RF Harmonic Oscillators Integrated in Silicon Technologies

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SSCS Distinguished Lecture

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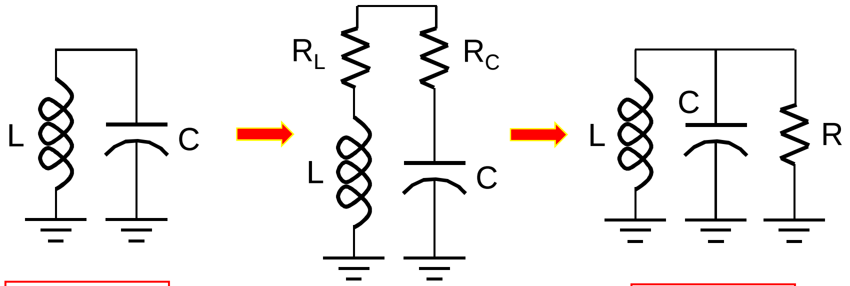
Overview

- Popular harmonic oscillators
  - Phase noise
- Architectures for low $1/f^2$ and/or $1/f^3$ phase noise
- Series-resonance oscillator
- Design techniques for very wide frequency tuning range RF CMOS VCOs

LC resonator

We begin with an inductor-capacitor resonator



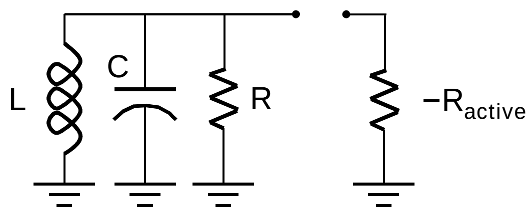
$$\omega_0 = \frac{1}{\sqrt{LC}}$$

$$\omega_0 \approx \frac{1}{\sqrt{LC}}$$

$$Z(\omega_0) = R$$

Building a harmonic oscillator

Tank losses are compensated by an active negative resistance in parallel to the tank

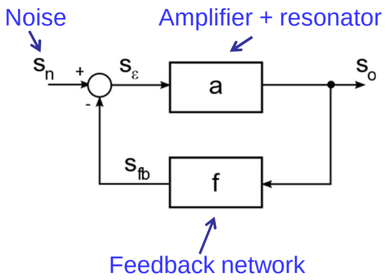


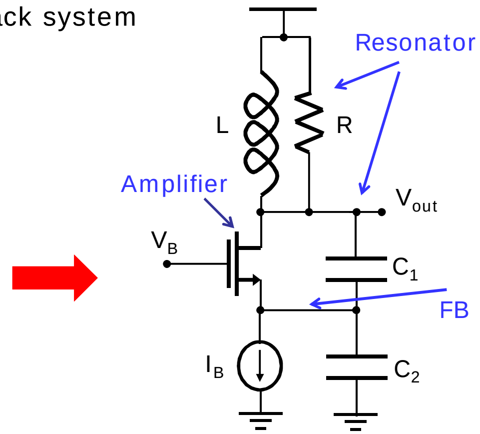
$$R_{active} < R$$

$$Q = \omega_0 RC = \frac{R}{\omega_0 L}$$

Colpitts oscillator

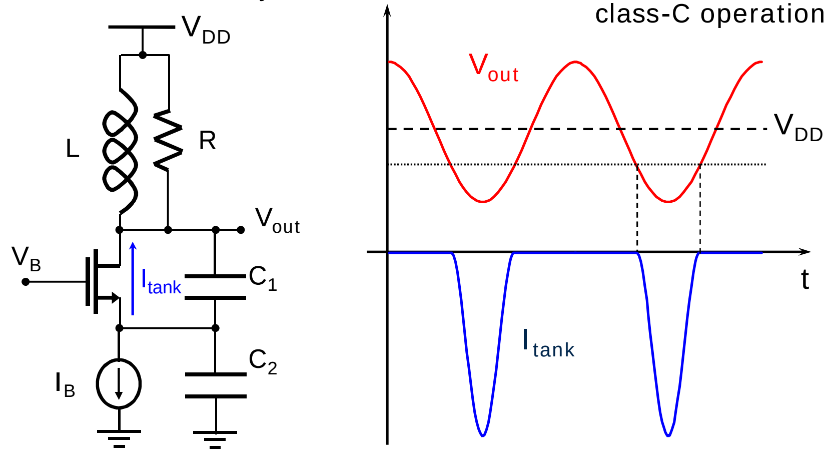
Oscillator as positive-feedback system



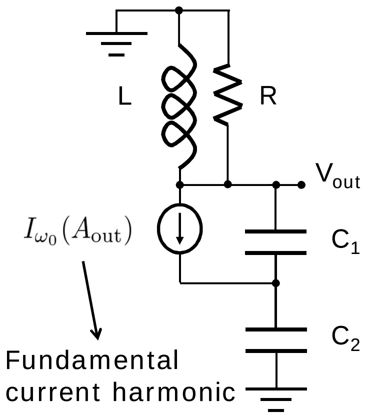


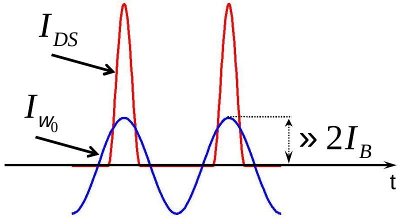
Colpitts oscillator

Classical embodiment of active negative resistance with only one active device



Analysis with Describing Function





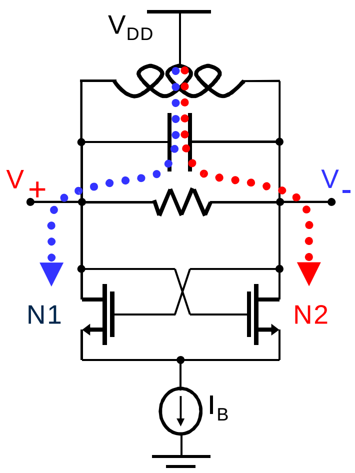
$$w_o = \frac{1}{\sqrt{LC}}$$

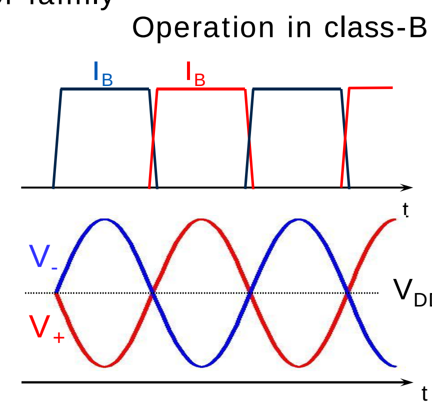
$$A_{out} = I_{w_0} R(1 - n) \approx 2 I_B R(1 - n)$$

$$C = \frac{C_1 C_2}{C_1 + C_2} \qquad n = \frac{C_1}{C_1 + C_2}$$

Cross-coupled differential-pair oscillator

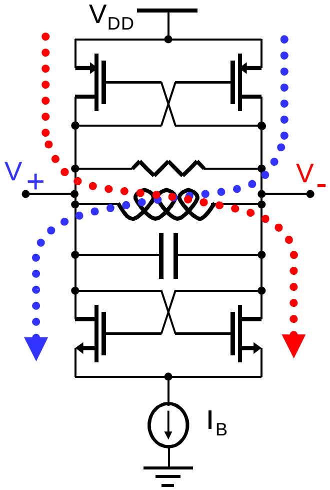
Extremely popular oscillator family

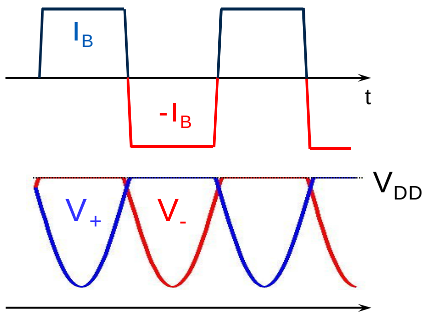




$$A_+ = A_- = \frac{2}{p} I_B \frac{R}{2} \Rightarrow A_{diff} = \frac{2}{p} I_B R$$

Class-B with double switch pair





$$A_{diff} = \frac{4}{\pi} I_B R$$

Compared to single-switch-pair: double amplitude, but also double V_{DD} (no swing above V_{DD})

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  - Phase noise
- Architectures for low $1/f^2$ and/or $1/f^3$ phase noise
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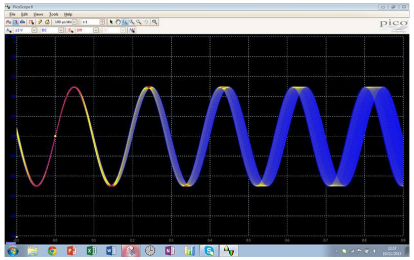
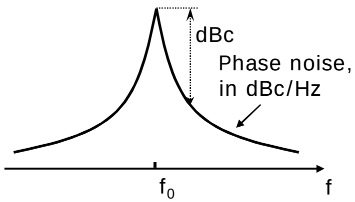
Real oscillations

