Digital Calibration for RF Transceivers

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Acknowledgements

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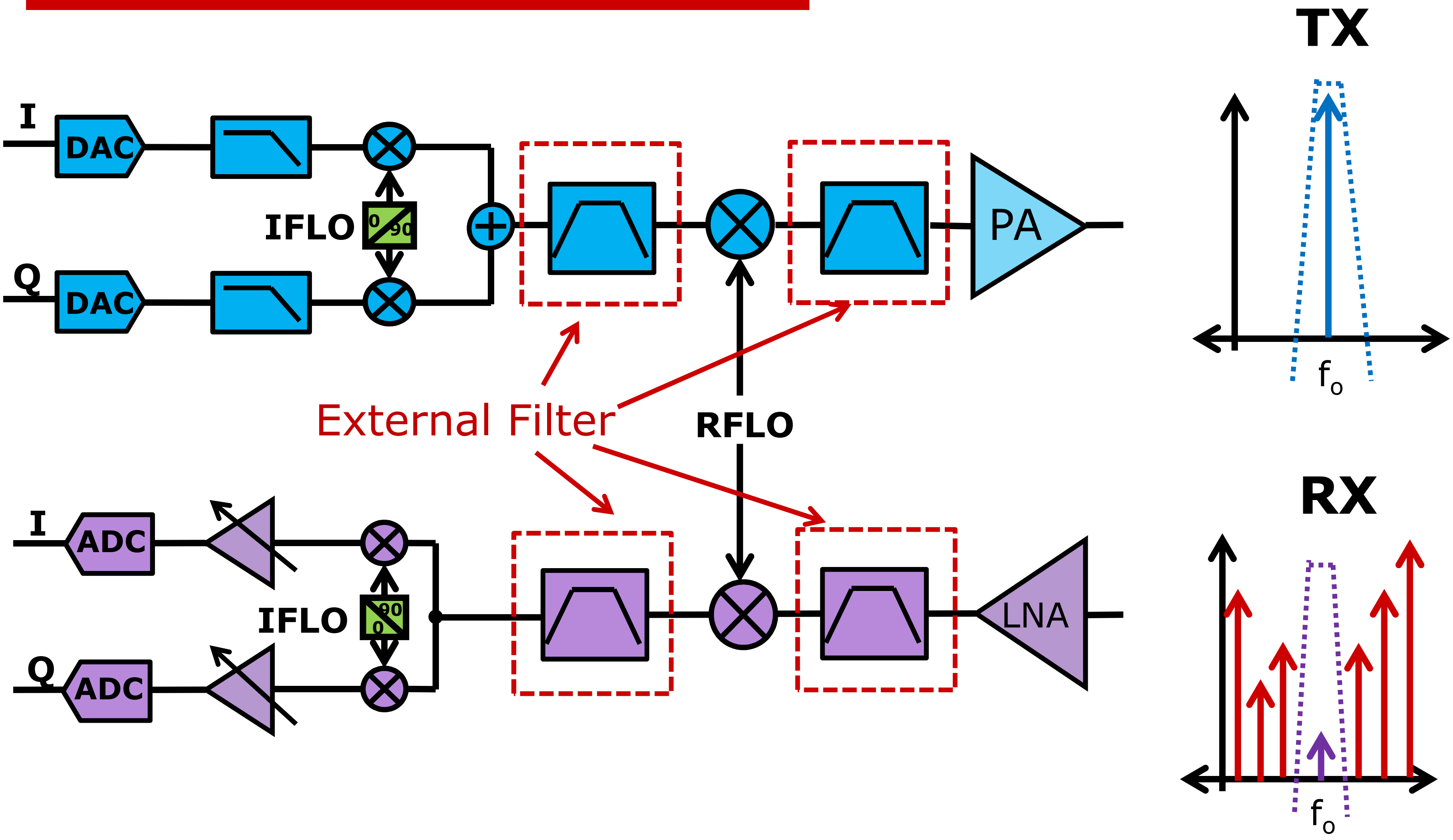
- Thanks to my Mediatek colleagues for their assistance and advice including Levine Chen, Shawn Chen, JC Tsai, Yuli Hsueh, LH Shen, Tiku Yu, CH Liao, Greg Wu, Benson Chen, and George Chien.

Outline

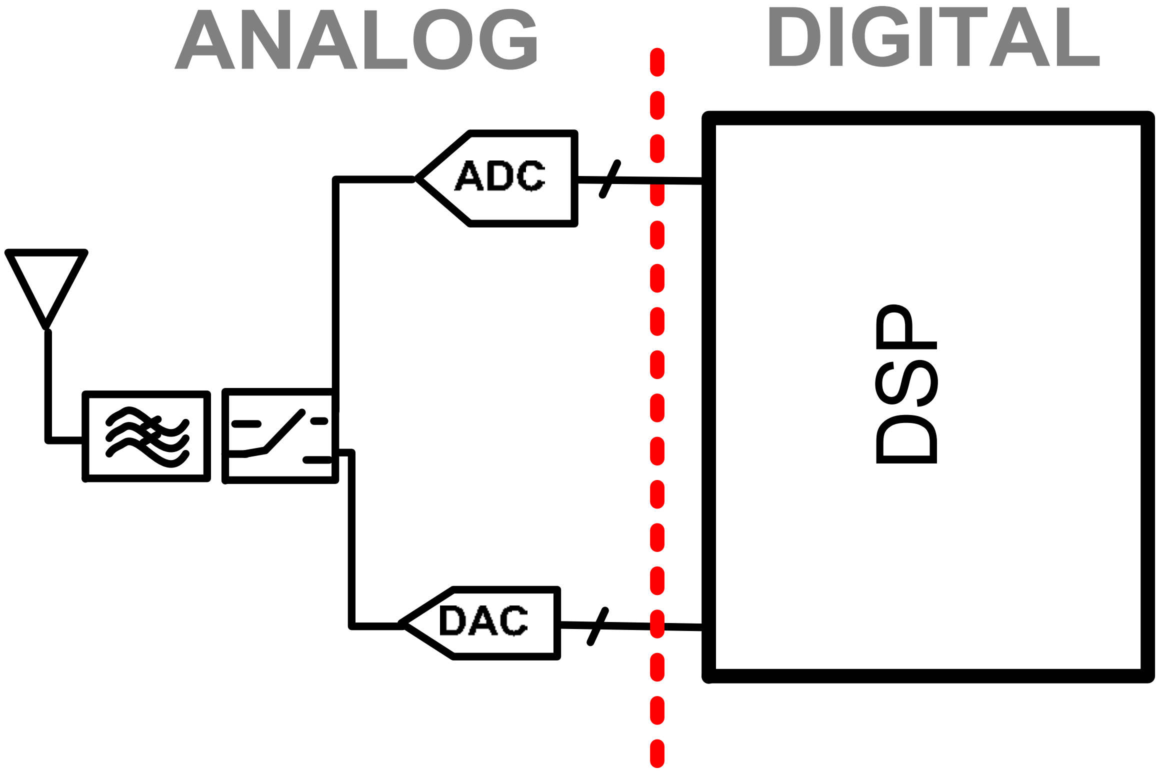
- Modern Transceiver Architecture
  - Direct Conversion Transceiver
  - Transceiver Building Blocks and their Impairments
  - Motivation for Digital Calibration

- Digital Calibration Techniques to Address
  - Device Mismatch
    - DC Offset Cancellation
    - IQ and LOFT Calibration
  - Process Variation
    - Analog Filter BW Calibration
    - Gain Calibration
    - VCO LC Tank Calibration
  - Non-Linearity
    - IMD Cancellation, PA Pre-Distortion

Super-Heterodyne Radio

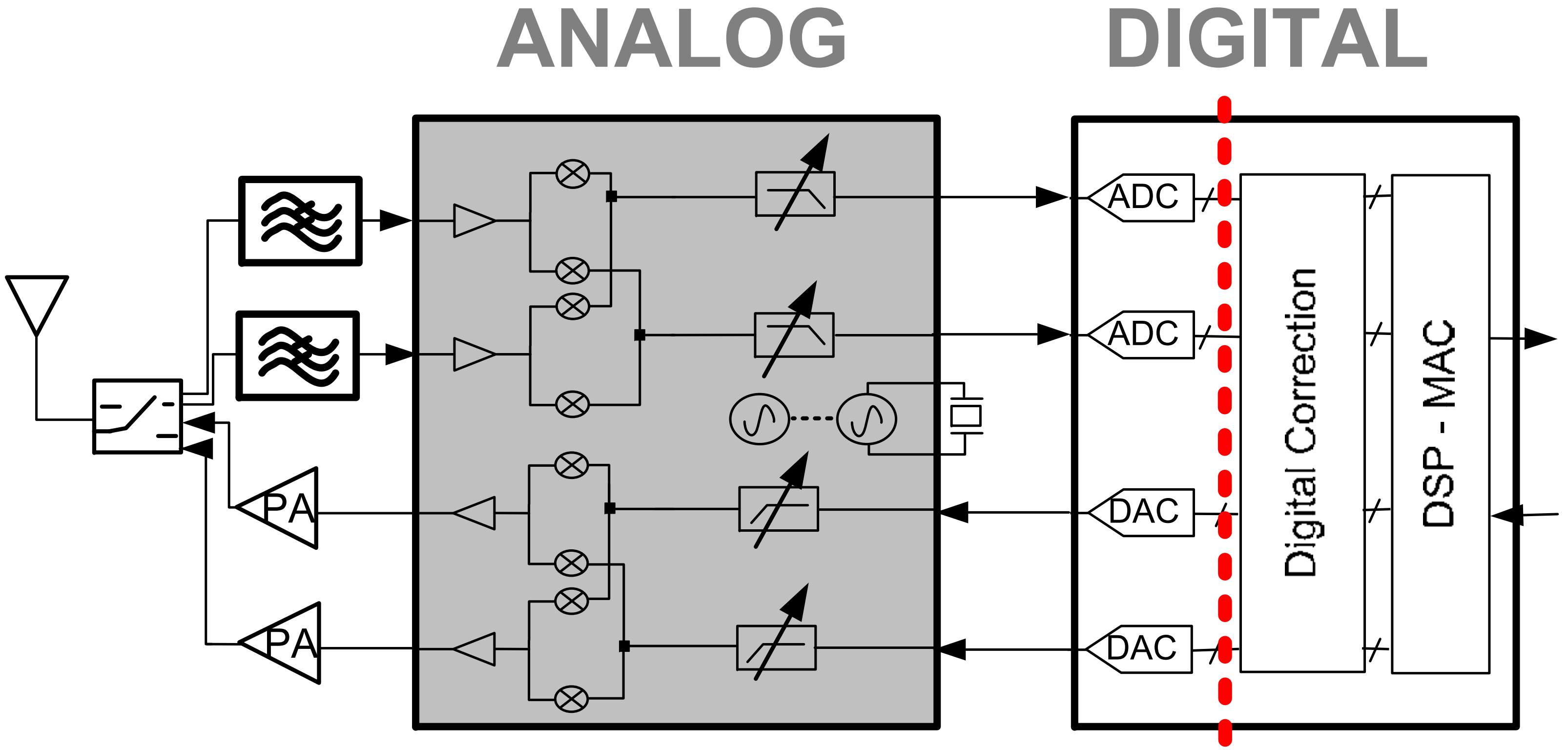


The Holy Grail



Ideal: Antenna to A/D not really possible (yet ??)

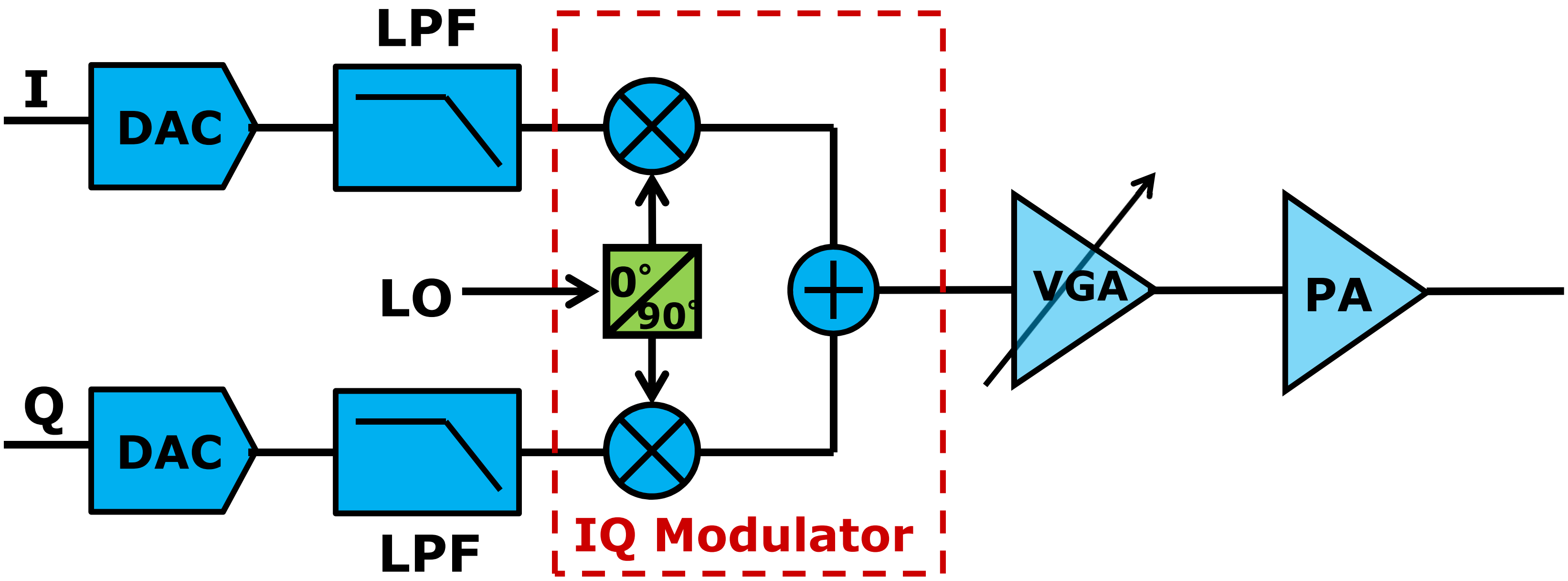
Direct Conversion Transceiver



- Next Best : Direct conversion (zero-IF or low-IF)
- Requires RF frequency IQ up/down-conversion
- Correct imperfections in digital domain

