPSK Receiver



QPSK Constellation Diagram



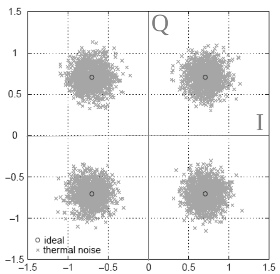
Receiver Circuit Non-Idealities

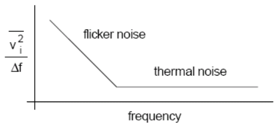
Circuit Noise (Thermal, 1/f)
Gain Mismatch
Phase Mismatch
DC Offset
Frequency Offset
Local Oscillator phase noise

Circuit Noise

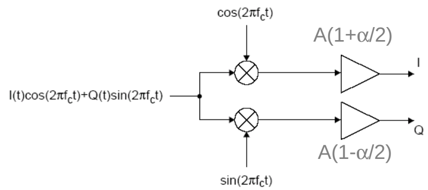
Circuit Noise:

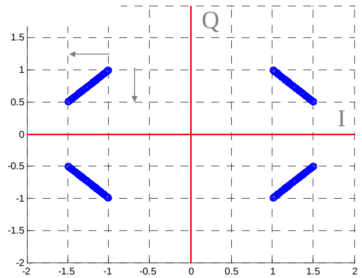
- Thermal Noise
  - Resistors
  - Transistors
- Flicker (1/f) Noise
  - MOS transistors



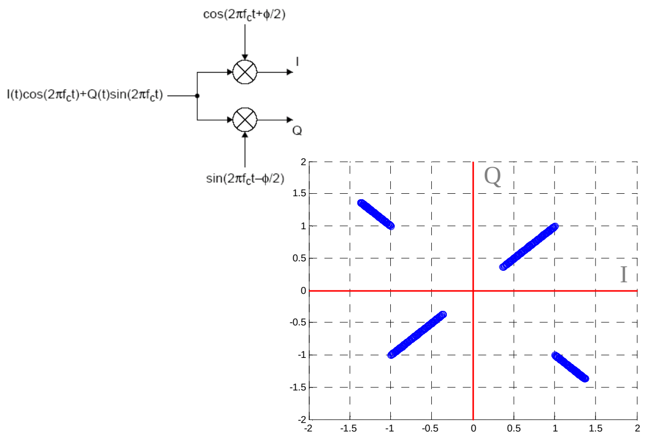


Gain Mismatch





Phase Mismatch



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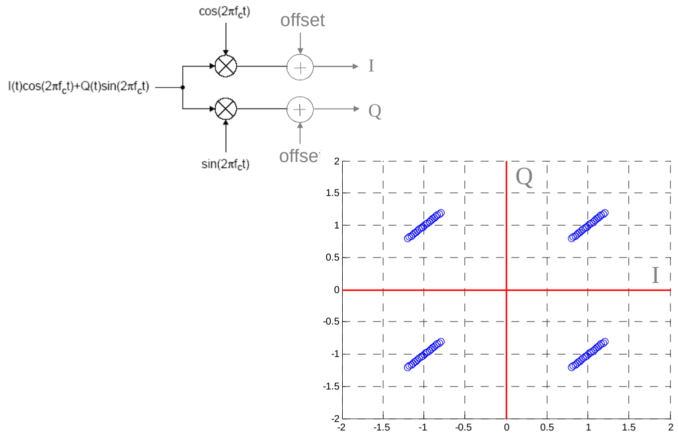
DC Offset



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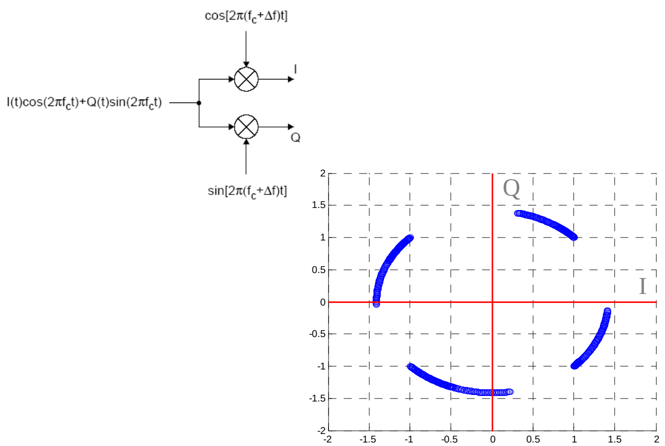
Frequency Offset



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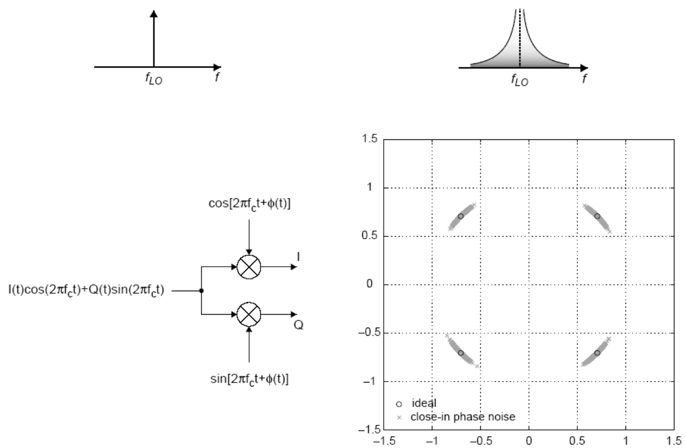
Local Oscillator Phase Noise