- Phase uncertainty grows with time à *jitter*
  - Caused by various noise sources
- Jitter increases without bound in a free-running oscillator
- In the frequency domain, the oscillator displays *phase noise*



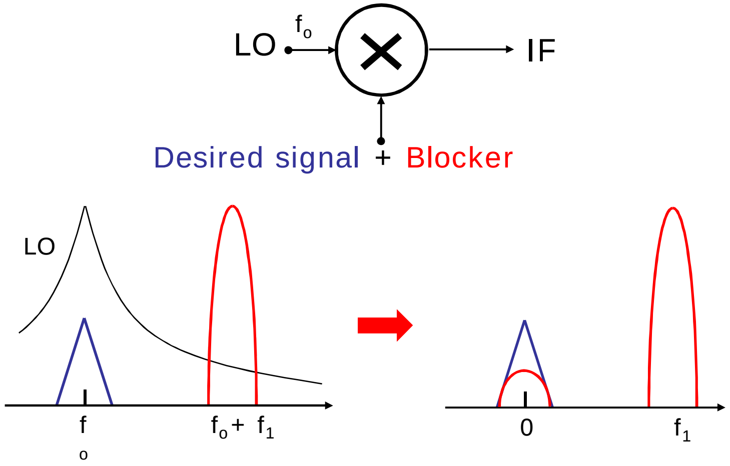


Why bother?

Phase noise in transceiver is important for at least three reasons:

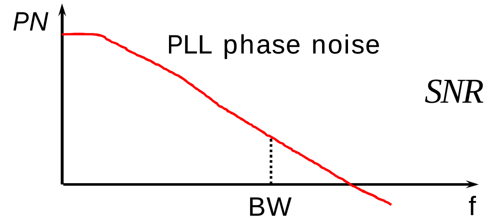
- In a receiver, it can downconvert large nearby signals on top of the desired signal
- In a transmitter, it can increase the noise floor in the receive band
- In both, it can directly corrupt the phase information in the signal
  – Not seldom, the phase noise of the VCO is the bottleneck for the whole radio performance

Reciprocal mixing

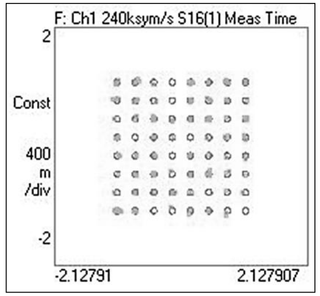


Noise floor raised by mixing of LO phase noise with blocker

Phase noise and SNR (EVM)



$$SNR=\int_{BW}^{BW} PN = 2\int_{0}^{BW} PN$$

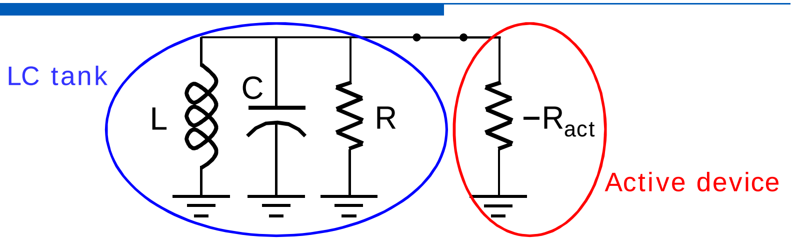


Error vector magnitude

$$EVM=10^{-\frac{SNR}{20}}$$

Example: 64QAM in WCDMA RX; EVM = 2.5%

Phase noise – LTI approach

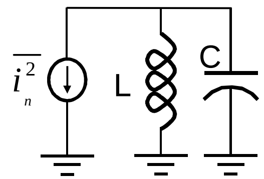


$$Y(\omega)=\left(G+j\omega C+\frac{1}{j\omega L}\right)-G_{act}=\frac{(1-\omega^2 LC)}{j\omega L}$$

$$Z(\omega_0+\Delta\omega)\approx\frac{j(\omega_0+\Delta\omega)L}{\left(1-(\omega_0+\Delta\omega)^2 LC\right)}\approx -j\frac{\omega_0 L}{2\frac{\Delta\omega}{\omega_0}}$$

$$\left|Z(\omega_0+\Delta\omega)\right|=\frac{\omega_0 L}{2\frac{\Delta\omega}{\omega_0}}=R\cdot\frac{\omega_0 L}{R}\cdot\frac{1}{2\frac{\Delta\omega}{\omega_0}}=R\cdot\frac{\omega_0}{2Q\Delta\omega}$$

Phase noise from tank losses



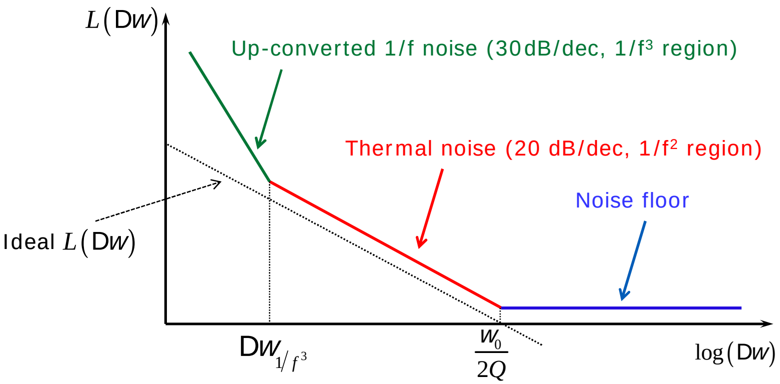
$$\frac{\overline{v_n^2}}{\Delta f}=\frac{\overline{i_n^2}}{\Delta f}\cdot\left|Z(\omega_0+\Delta\omega)\right|^2=4k_BTG\cdot\left(R\cdot\frac{\omega_0}{2Q\Delta\omega}\right)^2=4k_BTR\left(\frac{1}{2Q}\cdot\frac{\omega_0}{\Delta\omega}\right)^2$$

- Both amplitude and phase noise, but amplitude noise is rejected
- Thus, phase noise is defined as half the above expression, normalized to the output signal power (in dB below the carrier per Hertz, dBc/Hz):

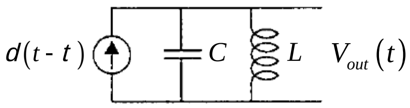
$$L(\Delta\omega)=10\log_{10}\left(\frac{\overline{v_n^2}/2}{A_{pk}^2/2}\right)=10\log_{10}\left(\frac{2k_BTR}{A_{pk}^2/2}\left(\frac{1}{2Q}\cdot\frac{\omega_0}{\Delta\omega}\right)^2\right)$$

Leeson's equation

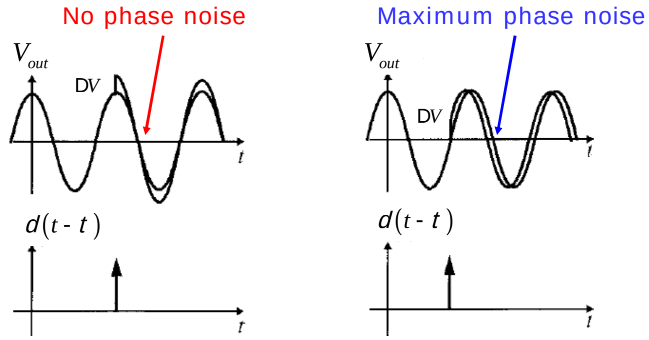
$$\mathcal{L}(\Delta\omega)=10\log_{10}\left\{F\frac{2k_BTR}{A_{pk}^2/2}\left[1+\left(\frac{1}{2Q}\cdot\frac{\omega_0}{\Delta\omega}\right)^2\right]\left(1+\frac{\Delta\omega_{1/f^3}}{\Delta\omega}\right)\right\}$$



Hajimiri and Lee's theory of phase noise



Conversion of noise into phase noise is time-dependent – LTV phase noise analysis needed!