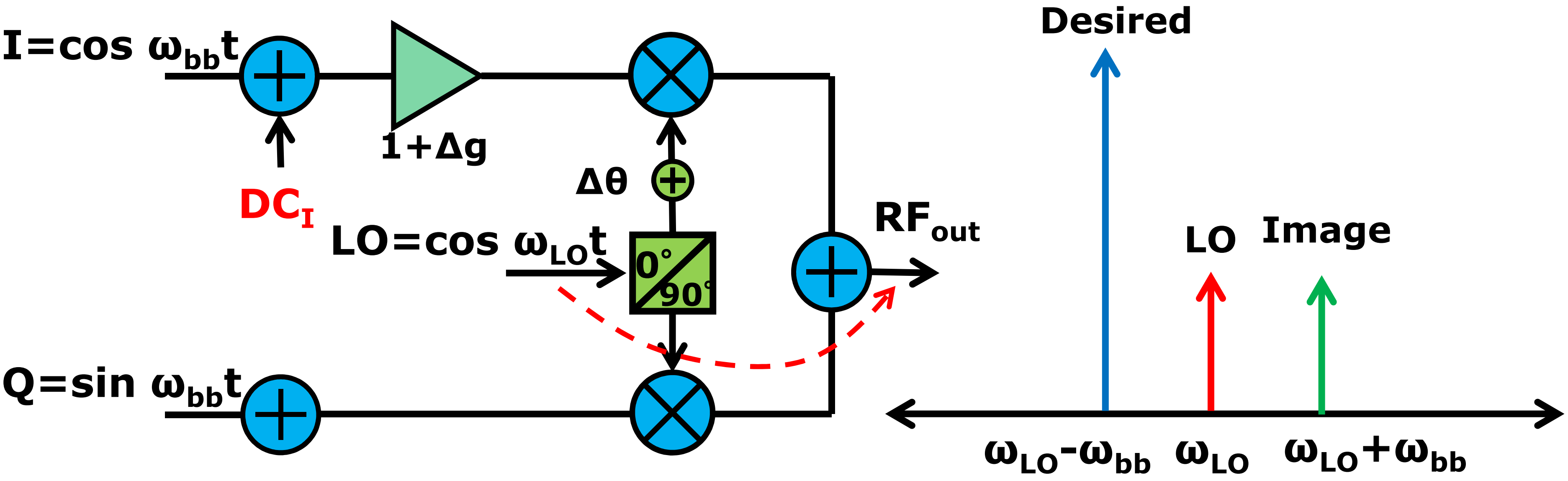
[Vassiliou, JSSC 03]

Direct Conversion Transmitter



- Direct Conversion
  - High Frequency IQ Up-Conversion
  - On-chip filter subject to process variation and mismatch
  - Mismatches cause LO leakage and image signal at RF
  - TX non-linearity degrades EVM and spectral mask

IQ Modulator Impairments



- Using single-tone analysis

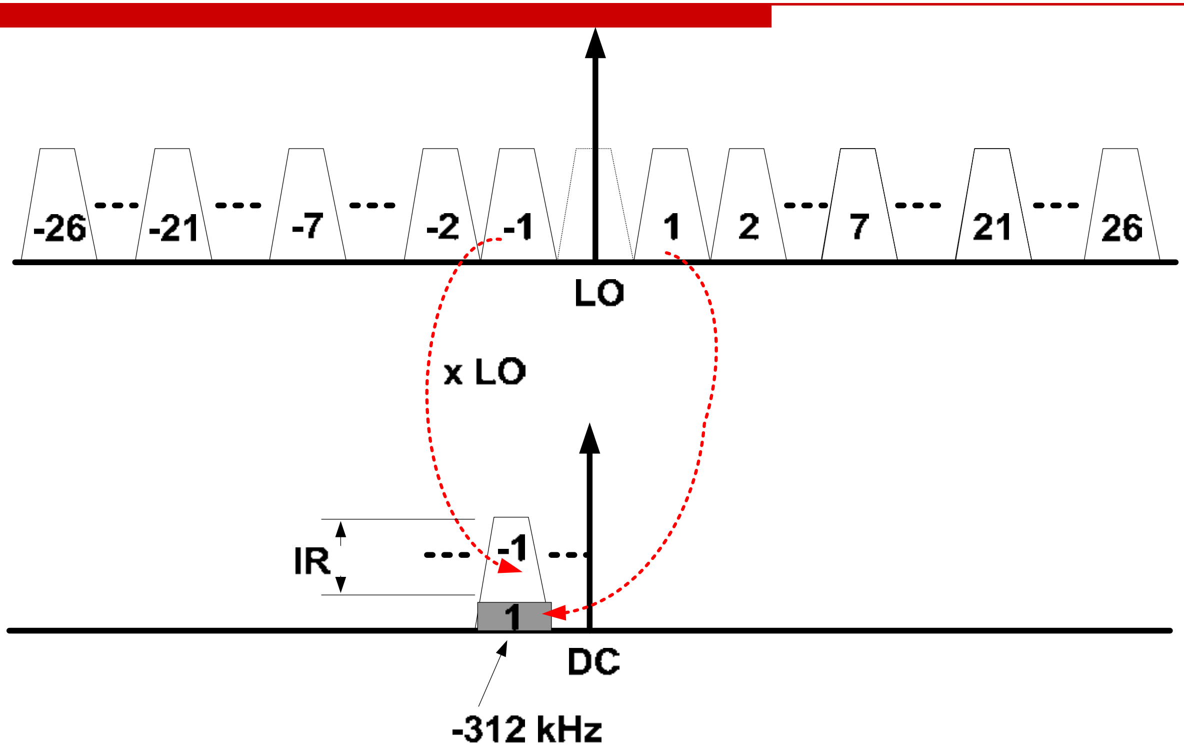
$$\text{RFout} = \frac{1}{2}[1+(1+\Delta g)\cos\Delta\theta]\cos(\omega_{bb}-\omega_{LO})t-\frac{1}{2}[(1+\Delta g)\sin\Delta\theta]\sin(\omega_{LO}-\omega_{bb})t$$

$$\text{Image}\ \left\{\quad +\frac{1}{2}[-1+(1+\Delta g)\cos\Delta\theta]\cos(\omega_{bb}+\omega_{LO})t-\frac{1}{2}[(1+\Delta g)\sin\Delta\theta]\sin(\omega_{LO}+\omega_{bb})t\right.$$

$$\text{LO}\ \left\{\quad +[DC_I]\cos(\omega_{LO})t\right.$$

[Razavi, RF Microelectronics]

TX/RX IQ Mismatch (OFDM)

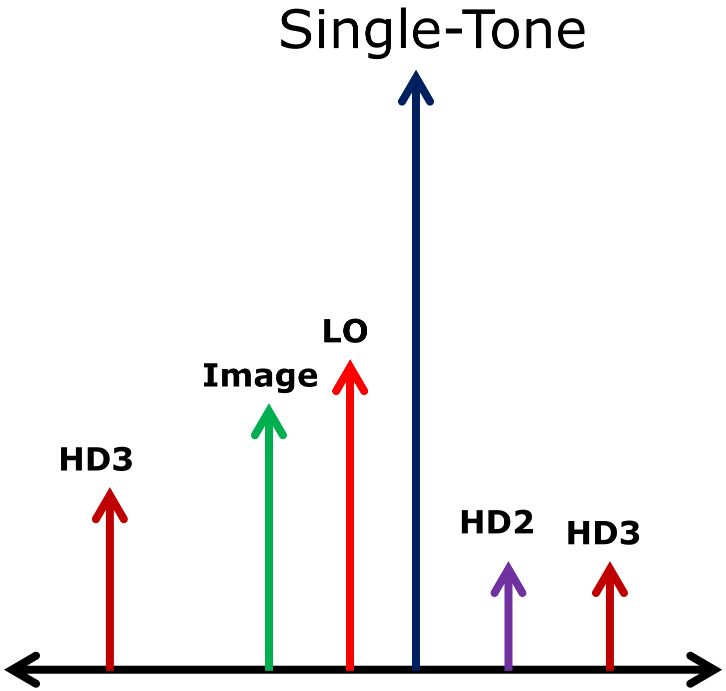


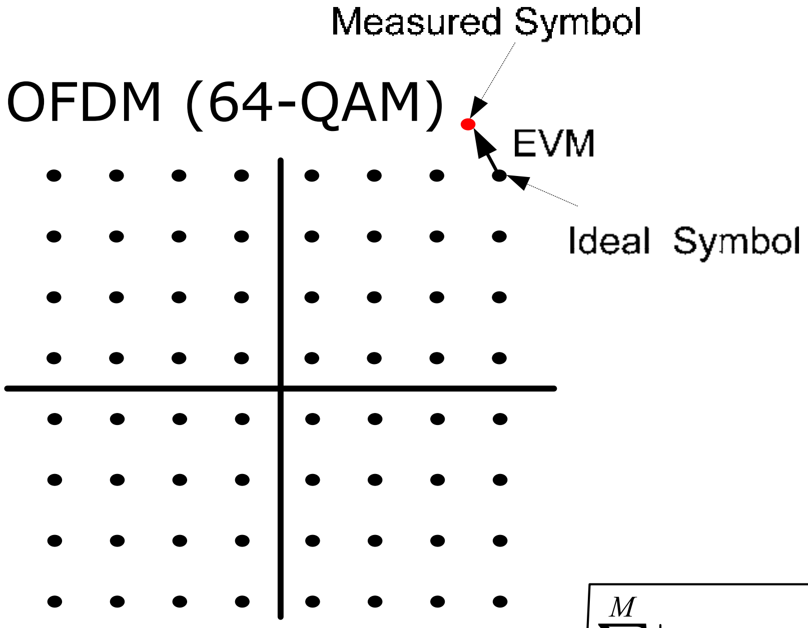
- OFDM receivers/transmitters
  - Each sub-carrier is modulated independently
  - IQ mismatch causes inter-sub-carrier interference
  - EVM Degradation

[Liu, IEEE Trans. 98]

Modulation Accuracy (EVM)

- EVM (Error Vector Magnitude) is a measure of the modulation accuracy of a transmitter





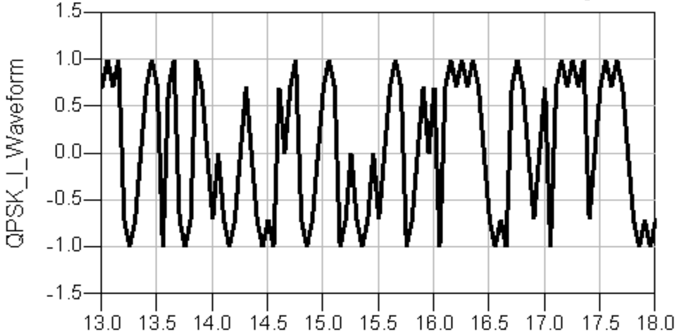
$$EVM = \sqrt{\frac{\sum_{i=1}^{M} \left| Z(i) - R(i) \right|^2}{\sum_{i=1}^{M} \left| R(i) \right|^2}}$$

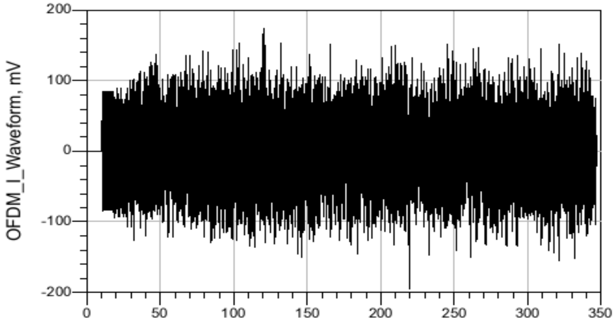
[Behzad, "Wireless LAN Radios"]

Digital Modulation Schemes

- Digital Modulation (bits/Hz)
  - Constant Envelope (FSK, PSK, GMSK)
  - Non-Constant Envelope (QAM, OFDM)

QPSK





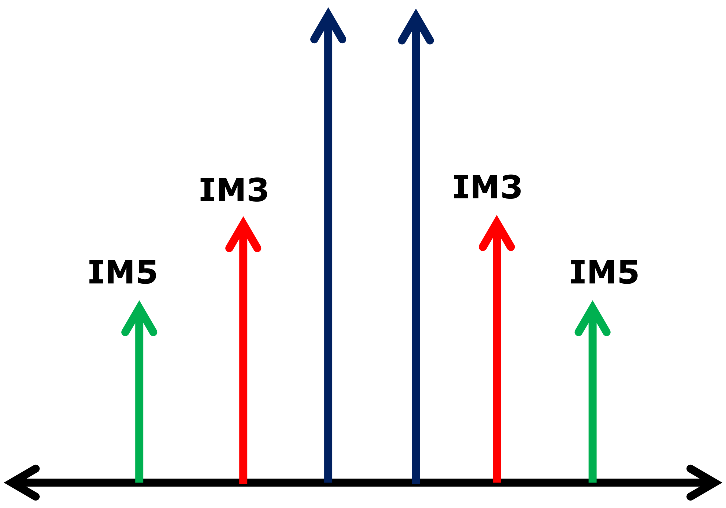
OFDM

- Maximize wireless users -> Spectral Efficiency
  - Use spectrally efficient modulation like 64-QAM
  - Band-limit PSK signals
    - Introduces amplitude variation
    - Need linear amplification to avoid spectral re-growth
  - FM/GMSK use non-linear PA but also use wider BW