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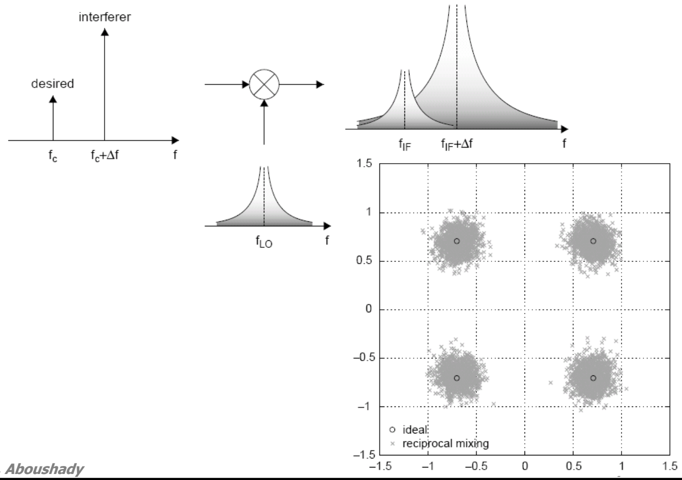
Reciprocal Mixing



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Lecture II

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

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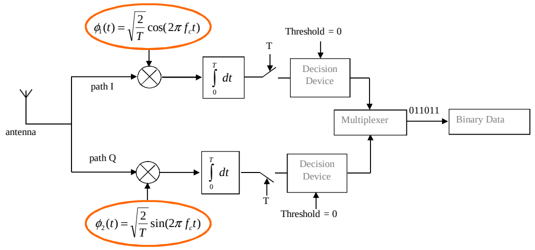
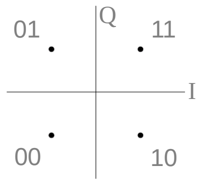
References

- 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.

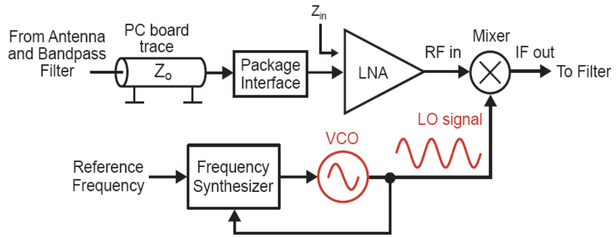
QPSK Receiver



QPSK Constellation Diagram

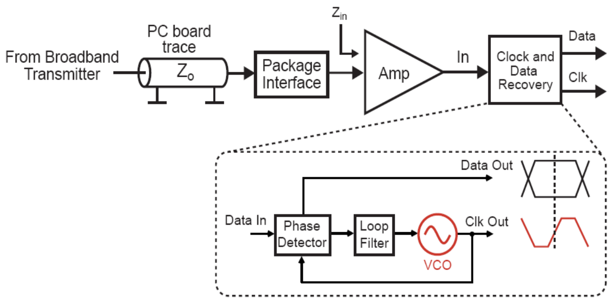


VCO Design for Wireless Systems



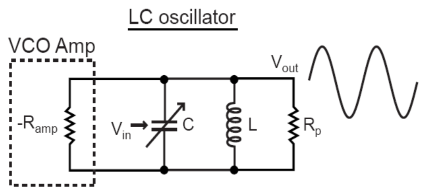
- **Design Issues**
  - **Tuning Range – need to cover all frequency channels**
  - **Noise – impacts receiver blocking and sensitivity performance**
  - **Power – want low power dissipation**
  - **Isolation – want to minimize noise pathways into VCO**
  - **Sensitivity to process/temp variations – need to make it manufacturable in high volume**

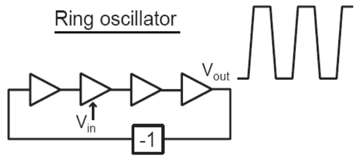
VCO Design For High Speed Data Links



- **Design Issues**
  - **Same as wireless, but:**
    - Required noise performance is often less stringent
    - Tuning range is often narrower

Popular VCO Structures



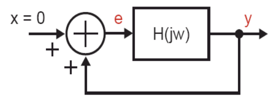


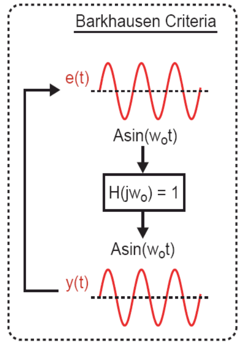
- **LC Oscillator: low phase noise, large area**
- **Ring Oscillator: easy to integrate, higher phase noise**

Further Info on Ring Oscillators

- Due to their relatively poor phase noise performance, ring oscillators are rarely used in RF systems
  - They are used quite often in high speed data links, though
- We will focus on LC oscillators in this lecture
- Some useful info on CMOS ring oscillators
  - Maneatis et. al., "Precise Delay Generation Using Coupled Oscillators", JSSC, Dec 1993 (look at pp 127-128 for delay cell description)
  - Todd Weigandt's PhD thesis – http://kabuki.eecs.berkeley.edu/~weigandt/

Barkhausen's Criteria for Oscillation





- Closed loop transfer function

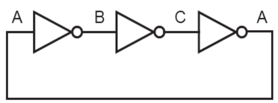
$$G(jw) = \frac{Y(jw)}{X(jw)} = \frac{H(jw)}{1 - H(jw)}$$