Hajimiri and Lee, JSSC Feb. '98

Impulse sensitivity function (ISF, Γ)

- Current noise source $i_n(\phi)$ is weighed by associated $\Gamma_{i_n}(\phi)$

  → effective current noise $i_{n,eff}(\phi)=i_n(\phi)\cdot\Gamma_{i_n}(\phi)$ $(\phi=\omega_0 t)$

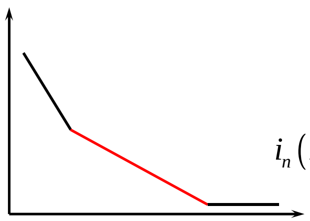
- ISF is dimensionless, frequency- and amplitude independent, with period 2π:

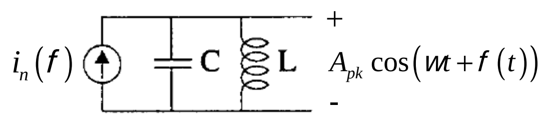
$$\Gamma(\phi)=\frac{c_0}{2}+\sum_{n=1}^{\infty}c_n\cos(n\phi+\phi_n)$$

Phase noise expression

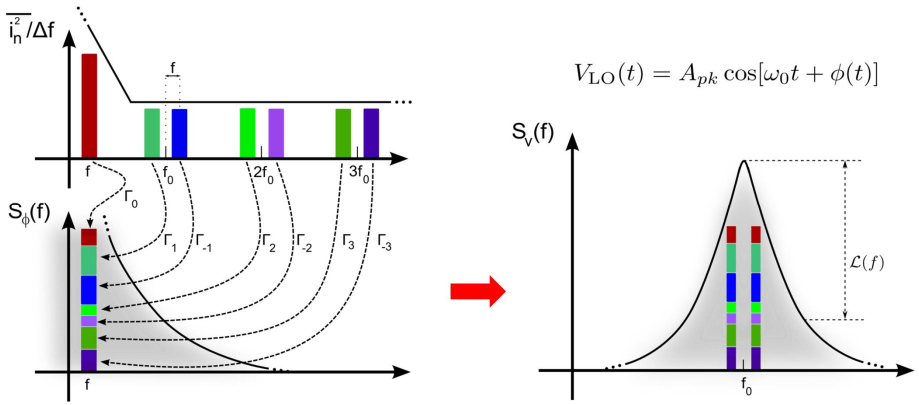
If $i_n(\phi)$ is a (cyclo)stationary white current noise source, its contribution to 1/f² phase noise is

$$\mathcal{L}(\Delta\omega)=10\log\left(\frac{\overline{i_{n,eff,rms}^2}}{2(CA_{pk})^2(\Delta\omega)^2}\right)$$



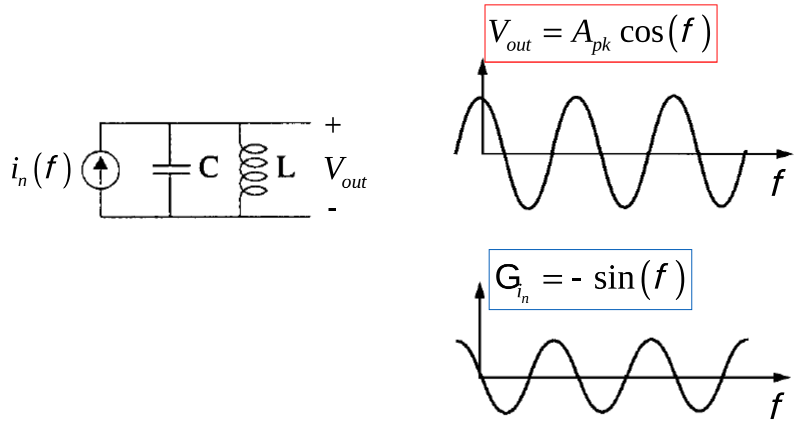


Graphical interpretation



$$\mathcal{L}(f) \approx S_\phi(f)$$

Example of ISF – LC oscillators



Hajimiri and Lee, JSSC Feb. '98; Andreani and Wang, JSSC Nov. '04

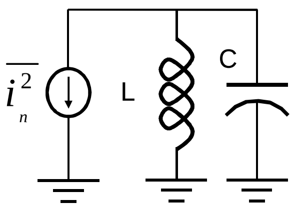
A particularly simple case

Parallel RLC resonator again – phase noise from tank losses:

$$L(\Delta\omega) = 10\log\left(\frac{\overline{i^2_{n,eff,rms}}}{2(CA_{pk})^2(\Delta\omega)^2}\right) = 10\log\left(\frac{\overline{i_n^2}\,\Gamma^2_{i_n,rms}}{2(CA_{pk})^2(\Delta\omega)^2}\right)$$

$$= 10\log\left(\frac{4k_BTG\cdot 1/2}{2(CA_{pk})^2(\Delta\omega)^2}\right)$$

$$= 10\log\left(\frac{2k_BTR}{A_{pk}^2/2}\left(\frac{1}{2Q}\frac{\omega_0}{\Delta\omega}\right)^2\right)$$

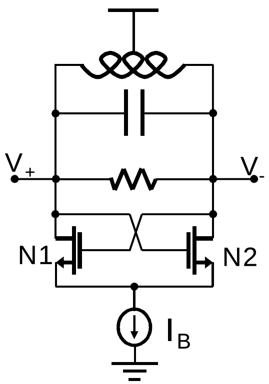


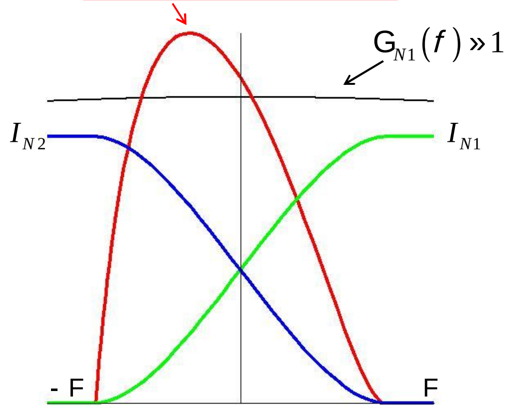
Leeson with F=1 recovered without any *ad hoc* assumptions!

Phase noise from MOS pair

$$\overline{i^2_{N1}}(\phi) = 4k_BT\gamma_n g_{m,N1}(\phi)$$

$$\overline{i^2_{N1,eff}}(\phi) = \overline{i^2_{N1}}(\phi)\cdot\Gamma^2_{N1}(\phi)$$





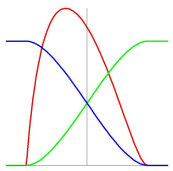
Two commutations in one oscillation period

Total phase noise

Tank — MOS pair

$$L(\Delta\omega)=10\log\left[\frac{k_BT}{RA_{tank}^2C^2(\Delta\omega)^2}(1+\gamma_n)\right]$$

$$F=1+\gamma_n$$



- Transistors appear only through channel noise factor γ_n
- Transistor phase noise always proportional to tank noise (60% from tank, 40% from MOS pair, if γ_n = 2/3)
- This is because: 1) transistor noise is proportional to commutation time, 2) which is inversely proportional to the oscillation amplitude, 3) which is proportional to the tank parallel resistance
- A simple-minded LTI analysis would yield very wrong predictions (i.e., MOS phase noise increases with MOS g_m)

MOS phase noise – invariance

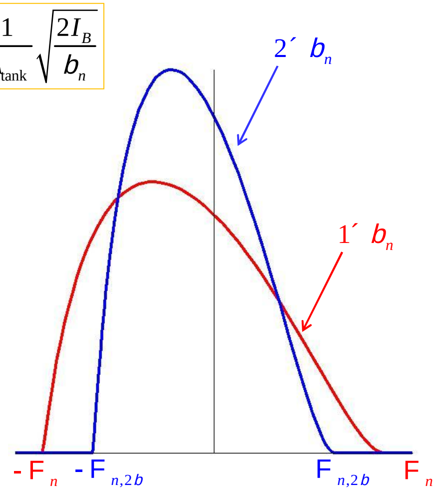
$$\overline{i_{N1}^2}(f)=4k_BT\gamma_n g_{m,N1}(f)$$

$$\Phi_n=\frac{1}{A_{tank}}\sqrt{\frac{2I_B}{\beta_n}}$$

Two effects balance each other:

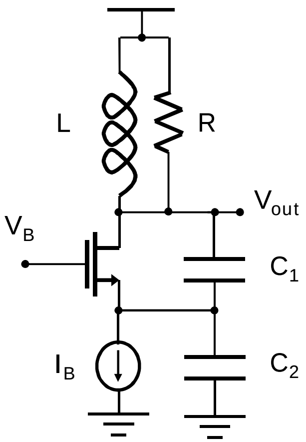
1) Larger MOS produces more noise during current commutation, and
2) Larger MOS allows a faster commutation

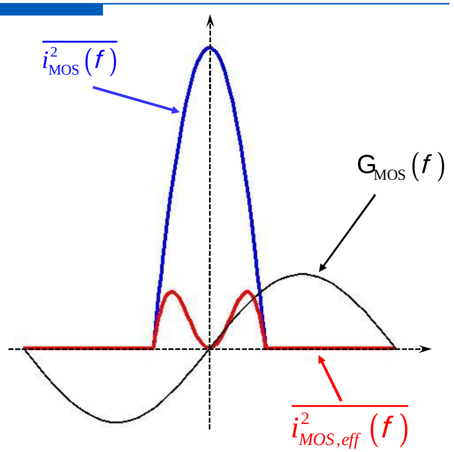
Result: the two areas are identical



Andreani *et al*., JSSC May '05

Single-ended Colpitts oscillator

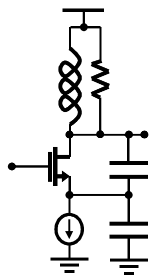




Noise injected into tank when ISF is near zero → excellent!