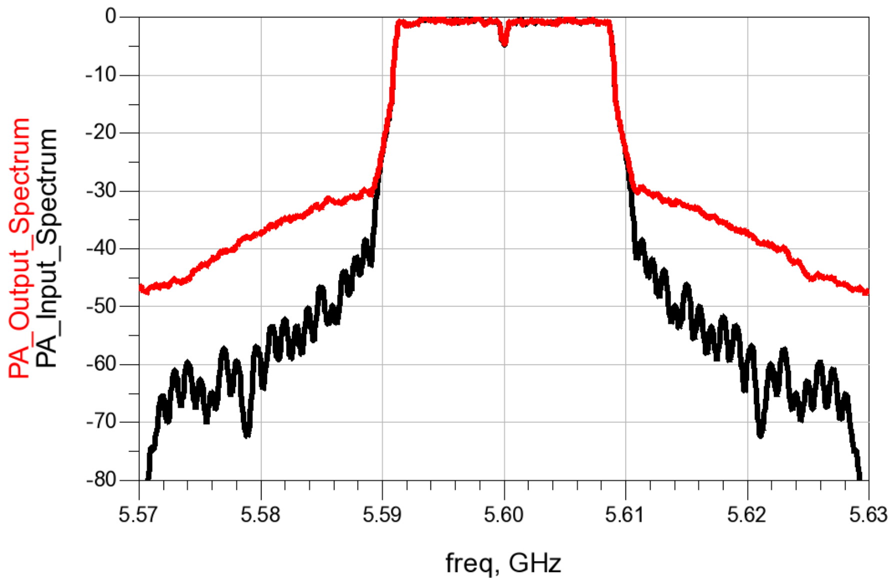
Spectral Mask

- Transmit spectral re-growth can
  - De-sensitize it’s receiver or nearby receivers
  - Create interference in adjacent channels
- Caused by inter-modulation distortion

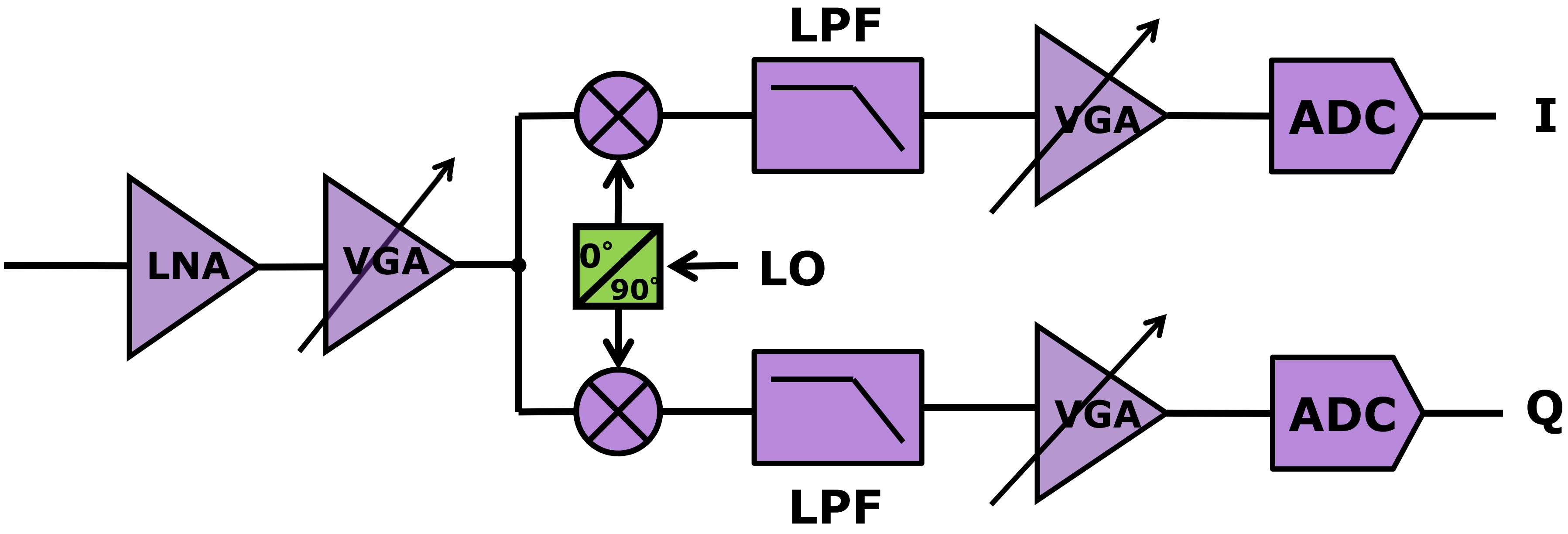
2-Tone Input



OFDM Spectrum

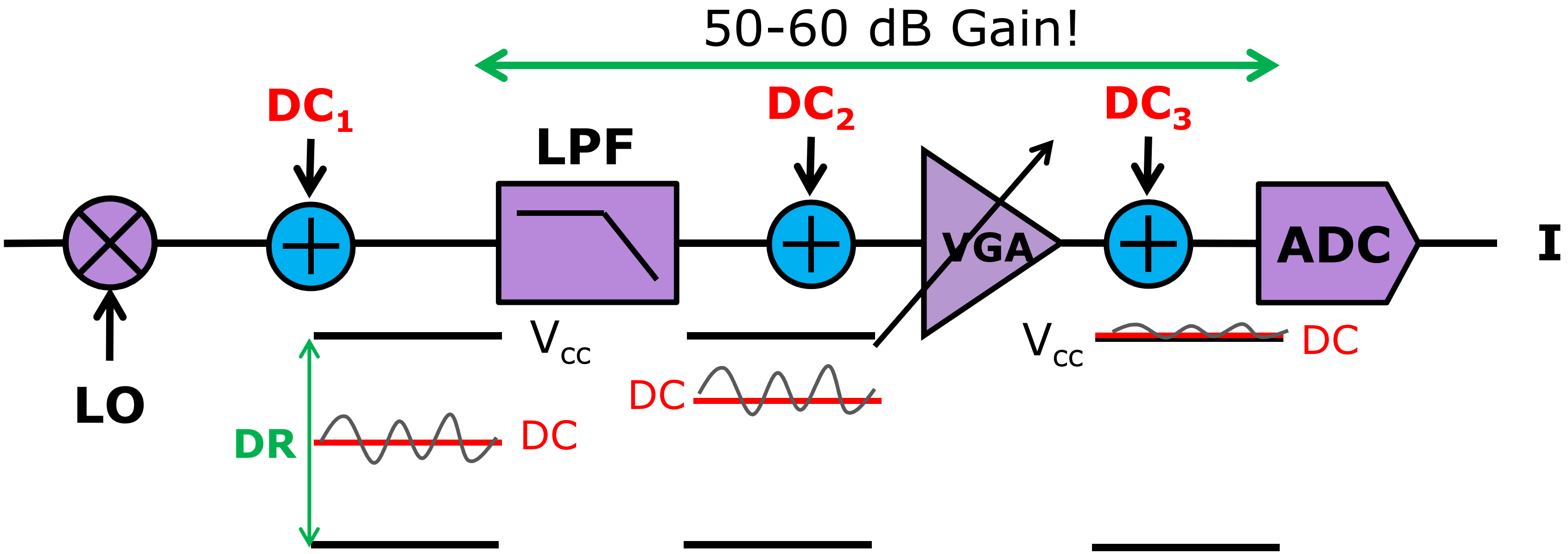


Direct Conversion Receiver



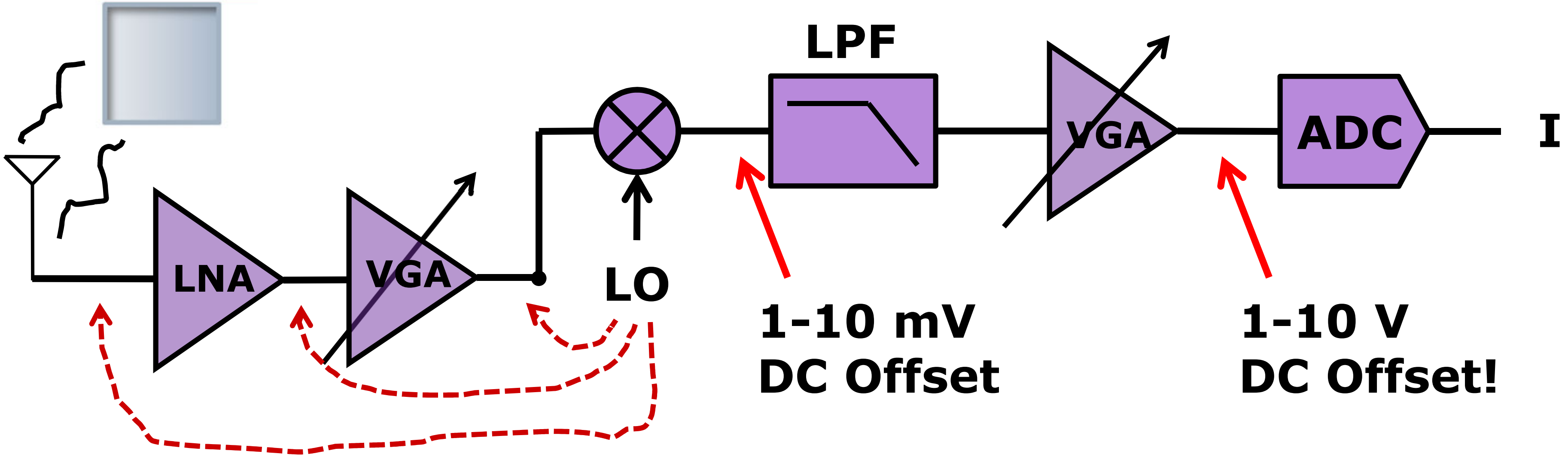
- Direct Conversion
  - High Frequency IQ Down-Conversion
  - On-chip filter subject to process variation and mismatch
  - DC Offset will be amplified by baseband gain
  - Blocker-induced IM2 may fall on top of desired channel

Static Receiver DC Offset



- ☐ DC Offset restricts dynamic range for signal
- ☐ DC Offset may be function of
  - VGA Gain Setting
  - Temperature
  - Supply voltage

Dynamic Receiver DC Offset

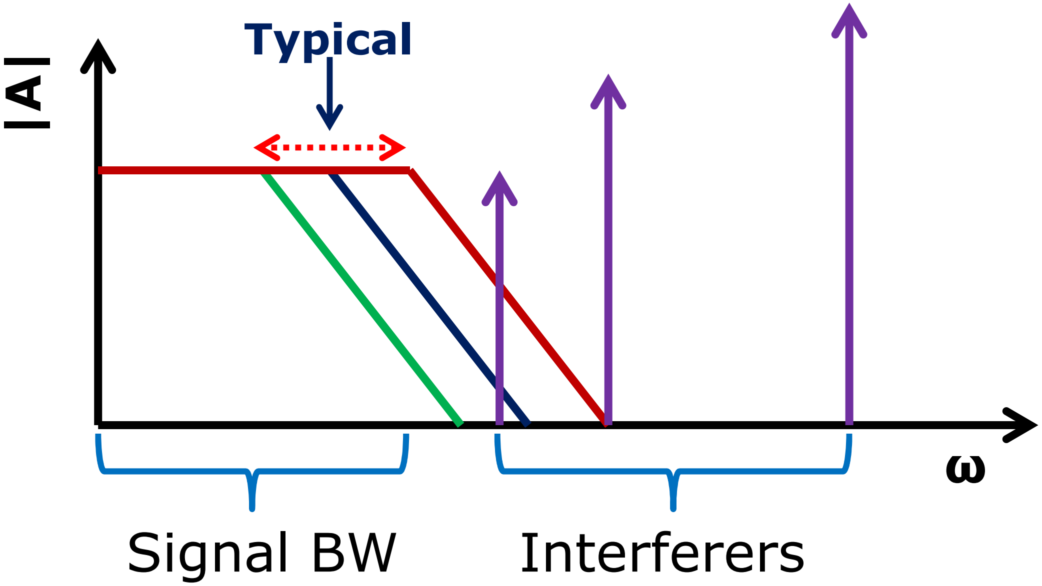


1-10 mV DC Offset

1-10 V DC Offset!

- LO Self-Mixing
  - Re-radiation, reflections
- Sensitive to Blockers (IM2)
- RF Gain-Dependent Mismatch

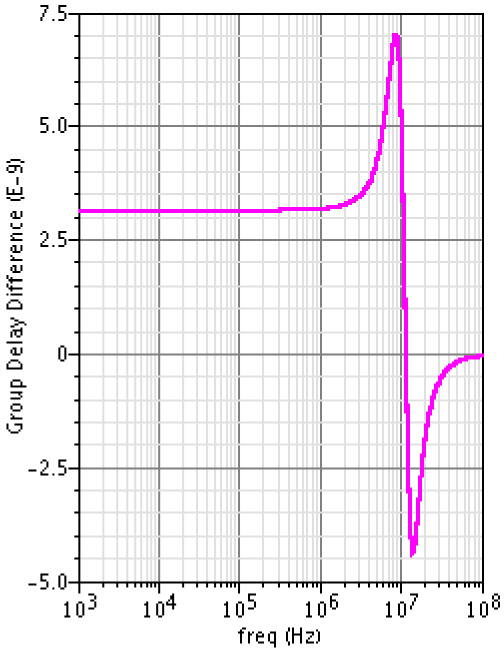
Analog Filter Variation



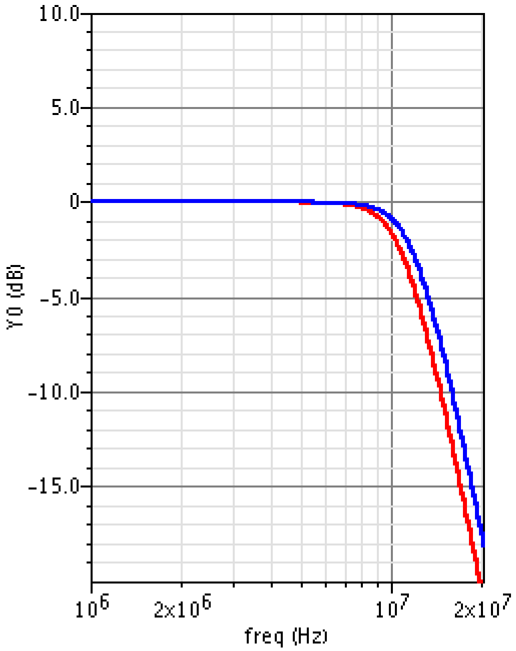
- Direct Conversion RX relies on accurate analog filtering
  - Channel Selectivity (reject adjacent/alternate channels)
  - Blocker Performance (interferers, co-existence)
  - Over-design of filter leads to noise/linearity/power penalty
- I and Q filter paths must match to level of -40 dB
  - Local random mismatch, gradient and proximity effects
  - Gain/phase matching may be frequency dependent