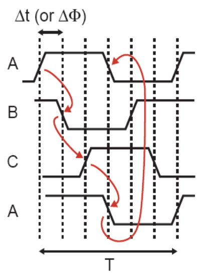
- Self-sustaining oscillation at frequency w_o if

$$H(jw_o) = 1$$

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

Ring Oscillator





- Gain is set to 1 by saturating characteristic of inverters
- Phase equals 360 degrees at frequency of oscillation
  - Assume N stages each with phase shift $\Delta\Phi$

$$2N\Delta\Phi = 360^o \;\Rightarrow\; \Delta\Phi = \frac{180^o}{N}$$

  - Alternately, N stages with delay Δt

$$2N\Delta t = T \;\Rightarrow\; \Delta t = \frac{T/2}{N}$$

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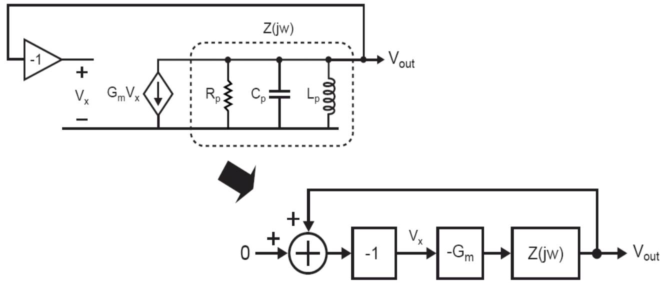
Lecture II

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

Hassan Aboushady
University of Paris VI

Resonator-Based Oscillator

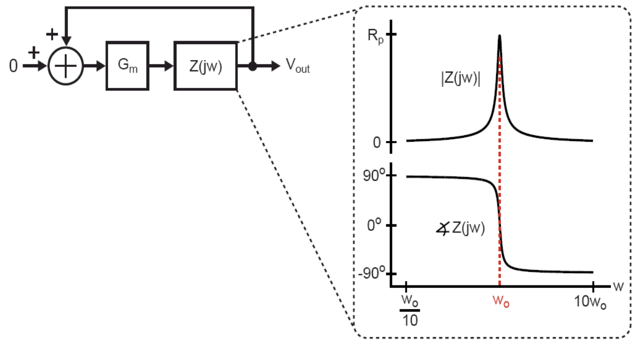


- **Barkhausen Criteria for oscillation at frequency w_o:**

$$G_m Z(jw_o) = 1$$

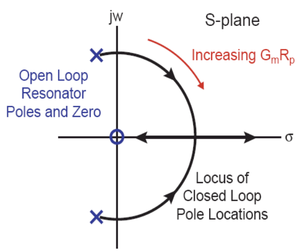
  - **Assuming G_m is purely real, $Z(jw_o)$ must also be purely real**

A Closer Look At Resonator-Based Oscillator



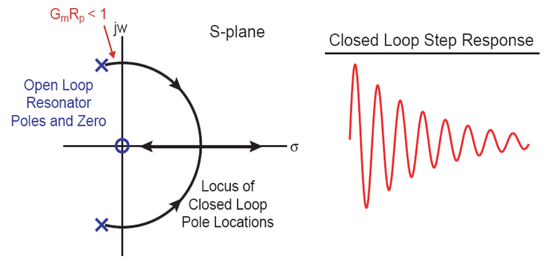
- **For parallel resonator at resonance**
  - **Looks like resistor (i.e., purely real) at resonance**
    - Phase condition is satisfied
    - Magnitude condition achieved by setting $G_m R_p = 1$

Impact of Different G_m Values



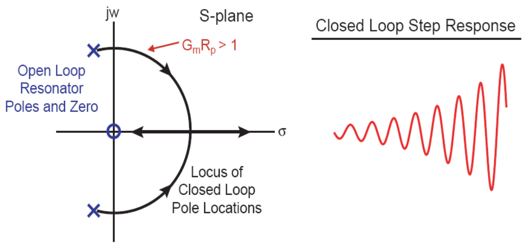
- **Root locus plot allows us to view closed loop pole locations as a function of open loop poles/zero and open loop gain ($G_m R_p$)**
  - **As gain ($G_m R_p$) increases, closed loop poles move into right half S-plane**

Impact of Setting G_m too low



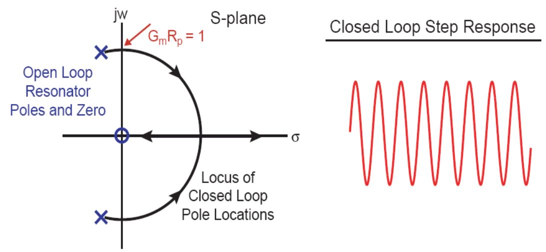
- **Closed loop poles end up in the left half S-plane**
  - **Underdamped response occurs**
    - Oscillation dies out

Impact of Setting G_m too High



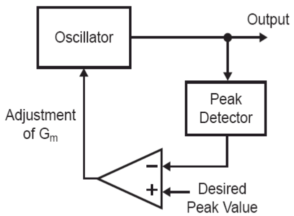
- Closed loop poles end up in the right half S-plane
  - Unstable response occurs
    - Waveform blows up!

Setting G_m To Just the Right Value



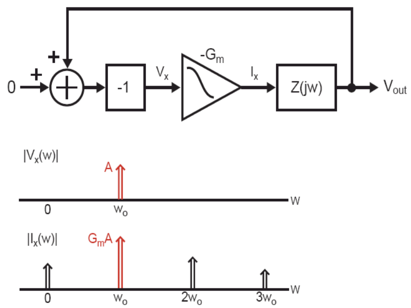
- Closed loop poles end up on jw axis
  - Oscillation maintained
- Issue – G_mR_p needs to *exactly* equal 1
  - How do we achieve this in practice?