Hajimiri and Lee, JSSC Feb. '98; Andreani *et al*., JSSC May '05

Phase noise in Colpitts oscillator



$$L(\Delta\omega)\approx10\log\left[\frac{k_BT}{4I_B^2R^3(1-n)^2C^2(\Delta\omega)^2}\left(1+\gamma_n\frac{1-n}{n}\right)\right]$$

Minimum for:

$$n_{opt}=\begin{cases}0.30 & \text{for } \gamma_n=2/3\\ 1/3 & \text{for } \gamma_n=1\\ 0.36 & \text{for } \gamma_n=1.5\end{cases}$$

However, contrary to what was once (justifiably) believed, Colpitts is <u>more</u> noisy than the differential-pair LC oscillator!

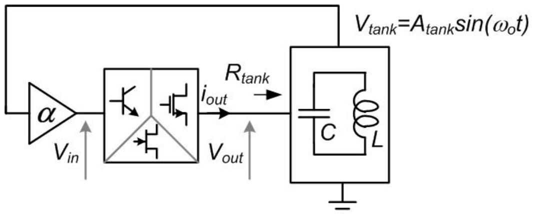
Andreani *et al*., JSSC May '05

Harmonic oscillators – a general result

1) G sinusoidal and in quadrature with tank voltage
2) Active devices work as transistors
3) Transistor current noise proportional to g_m

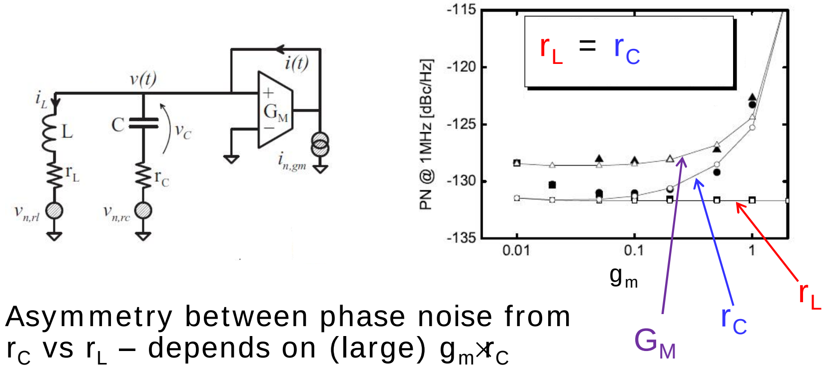
⬇

Transistor effective noise depends only on tank loss and topology



J. Bank, "A harmonic oscillator design methodology based on describing functions", PhD thesis, Gothenburg, Sweden, 2006
Mazzanti and Andreani, JSSC Dec. '08; Murphy et al, TCAS-I June '10

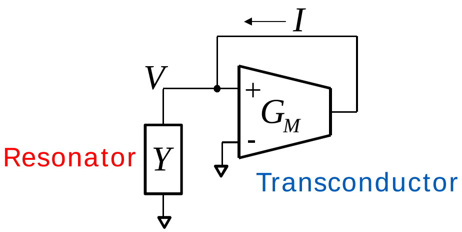
More on inductive vs capacitive losses



- Asymmetry between phase noise from r_C vs r_L – depends on (large) $g_m \times r_C$
- Lost with "equivalent" parallel tank losses
- PN from G_M always proportional to PN from r_C and r_L together (here, g = 1)

Pepe and Andreani, TCAS-II June 2016

Alternative phase noise analysis



Matrix-based Fourier-series LTV approach, starting from

$$\vec{I} = \mathbf{Y}\vec{V} \quad \text{and} \quad d\vec{I} = \mathbf{G_M}\, d\vec{V}$$

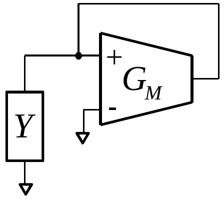
All quantities are functions of ω_0, $2\omega_0$, ... , $n\omega_0$

Pepe and Andreani, TCAS-I Feb. 2017

Results of new phase noise analysis

- Rigorous analysis under very broad hypotheses
  - G_M pure transconductance; Y linear; G_M noise proportional to G_M via g
- Phase noise from G_M always in proportion of g:1 to phase noise from Y, independently of resonator and transconductor nature
- Phase noise expressions as functions of V and Y
- Closed-form, explicit phase noise expressions if Q is high
  - General case of Y resonating at multiples of ω_0

What do these expressions look like?



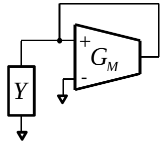
General equation

$$L(\Delta\omega)=10\log_{10}\left[2k_BT\frac{(1+\gamma)\,\omega_0^2}{\|\overrightarrow{\Delta V}\|^2}\,\frac{1}{\left|\overrightarrow{LV_1}^T\cdot\Delta\mathbf{Y}(\Delta\omega)\cdot\overrightarrow{RV_1}\right|^2}\left(\overrightarrow{LV_1}^T\cdot\mathrm{Re}(\mathbf{Y})\cdot\overrightarrow{LV_1}^*\right)\right]$$

Matrix algebra

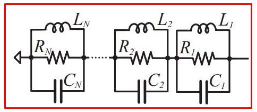
$$L(\Delta\omega)=10\log_{10}\left[2k_BT\frac{(1+\gamma)\,\omega_0^2}{\|\overrightarrow{\Delta V}\|^2}\,\frac{1}{\left|\overrightarrow{LV_1}^T\cdot\Delta\mathbf{Y}(\Delta\omega)\cdot\overrightarrow{RV_1}\right|^2}\left(\overrightarrow{LV_1}^T\cdot\mathrm{Re}(\mathbf{Y})\cdot\overrightarrow{LV_1}^*\right)\right]$$

$$M_0\cdot\overrightarrow{\Delta V}=\vec{0};\qquad \overrightarrow{RV_1}=\overrightarrow{\Delta V}\big/\|\overrightarrow{\Delta V}\|;\qquad \overrightarrow{LV_1}^T\cdot M_0=\vec{0}^T$$



$$M_0=\begin{bmatrix}\ddots & \vdots & \vdots & \vdots & \vdots & \vdots & ⋰\\ \cdots & Y_2^*-g_0 & -g_1^* & -g_2^* & -g_3^* & -g_4^* & \cdots\\ \cdots & -g_1 & Y_1^*-g_0 & -g_1^* & -g_2^* & -g_3^* & \cdots\\ \cdots & -g_2 & -g_1 & Y_0-g_0 & -g_1^* & -g_2^* & \cdots\\ \cdots & -g_3 & -g_2 & -g_1 & Y_1-g_0 & -g_1^* & \cdots\\ \cdots & -g_4 & -g_3 & -g_2 & -g_1 & Y_2-g_0 & \cdots\\ ⋰ & \vdots & \vdots & \vdots & \vdots & \vdots & \ddots\end{bmatrix}$$

Tank with multiple resonances



n high-Q resonances

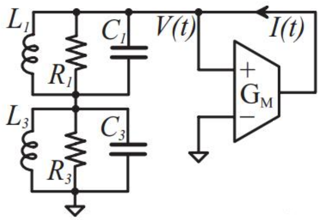
$$L(\Delta\omega)=10\log_{10}\left[\frac{k_BT(1+\gamma)}{\Delta\omega^2}\,\frac{\sum_1^N n^2A_n^2/R_n}{\left(\sum_1^N n^2A_n^2C_n\right)^2}\right]$$

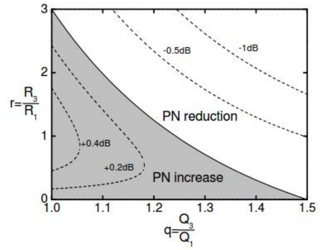
Single resonance → we recover the well-known equation

$$L(\Delta\omega)=10\log_{10}\left[\frac{k_BT(1+\gamma)}{\Delta\omega^2}\,\frac{1}{A_{pk}^2C^2R}\right]=10\log_{10}\left(\frac{2k_BTR}{A_{pk}^2/2}\left(\frac{1}{2Q}\cdot\frac{\omega_0}{\Delta\omega}\right)^2\right)$$