Frequency Dependent IQ Mismatch

Group Delay Mismatch

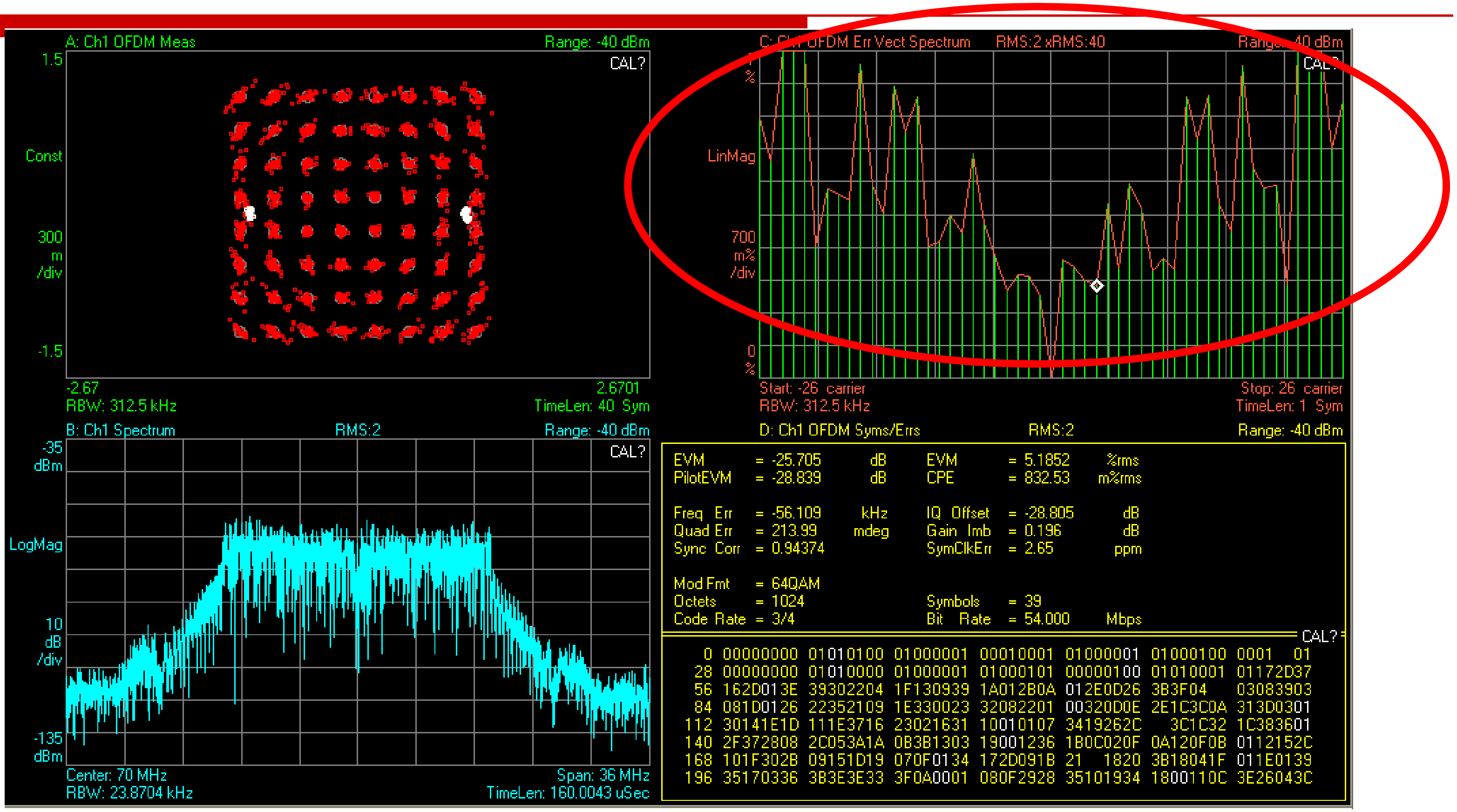


Filter Response



- Group Delay Mismatch causes frequency dependent phase mismatch (linear)
- Group Delay Mismatch itself may not be constant

Effect of Mismatch on EVM (802.11g)

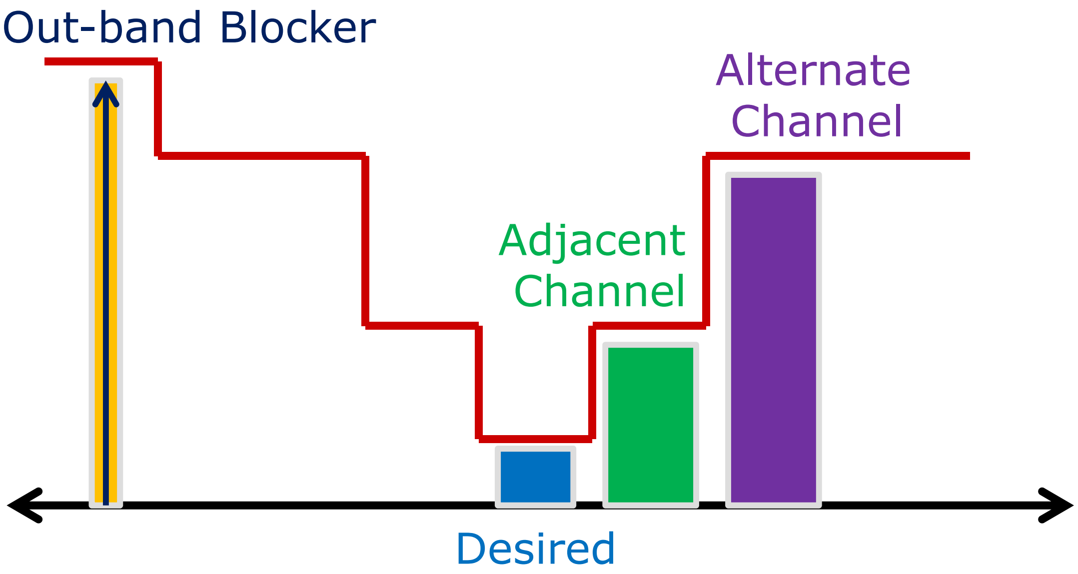


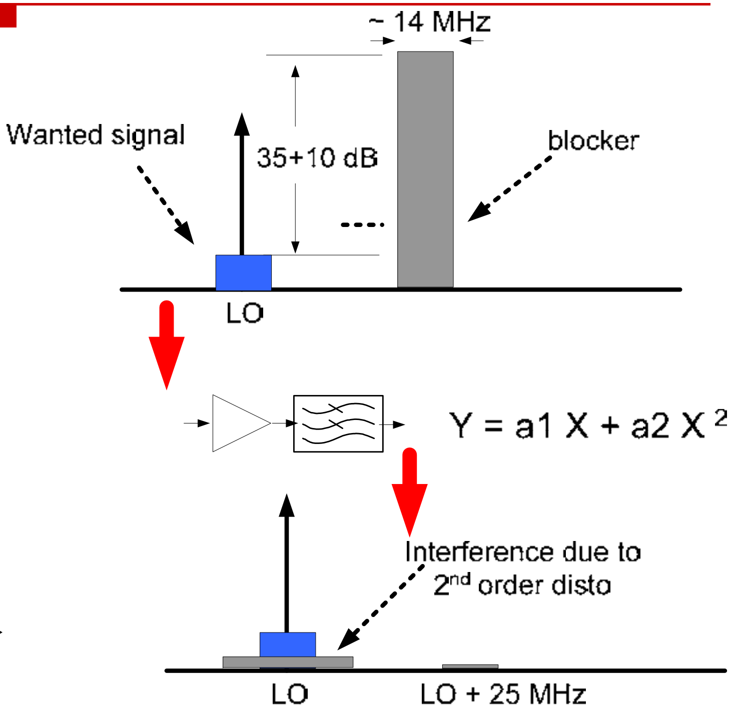
```
EVM       = -25.705    dB      EVM       = 5.1852    %rms
PilotEVM  = -28.839    dB      CPE       = 832.53   m%rms

Freq  Err = -56.109    kHz     IQ  Offset = -28.805    dB
Quad Err  = 213.99     mdeg    Gain  Imb  = 0.196      dB
Sync  Corr = 0.94374           SymClkErr  = 2.65       ppm

Mod Fmt   = 64QAM
Octets    = 1024               Symbols    = 39
Code Rate = 3/4                Bit   Rate = 54.000     Mbps

  0 00000000 01010100 01000001 00010001 01000001 01000100 0001     01
 28 00000000 01010000 01000001 01000101 00000100 01010001 01172D37
 56 162D013E 39302204 1F130939 1A012B0A 012E0D26 3B3F04   03083903
 84 081D0126 22352109 1E330023 32082201 00320D0E 2E1C3C0A 313D0301
112 30141E1D 111E3716 23021631 10010107 3419262C   3C1C32 1C383601
140 2F372808 2C053A1A 0B3B1303 19001236 1B0C020F 0A120F0B 0112152C
168 101F302B 09151D19 070F0134 172D091B 21   1820 3B18041F 011E0139
196 35170336 3B3E3E33 3F0A0001 080F2928 35101934 1800110C 3E26043C
```

- ☐ EVM deteriorates at higher frequency sub-carriers due to frequency dependent IQ mismatch

Intermodulation Distortion

RX Antenna Input

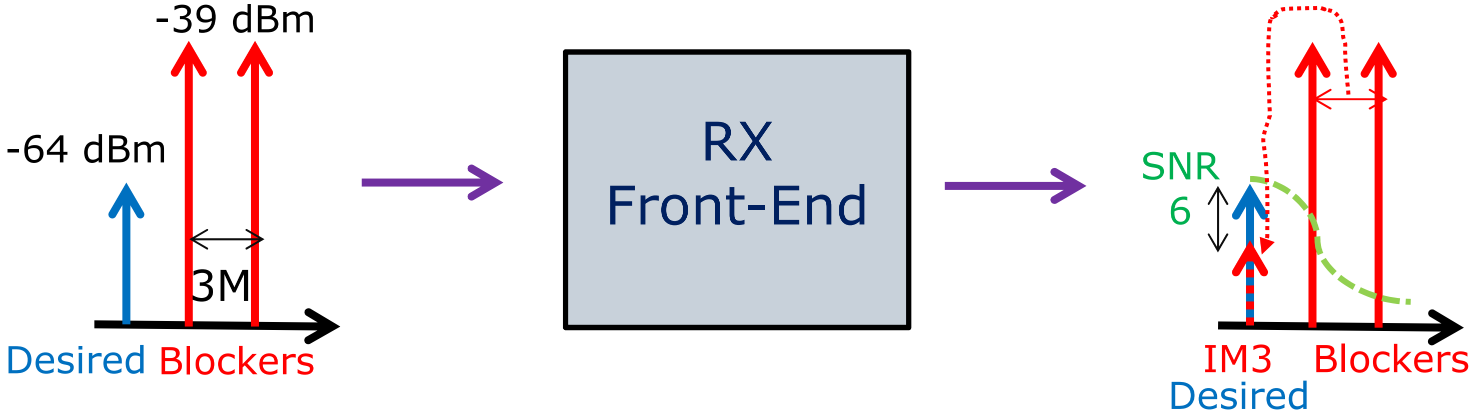




- RF input includes desired signal + large in-band and out-band blockers or interferers
- Interferers can degrade SNR of desired signal if folded over wanted signal through intermodulation distortion (IM3, IM2)

BT Intermodulation Specification

Packet Detection SNR: 6 dB
Min. desired signal: Ref. Sens. + 6 dB = -64 dBm
Level of blockers: -39 dBm
Freq. offset of blockers: 3M, 4M, or 5M



- BT In-band IIP3 Spec -

$$P_{IIP3} = P_{in} + \frac{P_{out} - P_{IM,out}}{2} = P_{in} + \frac{P_{in} - P_{IM,in}}{2} = \frac{3P_{in} - P_{IM,in}}{2}$$

$$P_{IIP3} = \frac{3P_{in} - P_{IM,in}}{2} = \frac{3*(-39)-(-64-6)}{2} = -23.5dBm$$

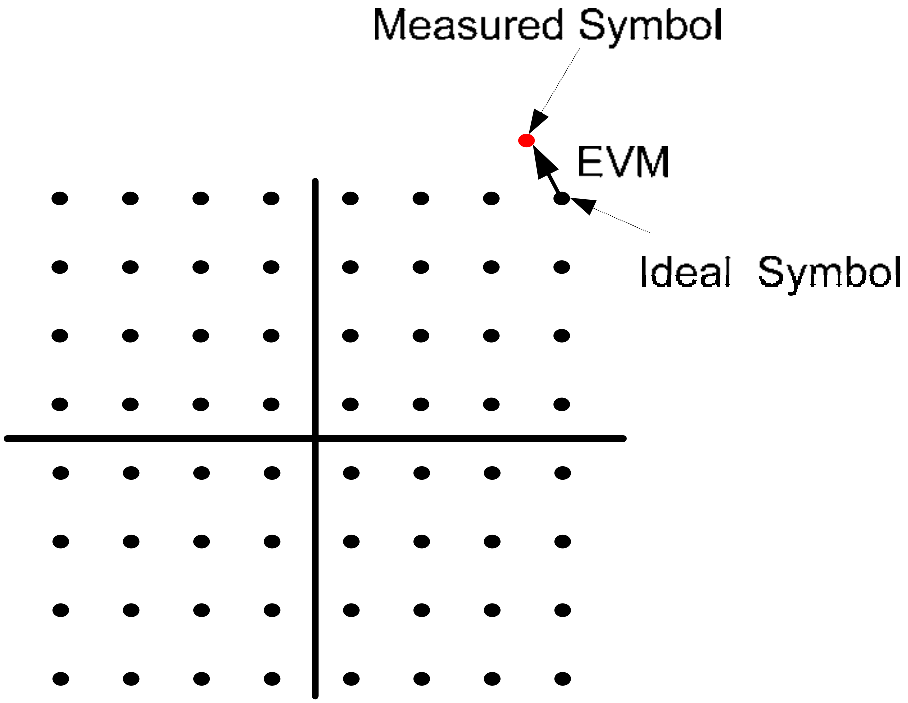
Current Challenges in Wireless

- 802.11 standards trend
  - Higher Throughput ->
    - Squeeze more bits/Hz (64 QAM -> 256 QAM)
    - Increase signal bandwidth (40 MHz -> 160 MHz)
  - Higher SNR requirement ->
    - Less tolerant to non-idealities