Amplitude Feedback Loop



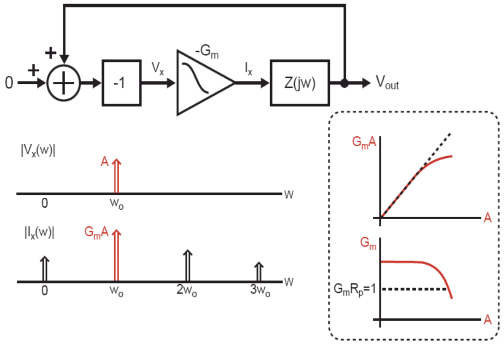
- One thought is to detect oscillator amplitude, and then adjust G_m so that it equals a desired value
  - By using feedback, we can precisely achieve $G_mR_p = 1$
- Issues
  - Complex, requires power, and adds noise

Leveraging Amplifier Nonlinearity as Feedback



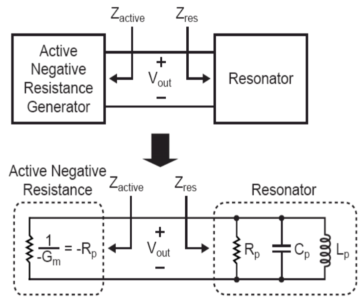
- Practical transconductance amplifiers have saturating characteristics
  - Harmonics created, but filtered out by resonator
  - Our interest is in the relationship between the input and the fundamental of the output

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

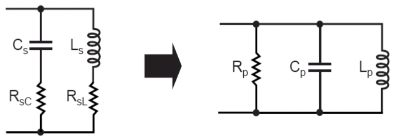


- Convenient for intuitive analysis
- Here we seek to cancel out loss in tank with a negative resistance element
  - To achieve sustained oscillation, we must have

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

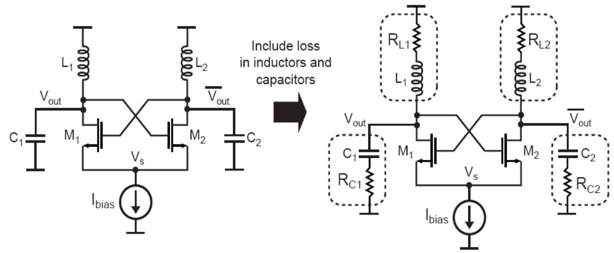
One-Port Modeling Requires Parallel RLC Network

- 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



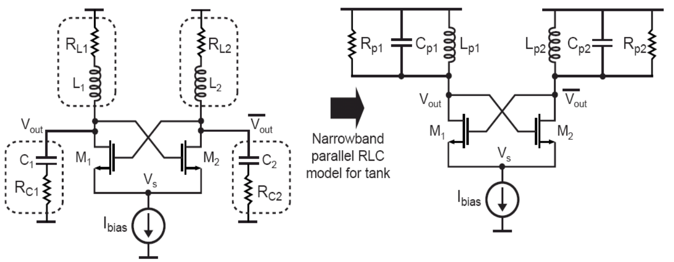
  - Warning – in practice, RLC networks can have secondary (or more) resonant frequencies, which cause undesirable behavior
    - Equivalent parallel network masks this problem in hand analysis
    - Simulation will reveal the problem

Example – Negative Resistance Oscillator



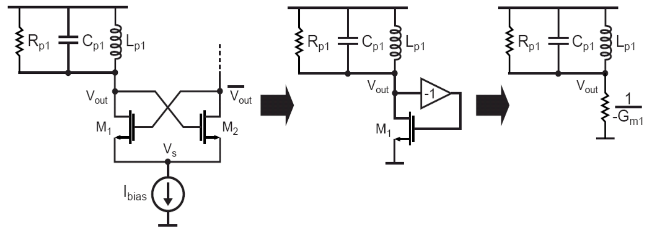
- This type of oscillator structure is quite popular in current CMOS implementations
  - Advantages
    - Simple topology
    - Differential implementation (good for feeding differential circuits)
    - Good phase noise performance can be achieved

Analysis of Negative Resistance Oscillator (Step 1)



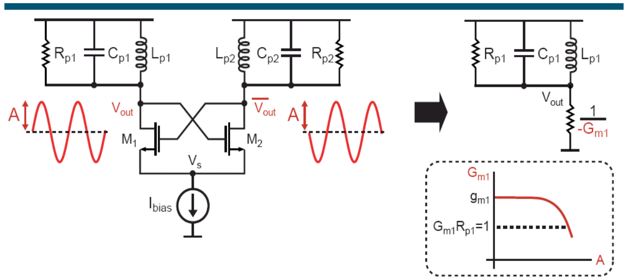
- Derive a parallel RLC network that includes the loss of the tank inductor and capacitor
  - Typically, such loss is dominated by series resistance in the inductor

Analysis of Negative Resistance Oscillator (Step 2)



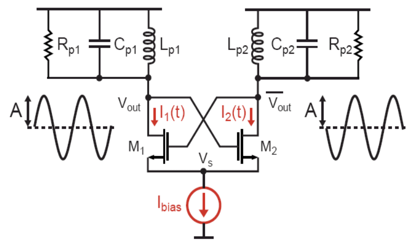
- Split oscillator circuit into half circuits to simplify analysis
  - Leverages the fact that we can approximate V_s as being incremental ground (this is not quite true, but close enough)
- Recognize that we have a diode connected device with a negative transconductance value
  - Replace with negative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

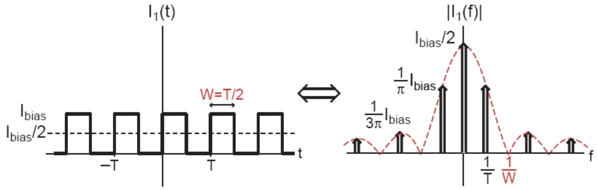
Calculation of Oscillator Swing



- 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)