Tank resonating at ω_0 and $3\omega_0$

$$L(\Delta\omega)=10\log_{10}\left[\frac{k_BT(1+\gamma)}{\Delta\omega^2}\,\frac{A_1^2/R_1+9A_3^2/R_3}{\left(A_1^2C_1+9A_3^2C_3\right)^2}\right]$$





- Steeper V-waveform, but now also R_3 contributes
- Advantageous only if Q_3 (much) larger than Q_1
- Difficult to enforce in practice

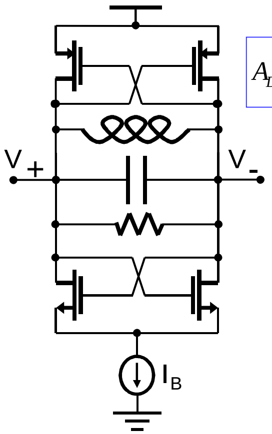
Pepe and Andreani, TCAS-I Feb. 2017; Garampazzi et al., JSSC Mar. 2014

Double-switch pair vs. single-switch pair

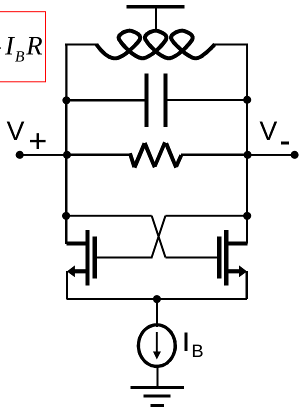
Double-switch (DS) pair oscillator

Single-switch (SS) pair oscillator

$A_{DS} = \frac{4}{\pi} I_B R$



$A_{SS} = \frac{2}{\pi} I_B R$



What phase noise difference should we expect?

DS pair vs. SS pair – phase noise

$$L_{DS}(\Delta\omega) = 10\log\left(\frac{2k_B TR}{A_{DS}^2/2}\left(\frac{1}{2Q}\frac{\omega_0}{\Delta\omega}\right)^2\left(1+\frac{\gamma_n+\gamma_p}{2}\right)\right)$$

$$L_{SS}(\Delta\omega) = 10\log\left(\frac{2k_B TR}{A_{SS}^2/2}\left(\frac{1}{2Q}\frac{\omega_0}{\Delta\omega}\right)^2(1+\gamma_n)\right)$$

- 60% from tank, 40% from transistors
  - If $\gamma_n = \gamma_p = 2/3$
- If $I_{B,DS} = I_{B,SS}$ and $\gamma_n = \gamma_p$ →

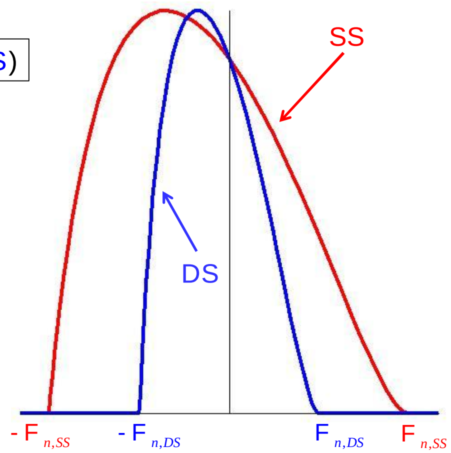
$$A_{DS} = 2A_{SS} \Rightarrow L_{DS} = L_{SS} - 6dB \quad (!)$$

Andreani and Fard, JSSC Dec. 2006

DS vs. SS – MOS noise

Area (SS) = 2 · Area (DS)

4 DS transistors make as much noise as 2 SS transistors!



Phase noise vs power consumption

- If $I_{B,SS} = 2I_{B,DS}$ → $A_{SS} = A_{DS}$ → same phase noise
- In this case, $V_{DD,SS} = ½\ V_{DD,DS}$ → same power consumption
- Thus, same maximum achievable "figure-of-merit" (FoM, phase noise for a given power consumption)
- If $V_{DD,DS} = V_{DD,SS}$ and $I_{B,DS} = ¼\ I_{B,SS}$ → $A_{SS,max} = 2A_{DS,max}$ → $L_{SS,min} = L_{DS,min} - 6dB$
- Again, $FoM_{SS,max} = FoM_{DS,max}$

Figure of merit (FoM)

Phase noise normalized to power consumption, oscillation frequency, and frequency offset

power efficiency

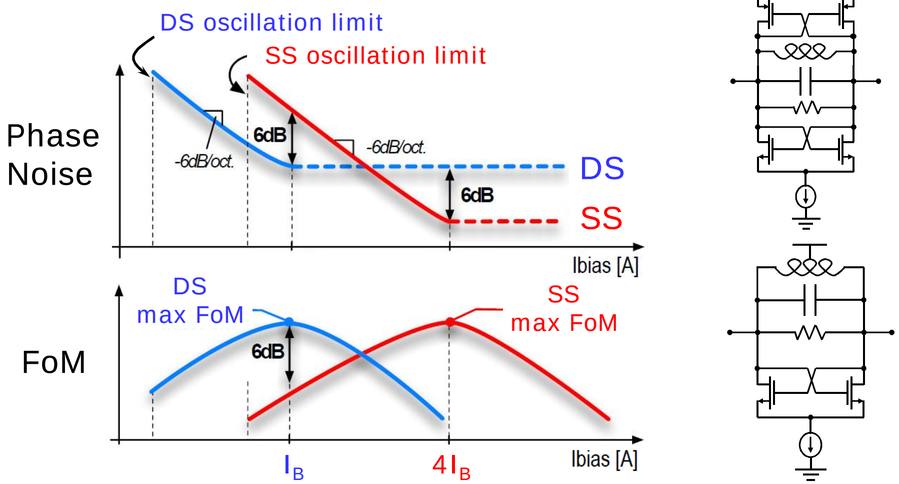
$$FoM = \frac{(\omega_0/\Delta\omega)^2}{L(\Delta\omega)\times P_{DC[mW]}} = 10^{-3}\times\frac{2Q^2}{k_B T}\times\frac{\eta}{F}$$

noise factor

High tank Q crucial for high FoM

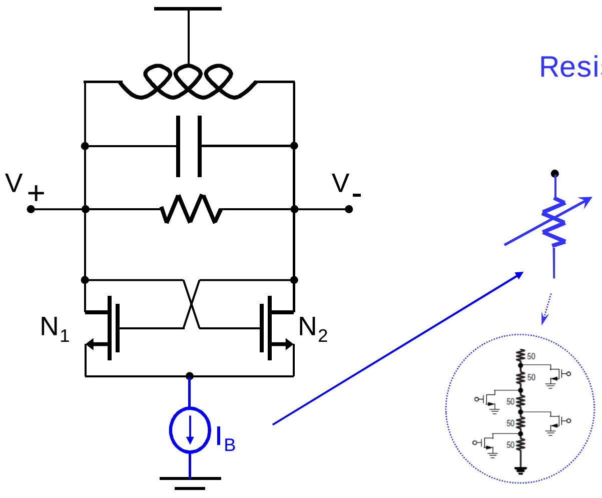
Andreani *et al*., JSSC Dec. 2006; Garampazzi *et al*., JSSC Jul. 2015; Murphy *et al*., ISSCC 2015

SS vs DS – PN and FoM with fixed V_{dd}



Liscidini *et al*., ISSCC 2012, JSSC Mar. 2014

Current bias – resistive source



Resistor

Very simple bias circuitry

No 1/f noise generation

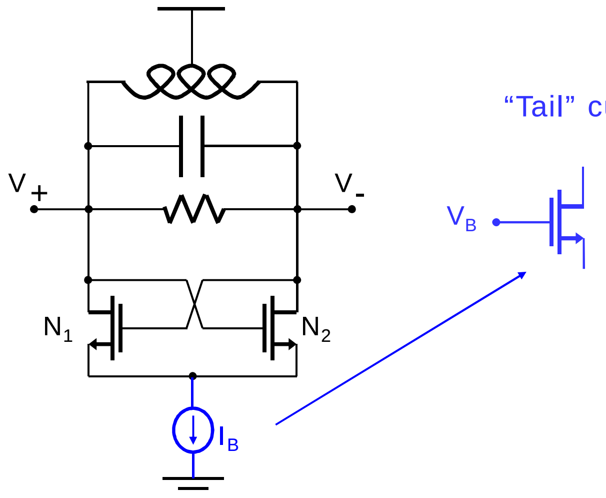
Low-Z source

Significant upconversion of 1/f noise from N_1-N_2

Abidi and Ismail, ISSCC 2003

Current bias – MOS source

“Tail” current source

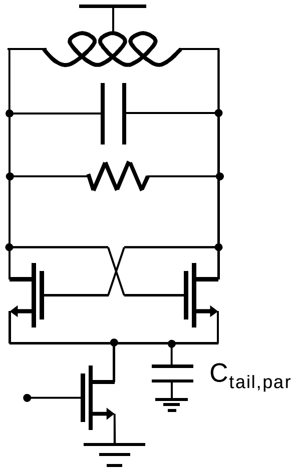


High-Z source

Less upconversion of 1/f noise from N_1-N_2

Own 1/f noise generation

Impact of parasitic tail capacitance

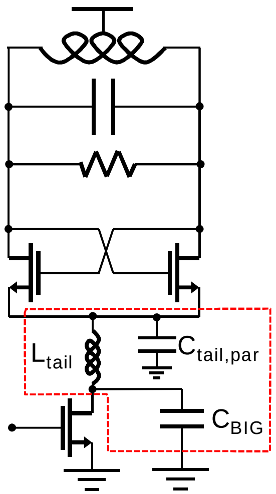


- $C_{tail,par}$ + cross-coupled MOS entering linear region à MOS contribution to phase noise increases, even by a large amount
- $C_{tail,par}$ good for filtering HF noise from bias
- But, increase of 1/f noise upconversion