WLAN Standard	**Typical Pout**	**Signal BW**	**EVM**	**LO Leakage**	**Image Rejection**
802.11g	17 dBm	20MHz	-25 dB	-18 dBc	-35 dBc
802.11n	17 dBm	40MHz	-28 dB	-21 dBc	-40 dBc
802.11ac wave 1/2	17 dBm	80MHz/ 160MHz	-32 dB	-24 dBc/ -27 dBc	-45 dBc

TX/RX EVM

- ☐ Linearity
- ☐ Phase Noise
- ☐ Thermal Noise
- ☐ LO I/Q Mismatch
- ☐ BB I/Q Filter Mismatch



Calibration helps to improve modulation accuracy

Digital Calibration Enables …

- Use advanced architectures -> Save area/cost
  - Direct Conversion
  - Polar Transmitter
- Relax analog specs -> Save power/area
  - Overcome impairments that degrade performance
  - More aggressive design approach
- Improve yield and reduce factory calibration
  - RF BIST
- Availability of cheap digital processing power
  - Less circuit overhead than analog calibration

Outline

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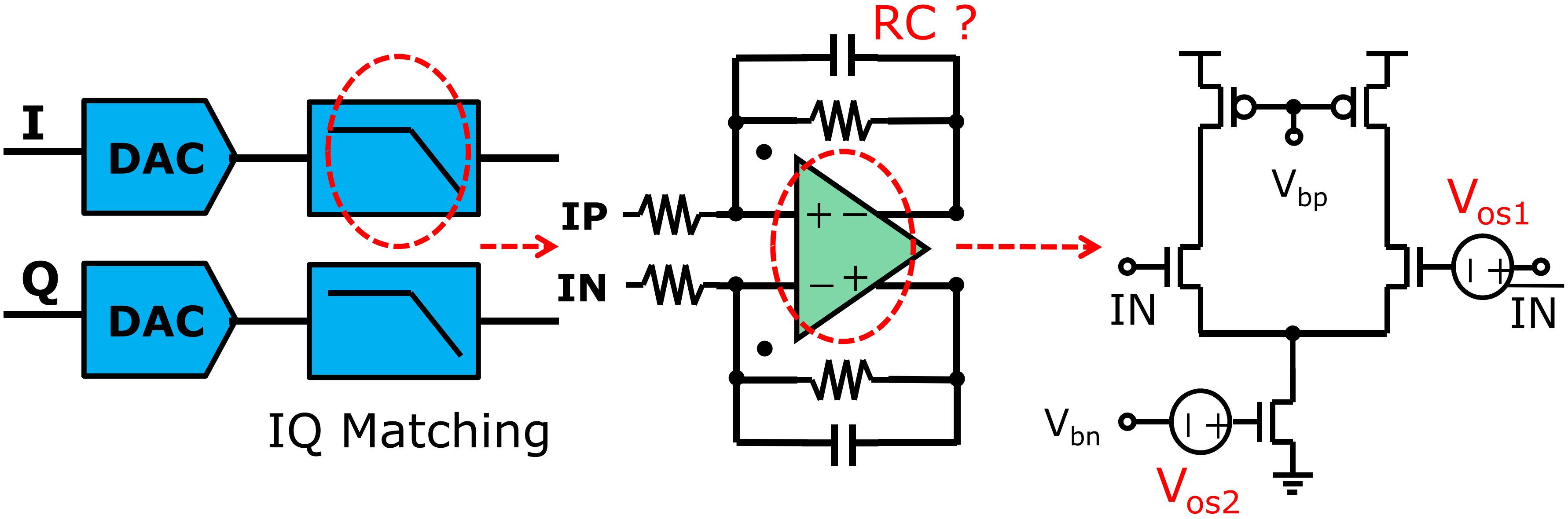
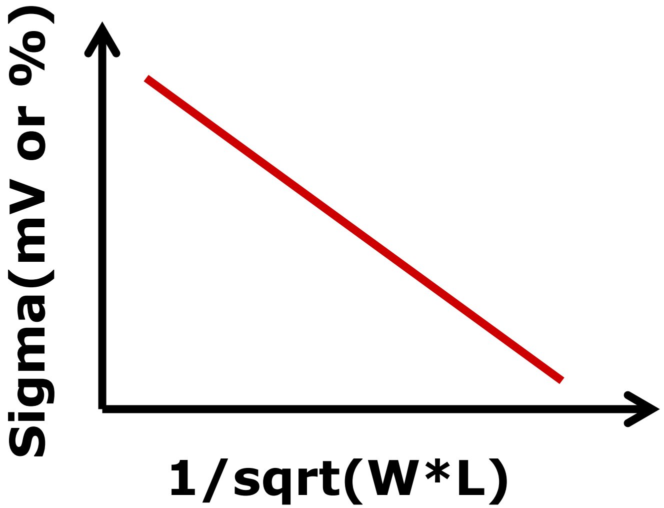
  - Direct Conversion Transceiver
  - Transceiver Building Blocks and their Impairments
  - Motivation for Digital Calibration
- Digital Calibration Techniques to Address
  - Device Mismatch
    - DC Offset Cancellation
    - IQ and LOFT Calibration
  - Process Variation
    - Analog Filter BW Calibration
    - Gain Calibration
    - VCO LC Tank Calibration
  - Non-Linearity
    - IMD Cancellation, PA Pre-Distortion

Device Mismatch



- Analog IC Design relies heavily on
  - Differential Matching
  - IQ Matching
  - Transistor and Passive Device Matching

Device Mismatch Equations



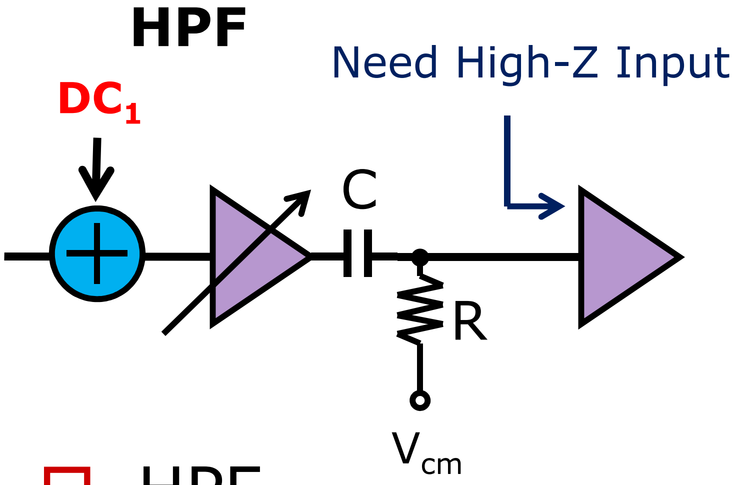
- Improve Matching by
  - Larger Area
  - Higher overdrive
    - Higher noise
    - Voltage Headroom
- Against scaling trend!

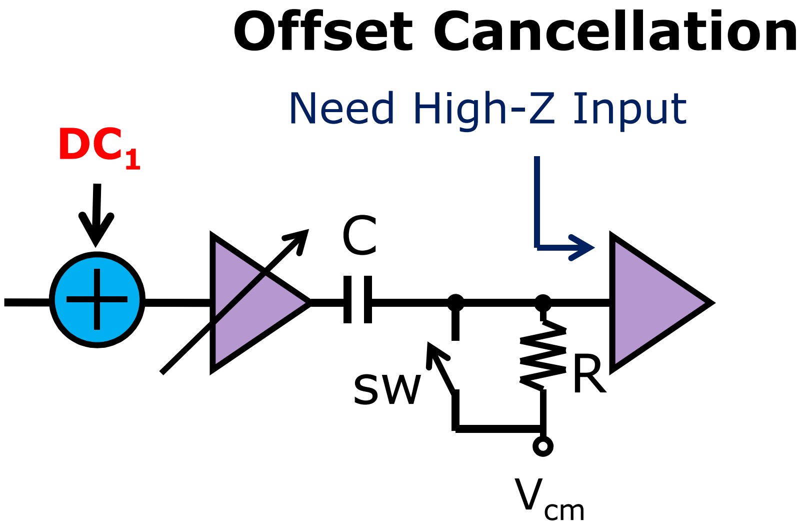
$$\left(\frac{\sigma(I_d)}{I_d}\right)^2 = \sigma^2\left(\frac{\Delta\beta}{\beta}\right) + \frac{4\sigma^2(\Delta V_t)}{(V_{gs}-V_t)^2} = \frac{1}{2WL}\left(A_\beta^2 + \frac{4A_{V_t}^2}{(V_{gs}-V_t)^2}\right)$$

$$\sigma\left(\frac{\Delta\beta}{\beta}\right) = \frac{A_\beta}{\sqrt{WL}} + B_\beta \qquad \sigma(\Delta V_t) = \frac{A_{V_t}}{\sqrt{WL}}$$

$$\left.\begin{array}{l} A_\beta \propto \% \cdot \mu m \\ A_{V_T} \propto mV \cdot \mu m \end{array}\right\} \text{Process Constants}$$

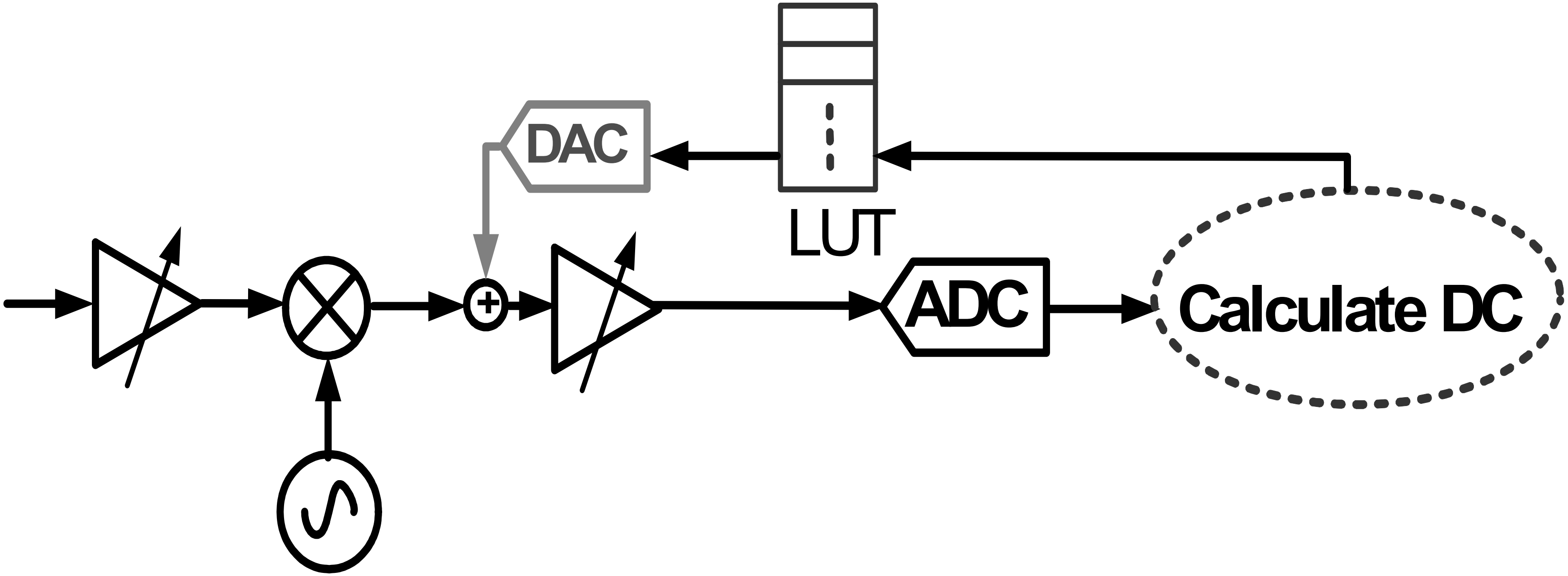
DC Offset Cancellation - Analog





- HPF
  - Large capacitor area
  - Settling time vs. low frequency corner (1/RC)
  - Low-IF allows higher corner and faster settling
- Offset Cancellation during idle period
  - "Variable" settling time, but still consumes area
  - Quickly settle DC offset due to gain change