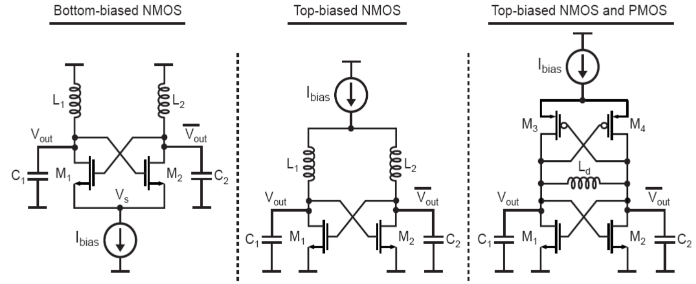


- Fundamental component is

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

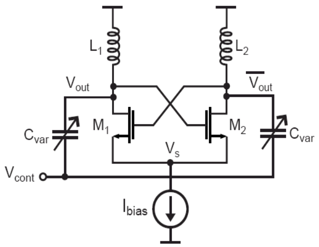
- Resulting oscillator amplitude $\boxed{A = \frac{2}{\pi} I_{bias} R_p}$

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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Lecture II

- ***Introduction***
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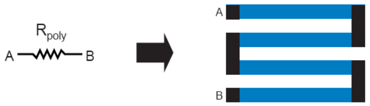
Hassan Aboushady
University of Paris VI

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

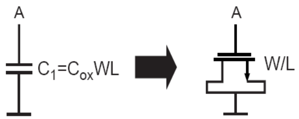
- Bias a MOS device in its triode region

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

- 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)

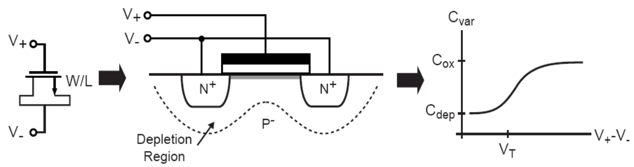
- MOS devices offer the highest capacitance per unit area
  - Limited to a one terminal device
  - Voltage must be high enough to invert the channel



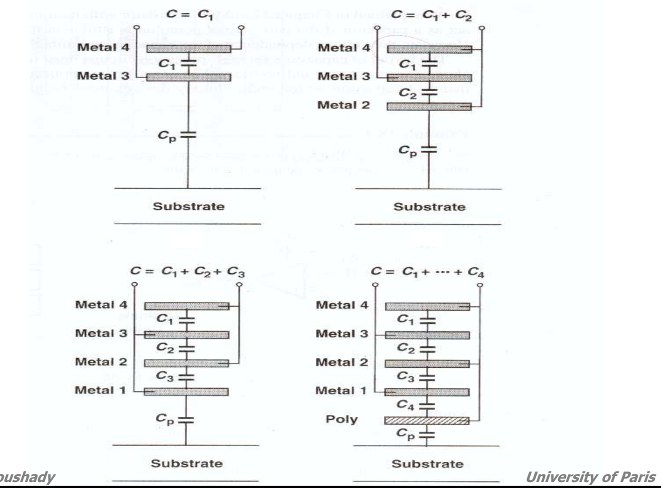
- Key parameters
  - Capacitance value
    - Raw cap value from MOS device is 6.1 fF/μ m^2 for 0.24u CMOS
  - Q (i.e., amount of series resistance)
    - Maximized with minimum L (tradeoff with area efficiency)

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

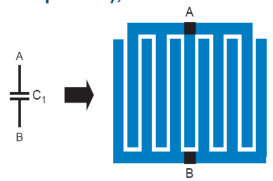


Vertical Metal Capacitors



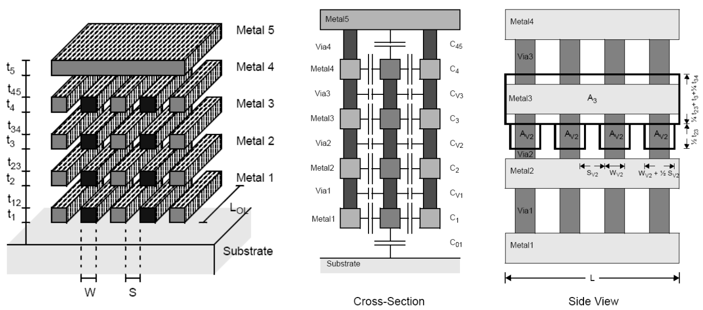
Lateral Metal Capacitors

- 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 $\pm 10\%$
  - Parasitic side cap is symmetric, less than 10% of cap value
- Example: C_T = 1.5 fF/μm^2 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



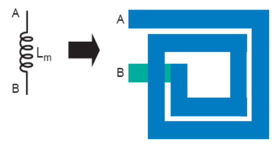
$$C_{mesh} = 2(C_1 + C_{V1} + C_2 + C_{V2} + C_3 + C_{V3} + C_4) + C_{45}$$

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

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

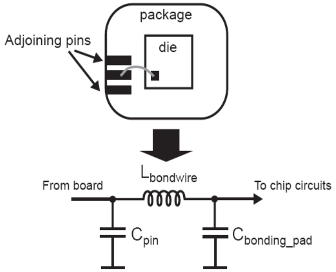
- Create integrated inductor using spiral shape on top level metals (may also want a patterned ground shield)