Overview

- Popular harmonic oscillators
  - Phase noise
- Architectures for low $1/f^2$ and/or $1/f^3$ phase noise
- Series-resonance oscillator
- Design techniques for very wide frequency tuning range RF CMOS VCOs

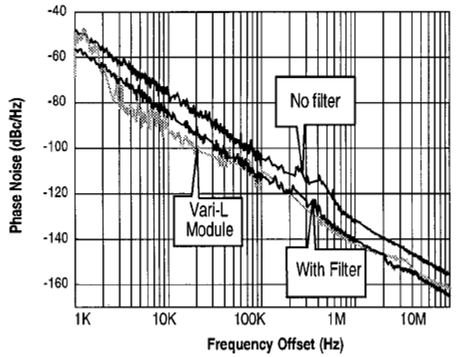
Possible solution – noise filter



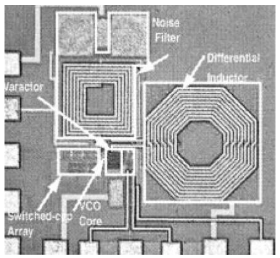
- Noise filter: $C_{tail,par}$ resonates with L_{tail} at $2w_0$ à MOS switches see high-Z at $2w_0$
- C_{BIG} filters tail noise and ac-grounds L_{tail}
- C_{BIG} includes C_{DB} of MOS tail à long and large MOS, low 1/f noise
- Drawbacks: narrow-band, $C_{tail,par}$ must be known with some precision, extra L_{tail}

Hegazi, Sjöland, Abidi, JSSC Dec. 2001

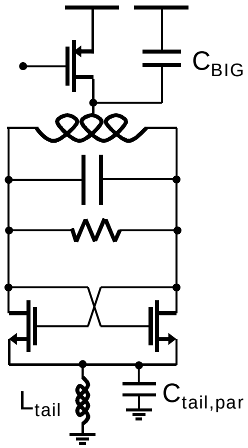
Dramatic performance improvement



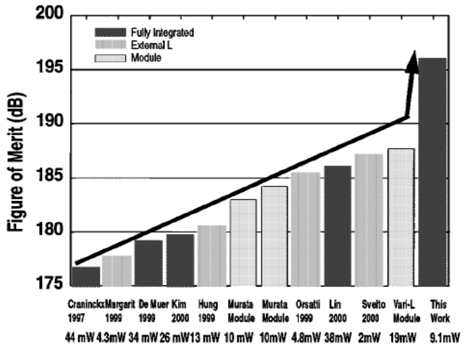
- 0.35µm CMOS
- 1.2GHz, 3.5mA, 2.5V
- L_{tail} = 10nH, C_{BIG} = 40pF
- FoM = 196dBc/Hz
- TR?



More on tail filter

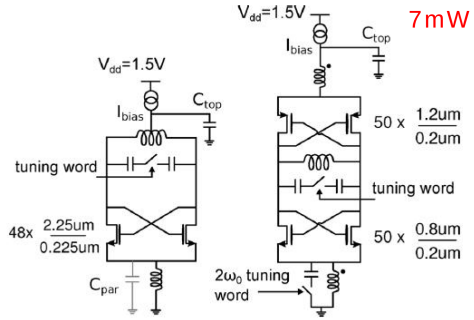


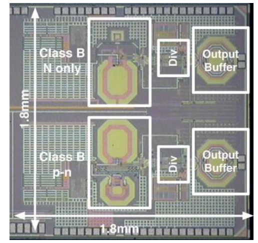
- Many variations on the same basic theme
- Extremely popular!

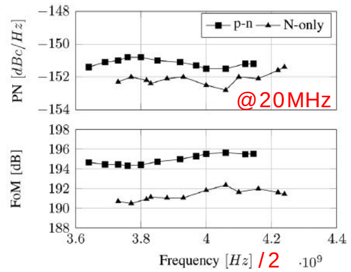


Hegazi, Sjöland, Abidi, JSSC Dec. 2001

A recent variation



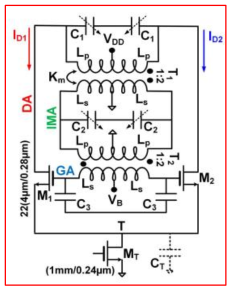
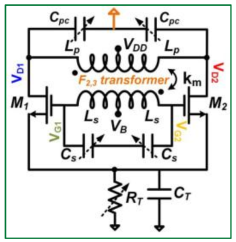
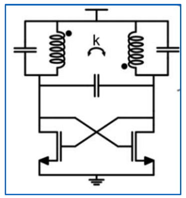




- Tail/top resonance with Xrfm
- Very low PN and great FoM (up to 195.6 dBc/Hz)
- 200-400 kHz $1/f^3$ corner

Garampazzi *et al.*, JSSC July 2015

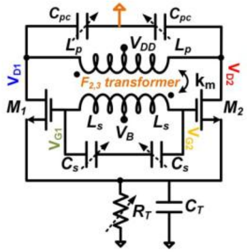
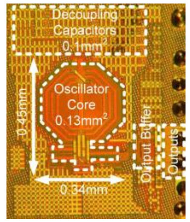
Alternative to tail resonance

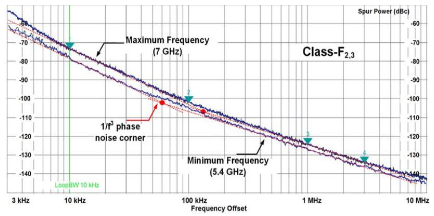




- Design tank for differential resonance at w_0 and common-mode resonance at $2w_0$
  - Also here, the $2w_0$ resonance must track the w_0 resonance – two capacitor banks

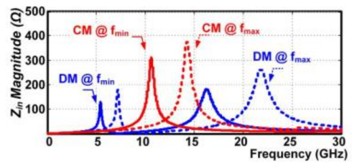
Babaie *et al.*, RFIC 2013, JSSC Mar. 2015; Shahmohammadi *et al.*, ISSCC 2015; Murphy *et al.*, ISSCC 2015

Class-F_2 (or, here, $F_{2,3}$) oscillator

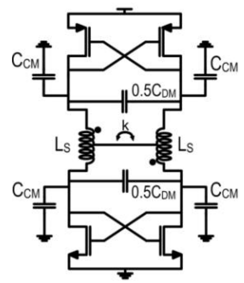


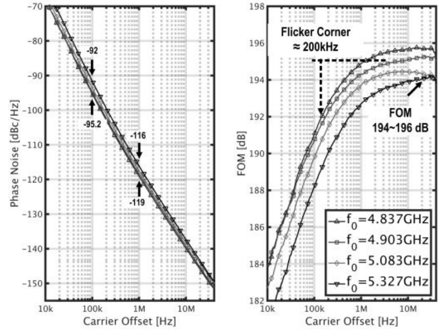
- 5.4-7.0 GHz; 1 V, 10-12 mW
- Low $1/f^3$ corner (60-130 kHz)
- Very good FoM (~191 dBc/Hz) at very low PN (-124 dBc/Hz @ 1 MHz)
- Very low V_{DD} pushing (12-23 MHz/V)



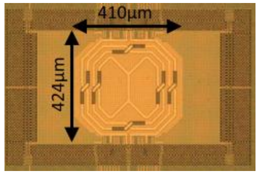
Shahmohammadi *et al.*, ISSCC 2015

Implicit common-mode resonance



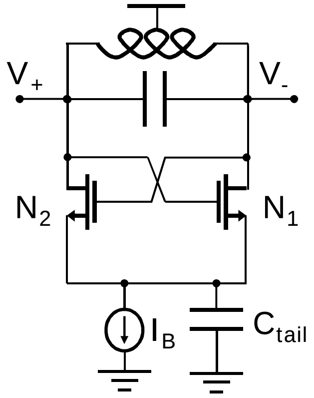


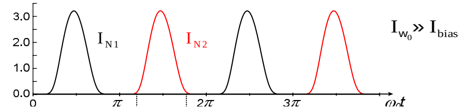
- 4.7-5.4 GHz; 0.7 V, 0.5 mW
- Low $1/f^3$ corner (200 kHz)
- Great FoM (194-196 dBc/Hz)