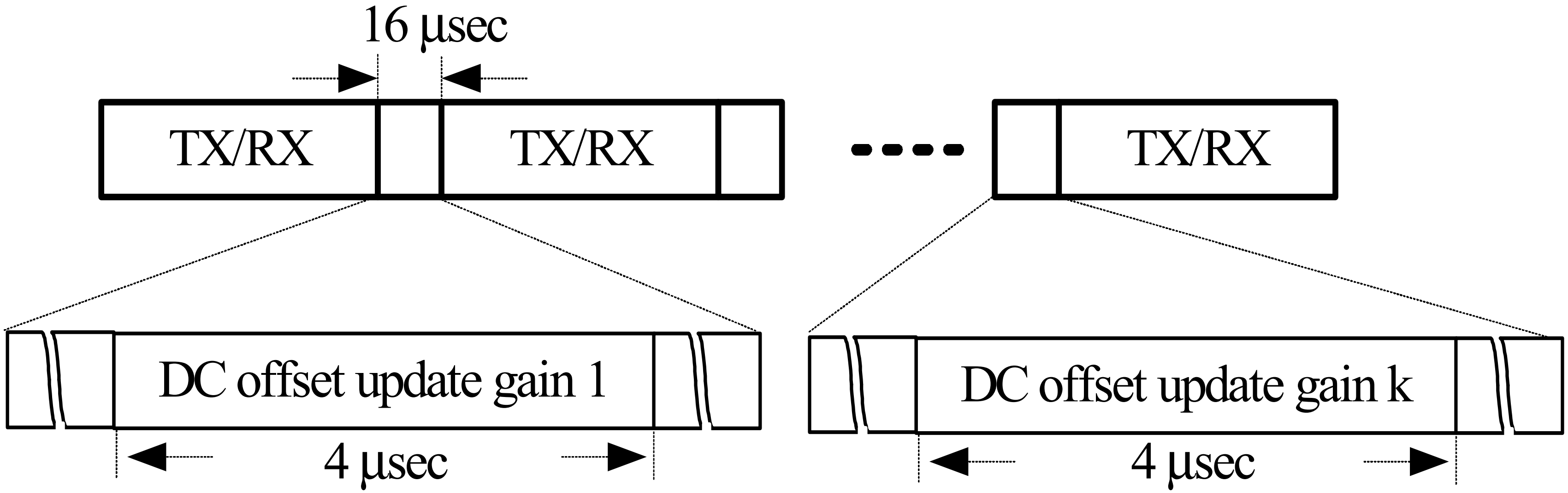
DC Offset Cancellation - Digital



- Phase 1
  - Calculate DC correction for different gains at power up
  - Use DACs at mixer output

[Vassiliou, JSSC 03]

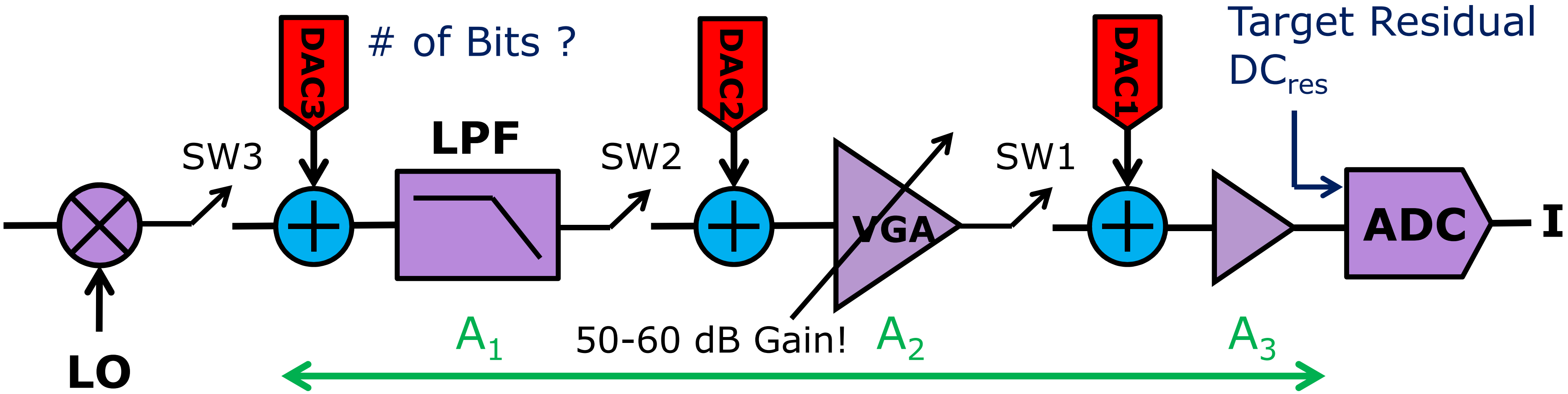
DC Offset Cancellation - Digital



- Phase 2
  - Track DC Offset
  - Alternatives
    - Track temperature and maintain multiple LUTs
    - Insert calibration cycle on RX based on temp sensor

[Vassiliou, JSSC 03]

DAC Cancellation Design



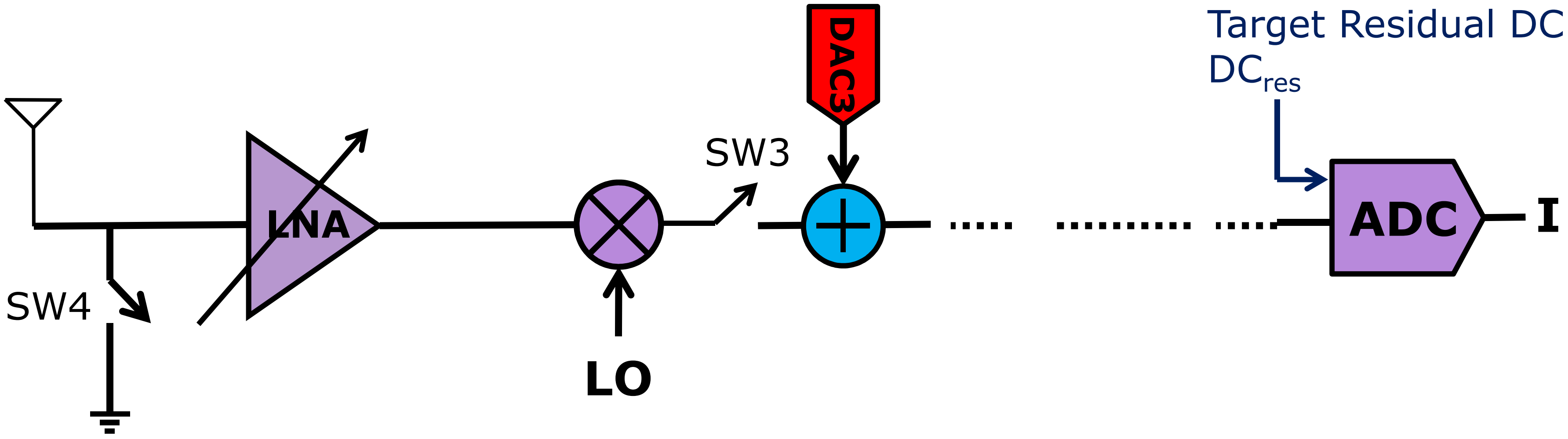
- ☐ Configure switches and calibrate each DAC independently to cancel it’s input-referred offset
- ☐ DAC Step Size determined by gain and DC spec
- ☐ DAC noise voltage < DAC LSB

$$DAC3_{LSB} = \frac{DC_{res}}{A_1 A_2 A_3}$$

Use Averaging Factor N :

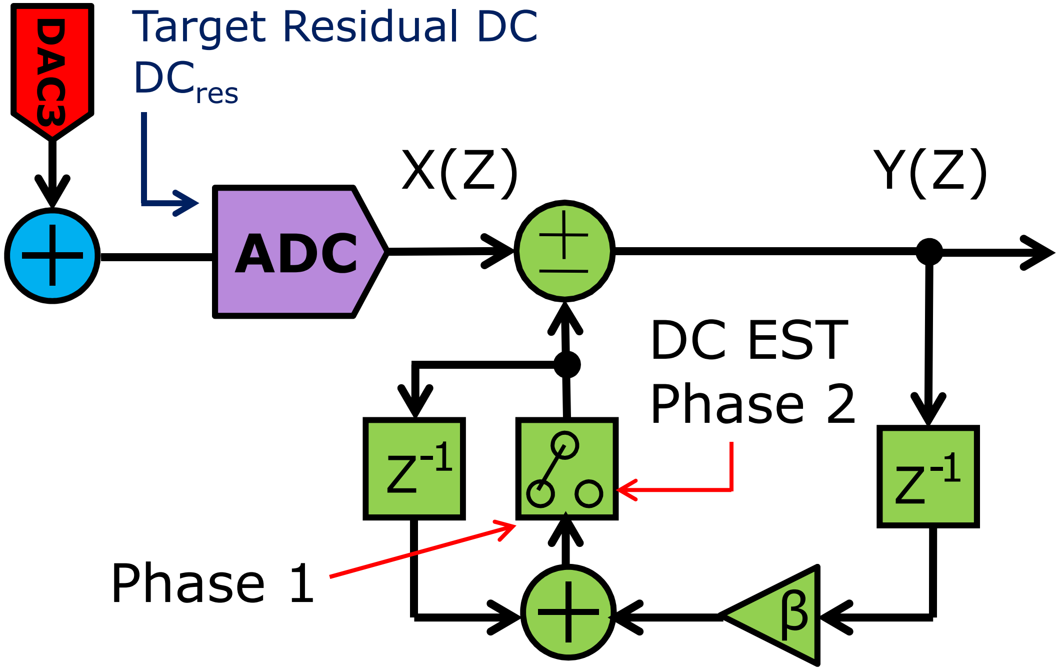
$$N > \left(\frac{\sigma_{n3}}{DAC3_{LSB}} \right)^2$$

Dynamic DC Offset



- ☐ Antenna signals may affect DC offset
  - Short LNA input to ground during calibration
- ☐ LO self-mixing DC offset may vary with LNA gain and antenna impedance
- ☐ Input referred DC offset < system noise floor
  - ADC Range and Digital DC Removal can relax this

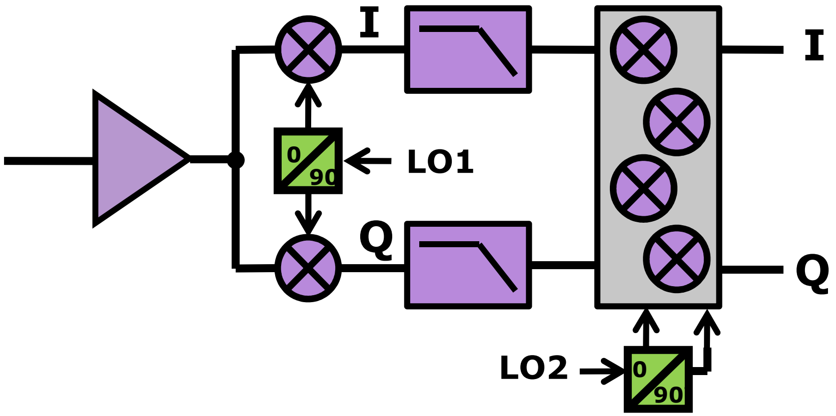
Digital DC Offset Correction



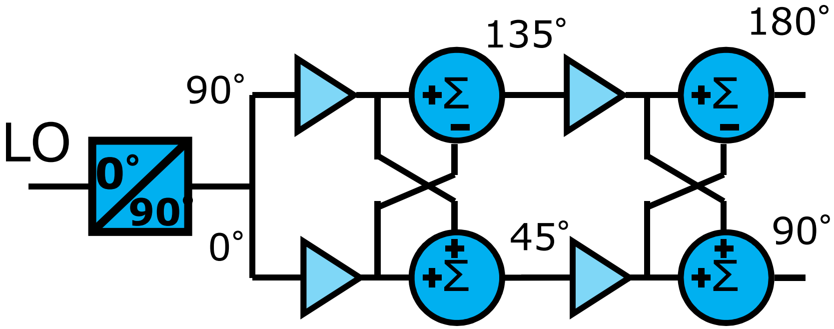
$$\frac{Y(Z)}{X(Z)} = \frac{1 - z^{-1}}{1 - (1 - \beta) z^{-1}}$$

- ☐ Phase 1 : High-pass Filter Mode
- ☐ Phase 2 : Open Loop Mode or Narrow-BW Mode
- ☐ If analog gain is low, can do DCOC all in digital

IQ Mismatch (Analog Techniques)



Double Quadrature Down-Converter



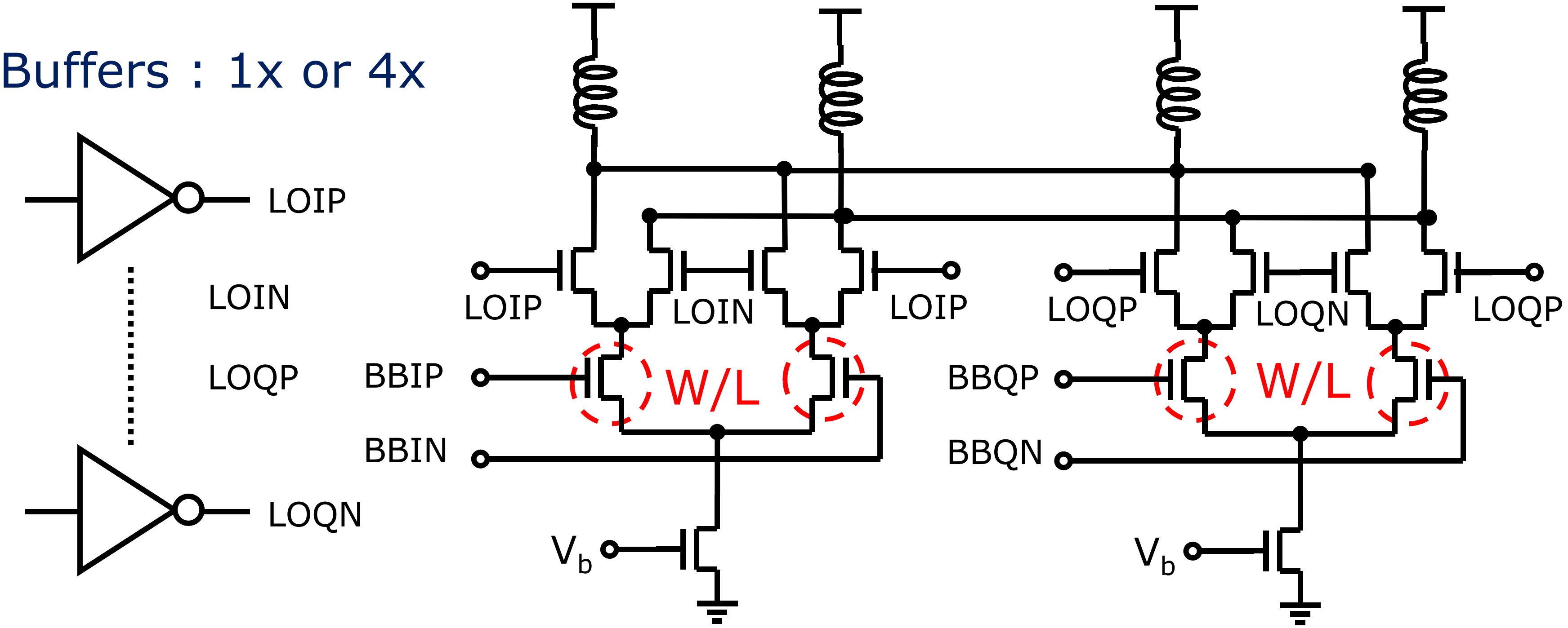
Havens Phase Corrector

- Before digital calibration
  - Brute Force Design (larger area)
  - Multi-Stage Poly-phase Filter
  - Havens Phase Corrector
  - Double Quadrature Mixer Architecture … and others
- Still subject to process variation