  - 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

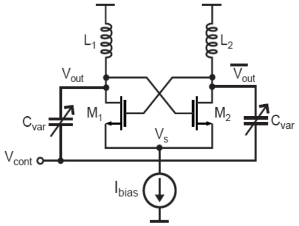


- Key parameters
  - Inductance (≈ 1 nH/mm – usually achieve 1-5 nH)
  - Q (much higher than spiral inductors – typically > 40)

Other Types of Resonators

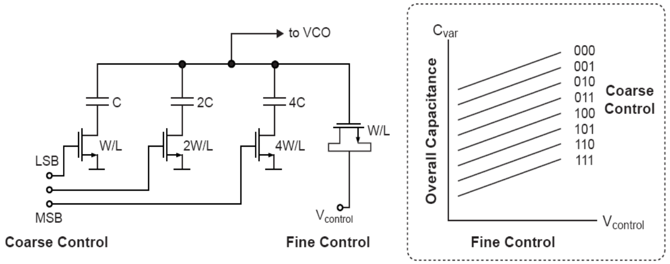
- Quartz crystal
  - Very high Q, and very accurate and stable resonant frequency
    - Confined to low frequencies (< 200 MHz)
    - Non-integrated
  - Used to create low noise, accurate, "reference" oscillators
- SAW devices
  - High frequency, but poor accuracy (for resonant frequency)
- MEMS devices
  - 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!

Supply Pulling and Pushing



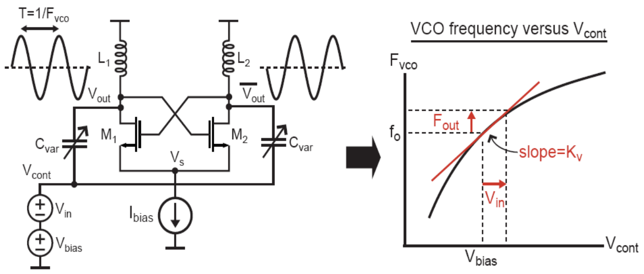
- Supply voltage has an impact on the VCO frequency
  - Voltage across varactor will vary, thereby causing a shift in its capacitance
  - Voltage across transistor drain junctions will vary, thereby causing a shift in its depletion capacitance
- This problem is addressed by building a supply regulator specifically for the VCO

A Method To Increase Q of MOS Varactor



- High Q metal caps are switched in to provide coarse tuning
- Low Q MOS varactor used to obtain fine tuning
- See Hegazi et. al., "A Filtering Technique to Lower LC Oscillator Phase Noise", JSSC, Dec 2001, pp 1921-1930

Model for Voltage to Frequency Mapping of VCO



- Model VCO in a small signal manner by looking at deviations in frequency about the bias point
  - Assume linear relationship between input voltage and output frequency

$$\boxed{F_{out}(t) = K_v v_{in}(t)}$$

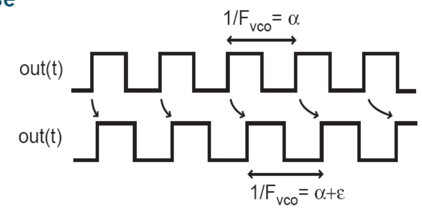
Model for Voltage to Phase Mapping of VCO

$$F_{out}(t) = K_v v_{in}(t)$$

- Phase is more convenient than frequency for analysis
  - The two are related through an integral relationship

$$\Phi_{out}(t) = \int_{-\infty}^{t} 2\pi F_{out}(\tau)d\tau = \int_{-\infty}^{t} 2\pi K_v v_{in}(\tau)d\tau$$

- Intuition of integral relationship between frequency and phase



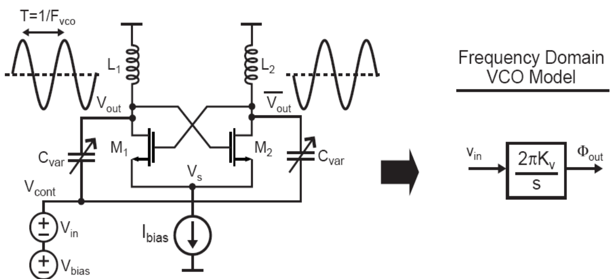
Frequency Domain Model of VCO

- Take Laplace Transform of phase relationship

$$\Phi_{out}(t) = \int_{-\infty}^{t} 2\pi K_v v_{in}(\tau)d\tau$$

$$\Rightarrow \quad \Phi_{out}(s) = \frac{2\pi K_v v_{in}(s)}{s}$$

  - Note that K_v is in units of Hz/V



CIRF
Circuit Intégré Radio Fréquence

Lecture II

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

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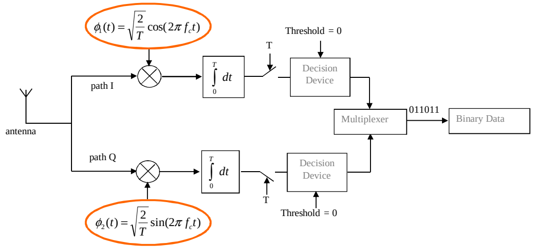
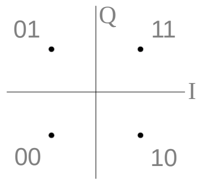
University of Paris VI