Murphy *et al.*, ISSCC 2016

A totally different approach – class-C



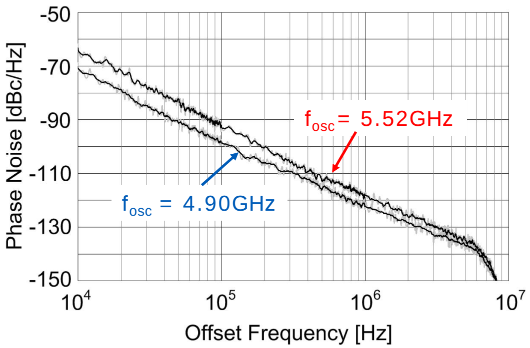


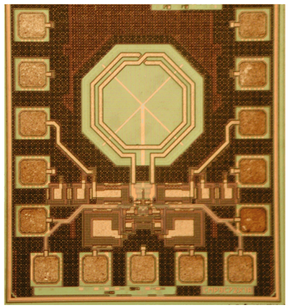
- C_{tail} turns class-B into class-C: optimal differential “Colpitts” oscillator
- Ideally, 3.9 dB lower phase noise for the same bias current
- Also here, C_{tail} filters off high-frequency noise from tail, and includes tail C_{DB} à long and large MOS, low 1/f noise

Mazzanti and Andreani, JSSC Dec. 2008

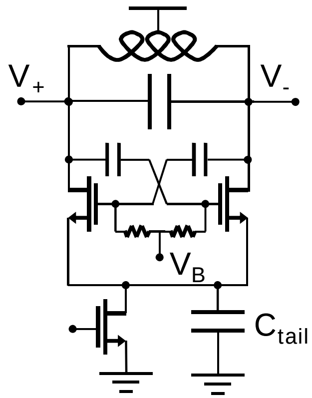
Original prototype

- 4.90 GHz < f_c < 5.65 GHz
- 1 V, 1.4 mW
- 193.5 dBc/Hz < FoM < 196 dBc/Hz





Design issues in class-C CMOS VCO

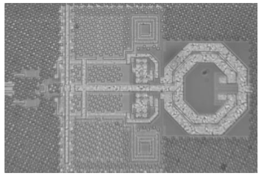


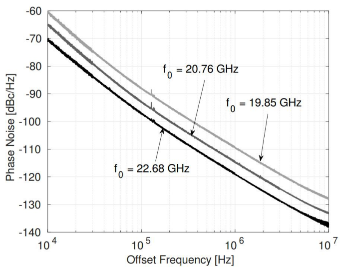
- Diff-pair must not enter the linear region (otherwise, large PN penalty) à shift of MOS DC gate voltage V_B (which may be generated with feedback loop)
- RC bias should not load tank (transformer feedback possible)
- Nevertheless, higher maximum oscillation amplitude in the *ideal* class-B CMOS oscillator
- Class-C very attractive for BJT VCOs

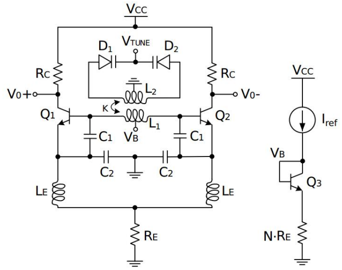
Mazzanti and Andreani, JSSC Dec. 2008; Fanori and Andreani, JSSC July 2013

Colpitts VCO in SiGe BiCMOS process

- 18.8-23.1GHz; 4.0V, 17.5mA
- PN = -119dBc/Hz @ 1MHz (best)
- FoM = 188dBc/Hz





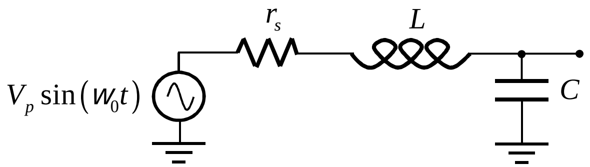


Boscolo *et al.*, ESSCIRC 2017

Overview

- Popular harmonic oscillators
  - Phase noise
- Architectures for low $1/f^2$ and/or $1/f^3$ phase noise
- Series-resonance oscillator
- Design techniques for very wide frequency tuning range RF CMOS VCOs

Oscillation with series resonance

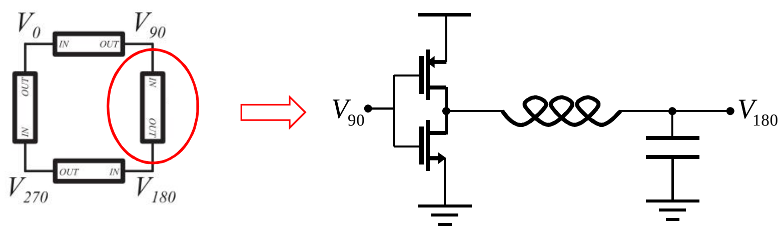


$Q \cdot V_p \sin\left(\omega_0 t - \frac{\pi}{2}\right)$

- Voltage driven
- Gain equal to quality factor à internal oscillation may be much higher than V_{DD}
  - Attractive for ultra-low phase noise
- $\pi/2$ phase shift between input and output

P. Andreani, L. Fanori, and T. Mattsson, “Series-resonance oscillator,” U.S. Patent 2015 0381 157, 2015

Phase shift by quadrature



- We disregard the (important) issue of start-up
- Square wave between V_{DD} and GND at LC input
- MOS devices work almost exclusively as switches à channel resistance in series with the tank’s

Tohidian *et al.*, MWCL Aug. 2015; Pepe, Bevilacqua, Andreani, TCAS-I Feb. 2018

Phase noise

$$L(\Delta\omega) = 10\log_{10}\left[\frac{4k_BTr_s}{(4I_{pk}L\,\Delta\omega)^2}(1+F)\right] = 10\log_{10}\left[\frac{4k_BT}{(4I_{pk}\Delta\omega)^2 r_s}\left(\frac{1}{Q}\frac{\omega_0}{\Delta\omega}\right)^2(1+F)\right]$$

$$I_{pk} = \frac{2}{\pi}\frac{V_{DD}}{r_s}$$

- MOS work as switches → previous phase noise theorems do not apply
- *F* accounts for 1) MOS are non-ideal switches, and 2) they do work as transconductors for a (tiny) fraction of the oscillation period
- Ideally, *F* is negligible!

Pepe, Bevilacqua, Andreani, TCAS-I Feb. 2018

Figure of merit

$$FoM = 10^{-3}\times\frac{2Q^2}{k_BT}\times\frac{\eta}{1+F}$$

Ideally, close to 1 (η)

Ideally, close to 0 (F)

- Usual dependence on Q^2
- Very large power consumption, ultra-low phase noise (plus quadrature phases for free)
- However:
  - MOS resistance is critical (current-based architectures such as class-B and class-C are much more robust)
  - Stray resistances of GND and power supply distribution are also critical
  - Very large internal voltages make frequency tuning difficult

Overview

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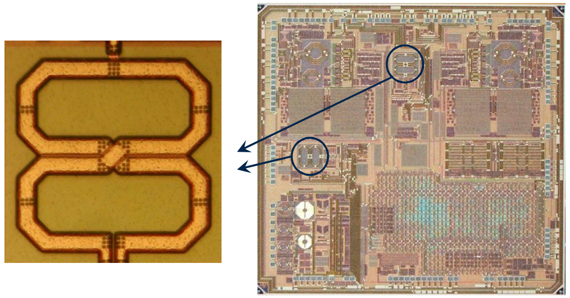
VCOs in modern radios – I

- Carrier aggregation requires several harmonic VCOs
  - Active at the same time
  - Should not pull one another
- Band proliferation favors VCOs with a very wide tuning range (TR)
  - Wider than 1 octave is particularly attractive

VCOs in modern radios – II

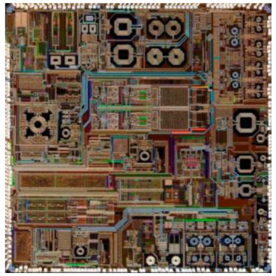
- VCO with 8-shaped tank inductor
  - Much less sensitive to external magnetic fields
  - Generates itself a vanishing magnetic field
  - Slightly lower Q acceptable
  - Often used