[Havens Patent]

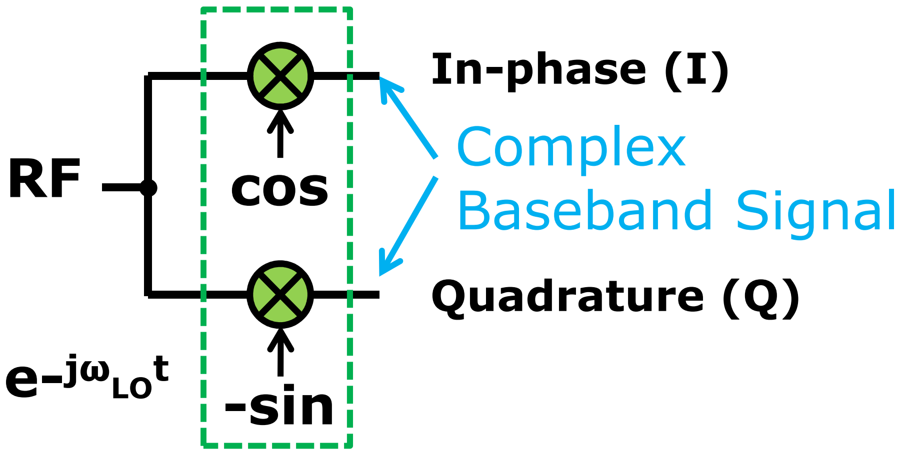
IQ LO Buffer Circuit Example

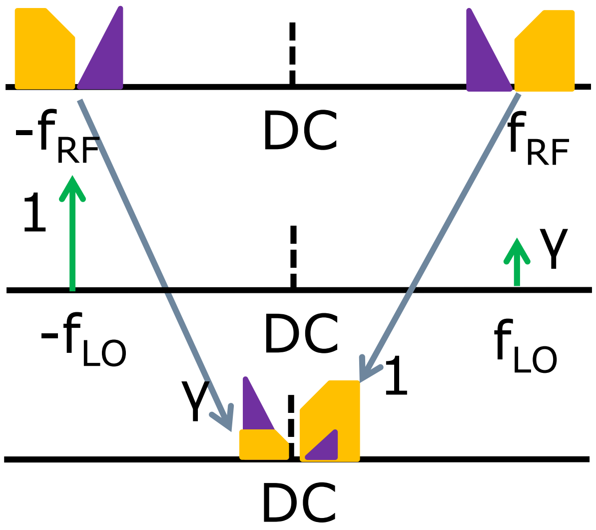
LO Buffers : 1x or 4x



LO Buffer Size	I_{rms} @ 5GHz	3σ Phase Mismatch
1X	2 mA	4.2°
4X	5.4 mA	2.5°

IQ Complex Mixer

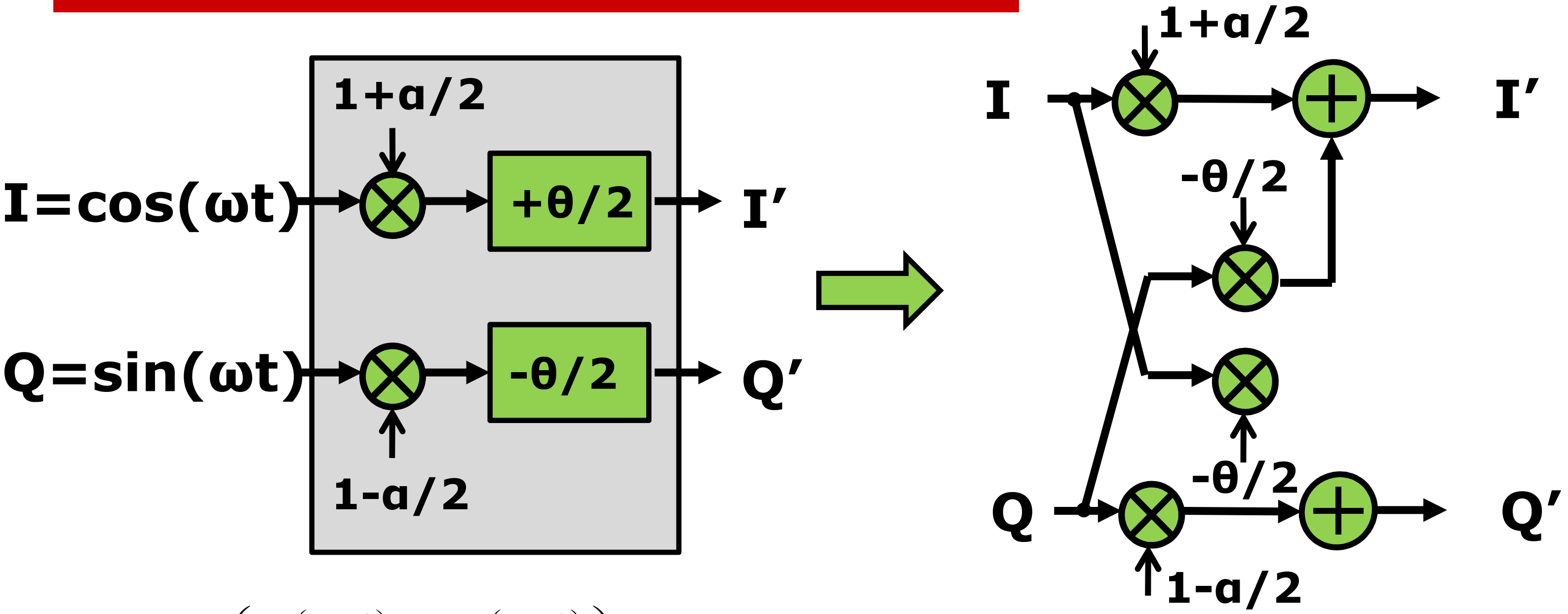




- Complex negative frequency local carrier is equivalent to multiplying by two real I/Q carriers with 90° phase delta
- The positive frequency is translated to DC and found by I+jQ if the I/Q carriers are ideal
- Imbalances allow the negative frequency to move to DC causing leakage of the image found by I-jQ

[S. Lerstaveesin, JSSC 06]

IQ Complex Non-Ideal Mixer Model



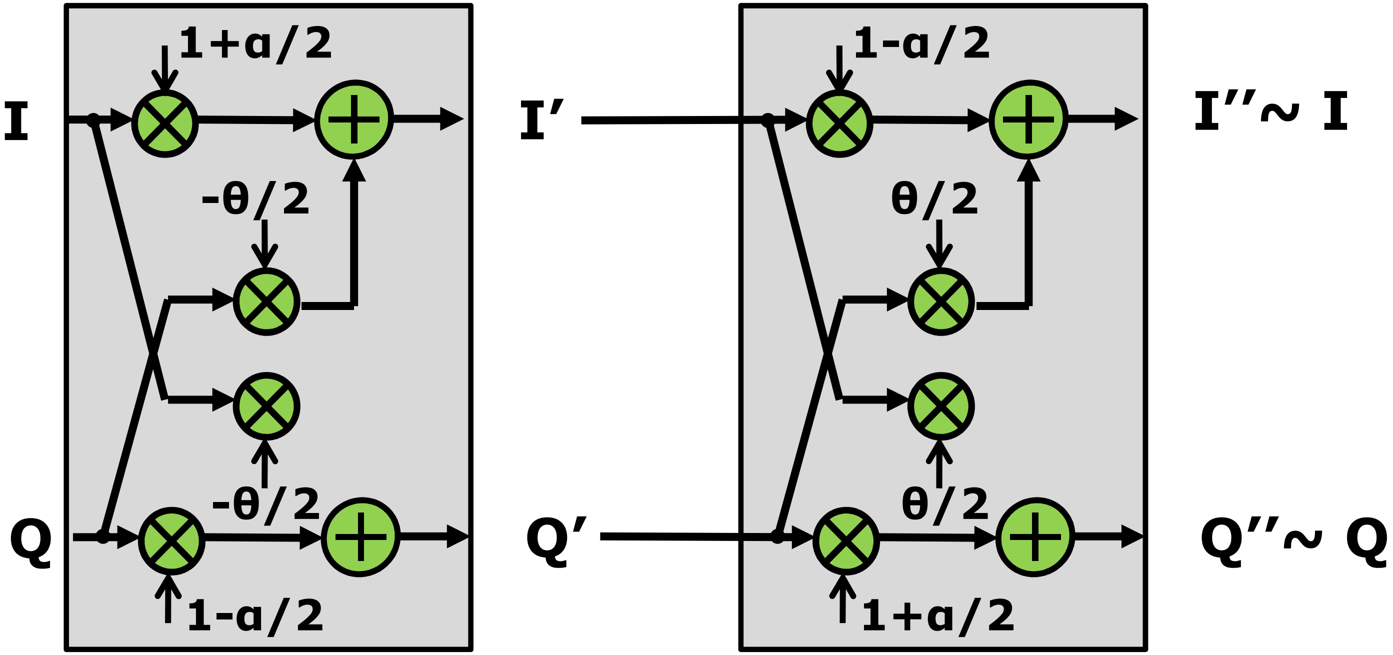
$$I' = \left(1+\frac{\alpha}{2}\right)\left(\frac{e^{j\left(\omega t+\frac{\theta}{2}\right)}+e^{-j\left(\omega t+\frac{\theta}{2}\right)}}{2}\right)$$

$$Q' = \left(1-\frac{\alpha}{2}\right)\left(\frac{e^{j\left(\omega t-\frac{\theta}{2}\right)}-e^{-j\left(\omega t-\frac{\theta}{2}\right)}}{2j}\right)$$

$$\begin{bmatrix} I' \\ Q' \end{bmatrix} = \begin{bmatrix} 1+\frac{\alpha}{2} & \frac{-\theta}{2} \\ \frac{-\theta}{2} & 1-\frac{\alpha}{2} \end{bmatrix} \begin{bmatrix} I \\ Q \end{bmatrix}$$

[S. Lerstaveesin, JSSC 06]

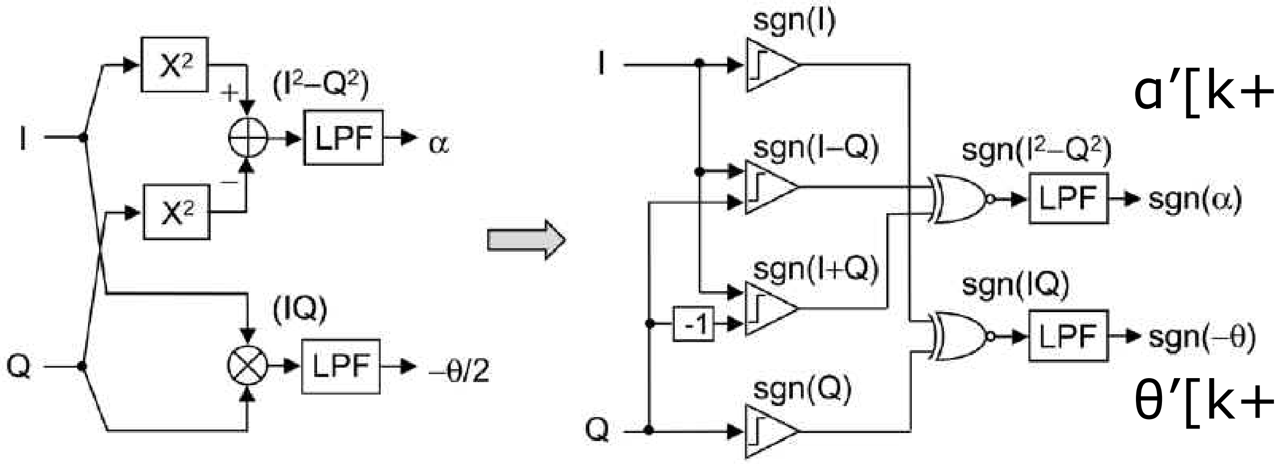
IQ Image Rejection Method



$$\begin{bmatrix} I'' \\ Q'' \end{bmatrix} = \begin{bmatrix} 1-\frac{\alpha}{2} & +\frac{\theta}{2} \\ +\frac{\theta}{2} & 1+\frac{\alpha}{2} \end{bmatrix} \begin{bmatrix} I' \\ Q' \end{bmatrix} = \begin{bmatrix} 1-\frac{\alpha^2}{4}-\frac{\theta^2}{4} & 0 \\ 0 & 1-\frac{\alpha^2}{4}-\frac{\theta^2}{4} \end{bmatrix} \begin{bmatrix} I \\ Q \end{bmatrix}$$

[S. Lerstaveesin, JSSC 06]

Gain and Phase Error Detection



$\alpha'[k+1]=\alpha'[k]+\mu_\alpha \mathrm{sgn}(\alpha)$

$\theta'[k+1]=\theta'[k]-\mu_\theta \mathrm{sgn}(-\theta)$

- Residual errors in I and Q can be reduced adaptively with a zero-forcing feedback loop using LMS update eqn.

$$LPF(I^2-Q^2)=\frac{\left(1+\frac{\alpha}{2}\right)^2}{2}-\frac{\left(1-\frac{\alpha}{2}\right)^2}{2}\approx\alpha$$