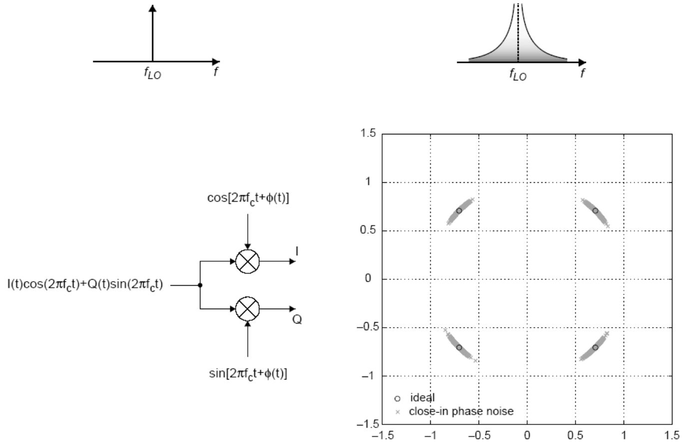
QPSK Receiver



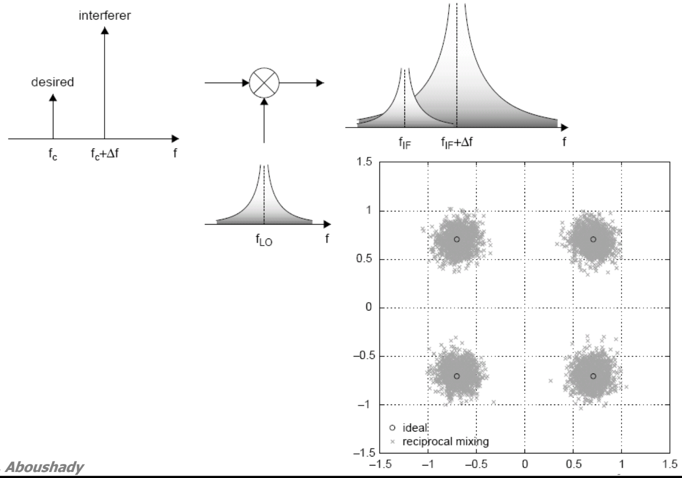
QPSK Constellation Diagram



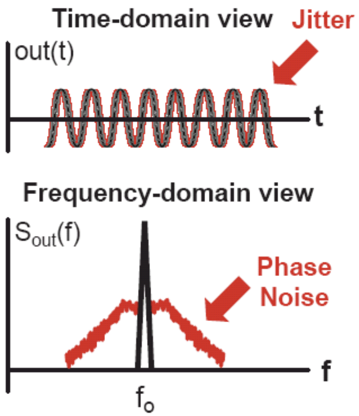
Local Oscillator Phase Noise



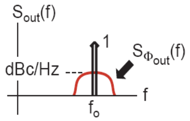
Reciprocal Mixing



Phase Noise in Oscillators



Measurement of Phase Noise in dBc/Hz



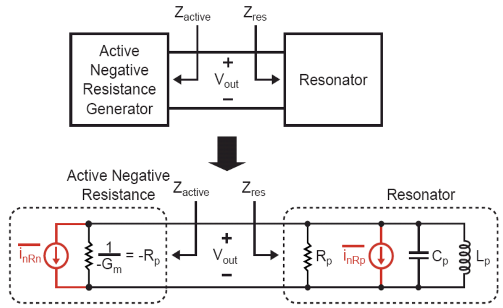
- **Definition of L(f)**

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

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

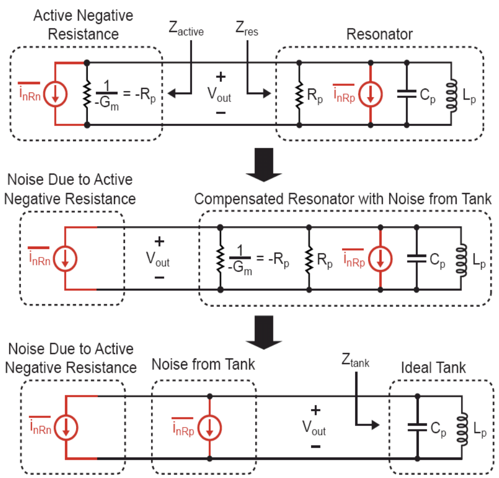
$$L(f) = 10\log\left(\frac{S_{\Phi_{out}}(f)}{1}\right) = 10\log(S_{\Phi_{out}}(f))$$

Calculation of Intrinsic Phase Noise in Oscillators

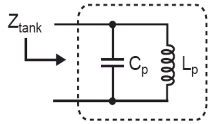


- **Noise sources in oscillators are put in two categories**
  - **Noise due to tank loss**
  - **Noise due to active negative resistance**
- **We want to determine how these noise sources influence the phase noise of the oscillator**

Equivalent Model for Noise Calculations



Calculate Impedance Across Ideal LC Tank Circuit



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

- **Calculate input impedance about resonance**

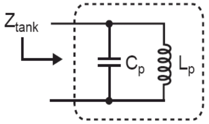
Consider $w = w_o + \Delta w$, where $w_o = \frac{1}{\sqrt{L_p C_p}}$

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

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

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

A Convenient Parameterization of LC Tank Impedance



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

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

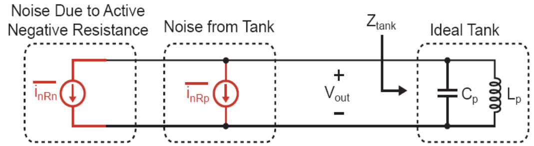
$$Q = R_p w_o C_p \;\Rightarrow\; \frac{1}{w_o C_p} = \frac{R_p}{Q}$$

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

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

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

Overall Noise Output Spectral Density



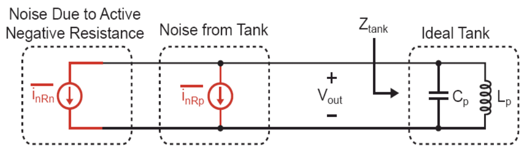
- Assume noise from active negative resistance element and tank are uncorrelated