M. Nilsson *et al.*, ISSCC 2011

Very-Wide-TR VCOs – I

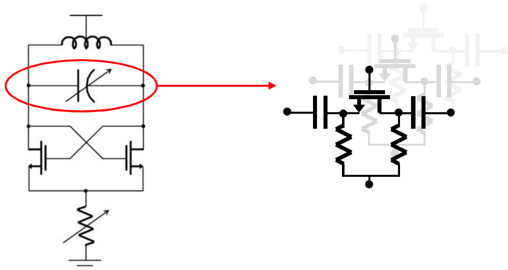
- Two or more VCOs with overlapping TRs
  - Saves power, costs area
  - Very popular choice in real-life products



Hadjichristos *et al.*, ISSCC 2009

Very-Wide-TR VCOs – II

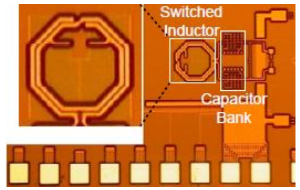
- Large switchable C in parallel to small L
  - floating switches
  - power wasted at low frequencies, compared to reasonable phase-noise specs
  - power cannot be decreased without killing the oscillation

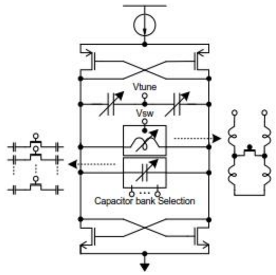


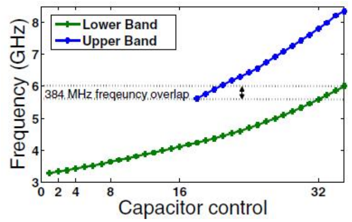
Sjöland, TCAS-II May 2002

Very-Wide-TR VCOs – III

- Switchable L
  - Ultra-wide TR possible
  - Difficult to obtain low PN at high FoM
  - Additional issue: switchable 8-shaped inductor



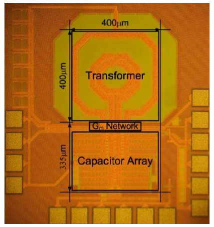


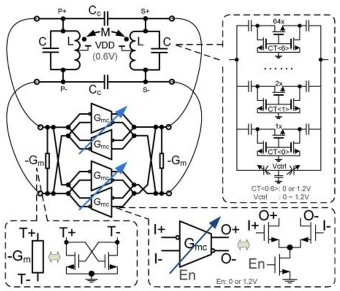


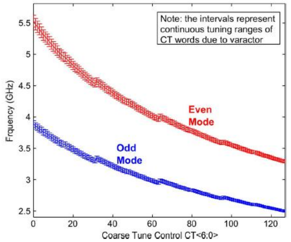
Sadhu *et al.*, CICC 2009

Very-Wide-TR VCOs – IV

- Transformer-based VCOs
  - Two resonances with overlapping TRs
  - TR > 1 octave
  - Difficult to design an 8-shaped transformer



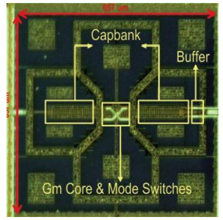


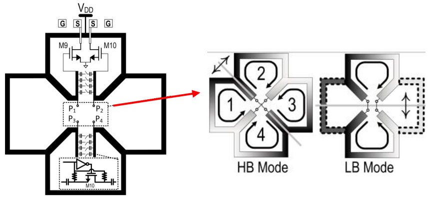


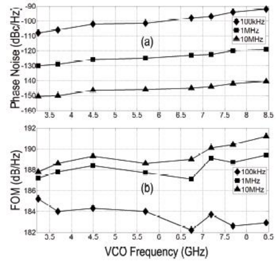
Bevilacqua *et al*., TCAS-II Apr. 2007; Li *et al*., JSSC June 2012

Very-Wide-TR VCOs – V

- Mode-switching VCO
  - 4 inductors, two oscillation modes
  - Rejects external magnetic fields
  - TR > 1 octave
  - Excellent PM and FoM
  - Large area



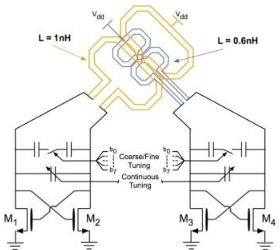


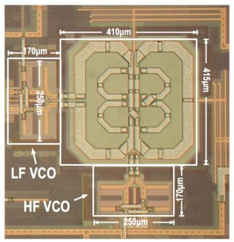


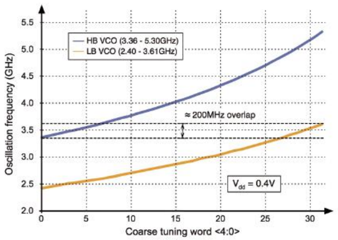
Taghivand *et al*., ISSCC 2014

Very-Wide-TR VCOs – VI

- Double-core VCO
  - Two concentric 8-shaped coils – do not interfere (much) with each other
  - TR > 1 octave; saves inductor area, sub-optimal Q







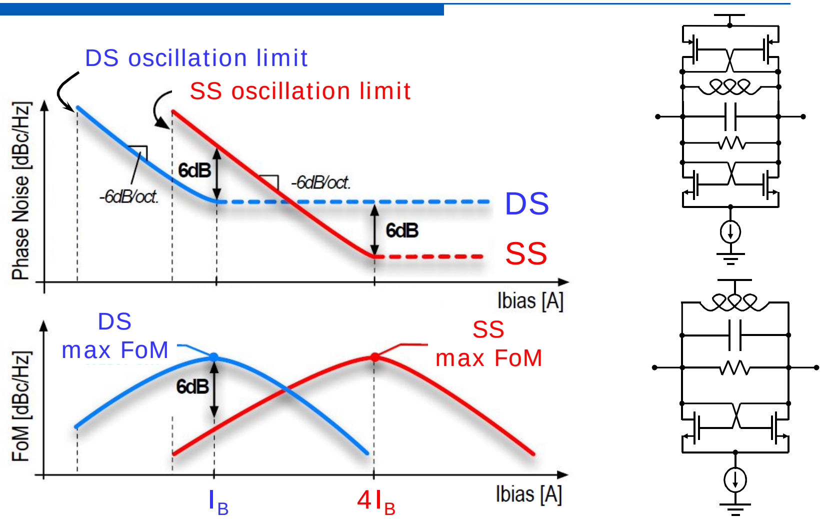
Fanori *et al*., ISSCC 2014

Very-Wide-TR VCOs – VII

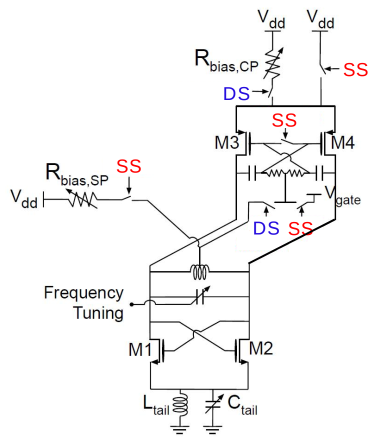
- Reconfigurable active core
  - Standard LC tank design (i.e., with very large capacitance)
  - Negative resistance: either single-switch (nMOS) pair – SS mode
  - or, double (complementary nMOS-pMOS) switch pair – DS mode
  - DS mode avoids power waste at lower frequencies

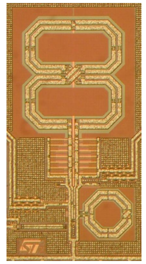
Liscidini *et al*., ISSCC 2012, JSSC Mar. 2014

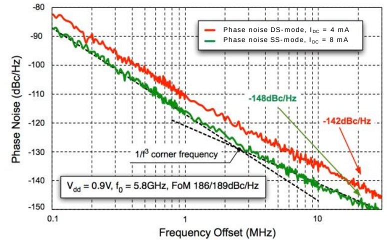
SS pair vs. DS pair, again



Very-Wide-TR reconfigurable VCO







- STM 28nm UTBB FD-SOI CMOS
- 2.8-5.8 GHz
- -154 < PN (dBc/Hz @ 20MHz) < -142
- 186 < FoM (dBc/Hz) < 189
- 300 kHz < $1/f^3$ corner < 3 MHz

Fanori *et al.*, RFIC 2015

Conclusions

- Rigorous phase noise results
  - For transconductor-based oscillators
- Class-B VCOs à simple, robust, ubiquitous
  - Tail filter improves phase noise, even largely
  - Recent proposals: common-mode tank resonance at $2w_0$
- Class-C VCOs à better efficiency than standard class-B, but more complicated
  - Class-C must be enforced for all working conditions
  - Excellent for BJT VCOs
- Series-resonance oscillator à great potential, but important issues to be solved
- Several techniques for very wide frequency tuning range
  - None is a clear winner

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