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

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

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

Parameterize Noise Output Spectral Density



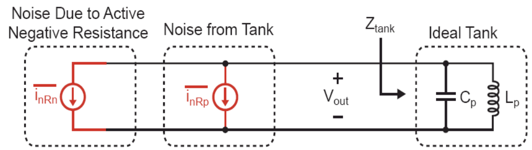
- From previous slide

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

- F(Δf) 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

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

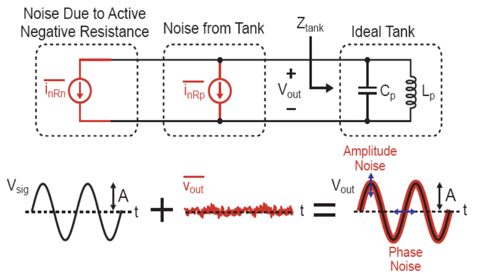
- $Z_{tank}(\Delta f)$ found previously

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

- Output noise spectral density expression (single-sided)

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

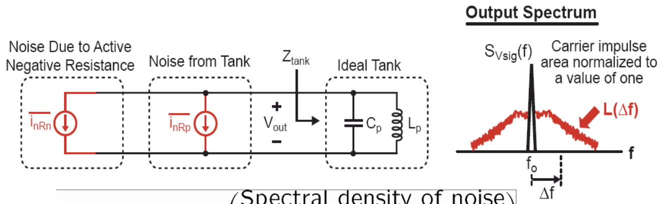
Separation into Amplitude and Phase Noise



- 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

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

Output Phase Noise Spectrum (Leeson's Formula)



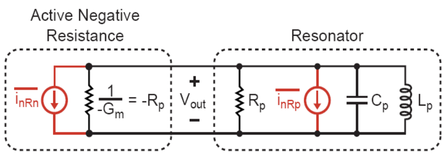
$$L(\Delta f) = 10\log\left(\frac{\text{Spectral density of noise}}{\text{Power of carrier}}\right)$$

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

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

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

Example: Active Noise Same as Tank Noise

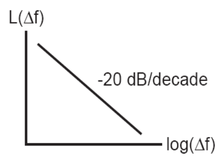


- Noise factor for oscillator in this case is

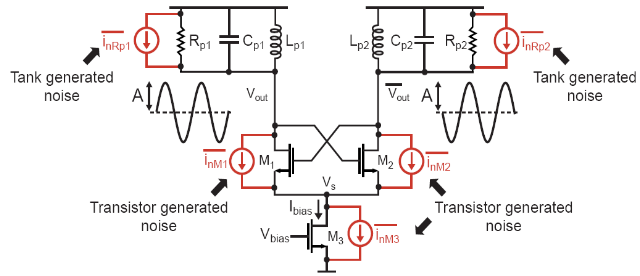
$$F(\Delta f) = 1 + \frac{\overline{i_{nRn}^2}}{\Delta f} \Big/ \frac{\overline{i_{nRp}^2}}{\Delta f} = 2$$

- Resulting phase noise

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

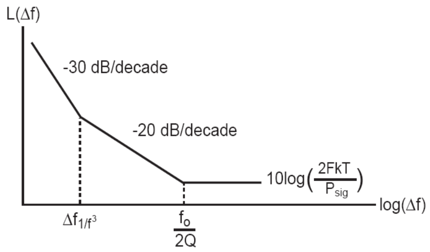


The Actual Situation is Much More Complicated



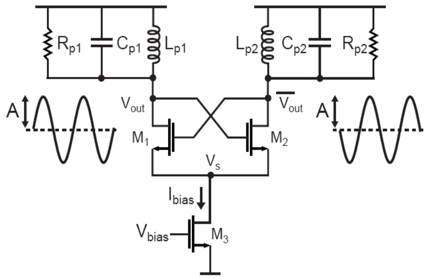
- Impact of tank generated noise easy to assess
- Impact of transistor generated noise is complicated
  - Noise from M_1 and M_2 is modulated on and off
  - Noise from M_3 is modulated before influencing V_{out}
  - Transistors have 1/f noise
- Also, transistors can degrade Q of tank

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

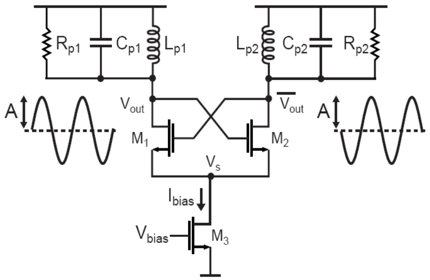


- Recall Leeson's formula

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

  - Key question: how do you determine F(Δf)?

F(Δf) Has Been Determined for This Topology



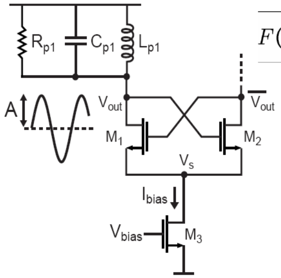
- Rael et. al. have come up with a closed form expression for F(Δf) for the above topology
- In the region where phase noise falls at -20 dB/dec:

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

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



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

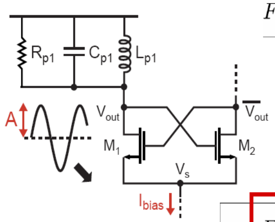
(A) Noise from tank resistance

(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)



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

- Recall

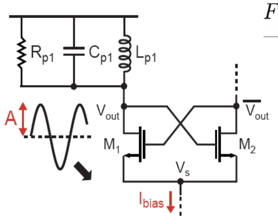
$$A = \frac{2}{\pi} I_{bias} R_p$$

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

- 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



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

- 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!