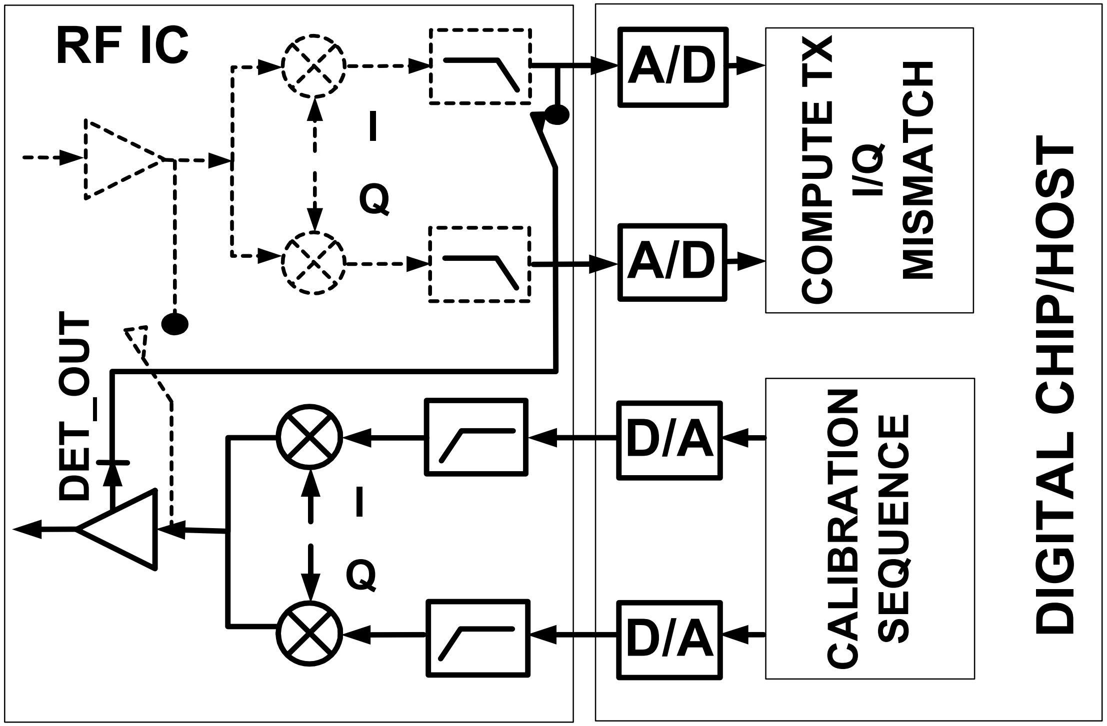
$$LPF(I*Q)=-\theta\frac{\left(1-\alpha^2/4\right)}{2}\approx-\frac{\theta}{2}$$

Note that $(I+jQ)(I-jQ)^*=I^2-Q^2+2jIQ=\alpha-j\theta$

We are minimizing correlation between signal and image.

[S. Lerstaveesin, JSSC 06]

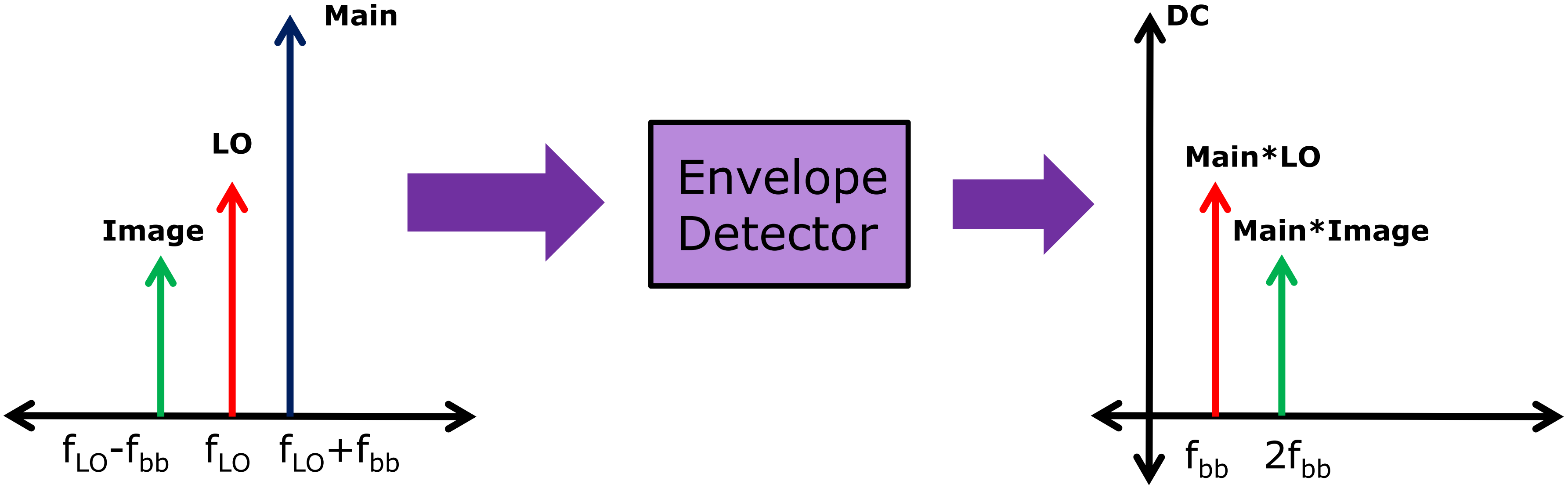
TX IQ/LOFT Calibration



1. Use on-chip envelope detector
2. Loopback to RX A/D
3. Compute TX I/Q mismatch and LO leakage
4. Pre-distort digital data real-time in baseband

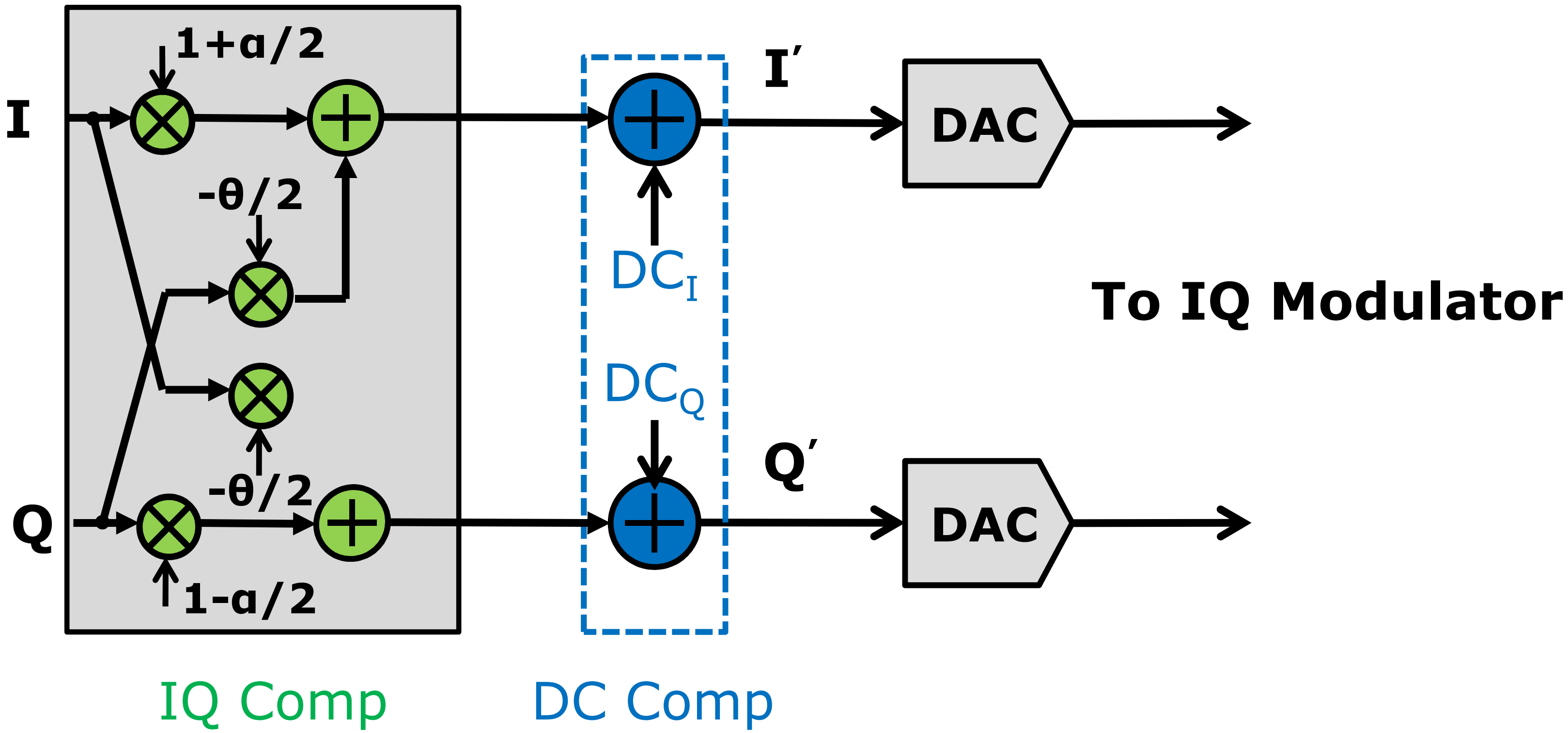
[Vassiliou, JSSC 03]

Envelope Detector



- Envelope detector extracts ripple from RF signal
- The envelope signal has 2 “beat-frequency” components
    1. $(f_{LO}+f_{bb})-f_{LO} = f_{bb}$
    2. $(f_{LO}+f_{bb})-(f_{LO}-f_{bb}) = 2f_{bb}$
- Amplitudes are proportional to LO/image tones
- Ignore DC and high frequency products

Digital IQ/LOFT Compensation

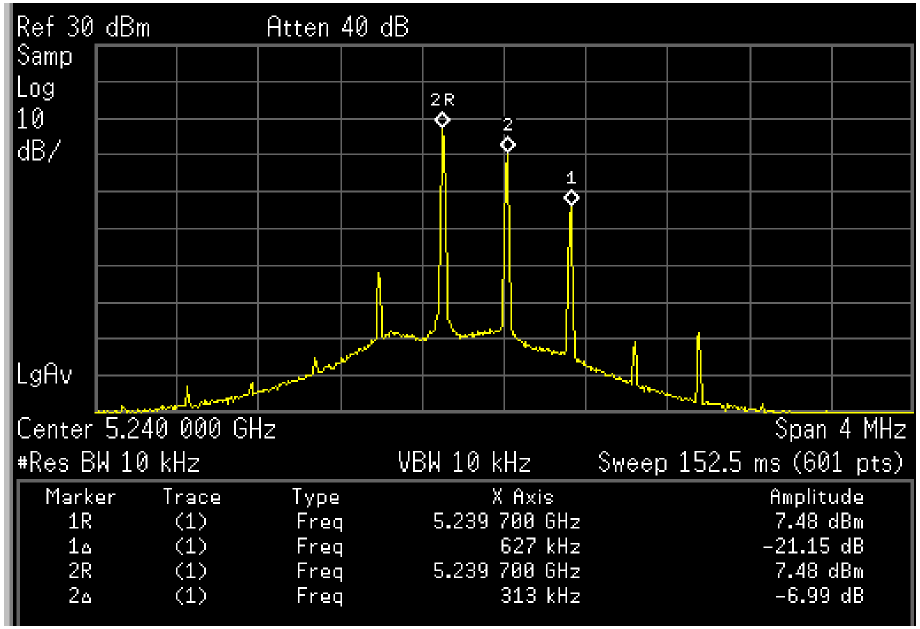


- Single-tone training or multiple amplitude tones
- LMS-based algorithm to find coefficients α, θ, DC_I, DC_Q

[Cavers, IEEE Trans. 97]

TX IQ CAL – Example Results

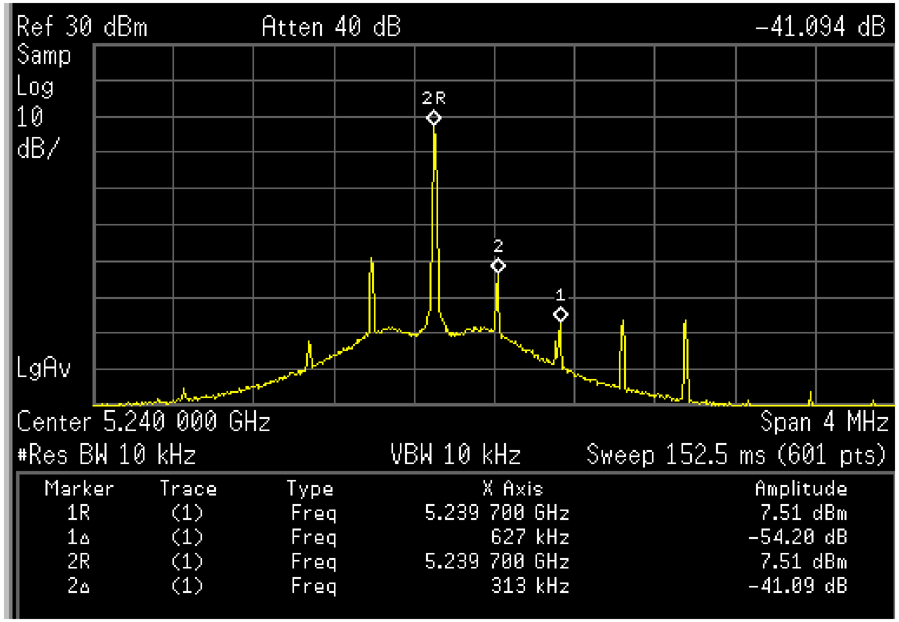
Pre-calibration



```
Ref 30 dBm          Atten 40 dB
Samp
Log
10
dB/

LgAv
Center 5.240 000 GHz                                  Span 4 MHz
#Res BW 10 kHz            VBW 10 kHz      Sweep 152.5 ms (601 pts)
 Marker   Trace    Type         X Axis                 Amplitude
   1R      (1)     Freq    5.239 700 GHz               7.48 dBm
   1Δ      (1)     Freq          627 kHz             -21.15 dB
   2R      (1)     Freq    5.239 700 GHz               7.48 dBm
   2Δ      (1)     Freq          313 kHz              -6.99 dB
```

- Image Rejection -21 dBc
- LO Leakage -7 dBc

Post-calibration

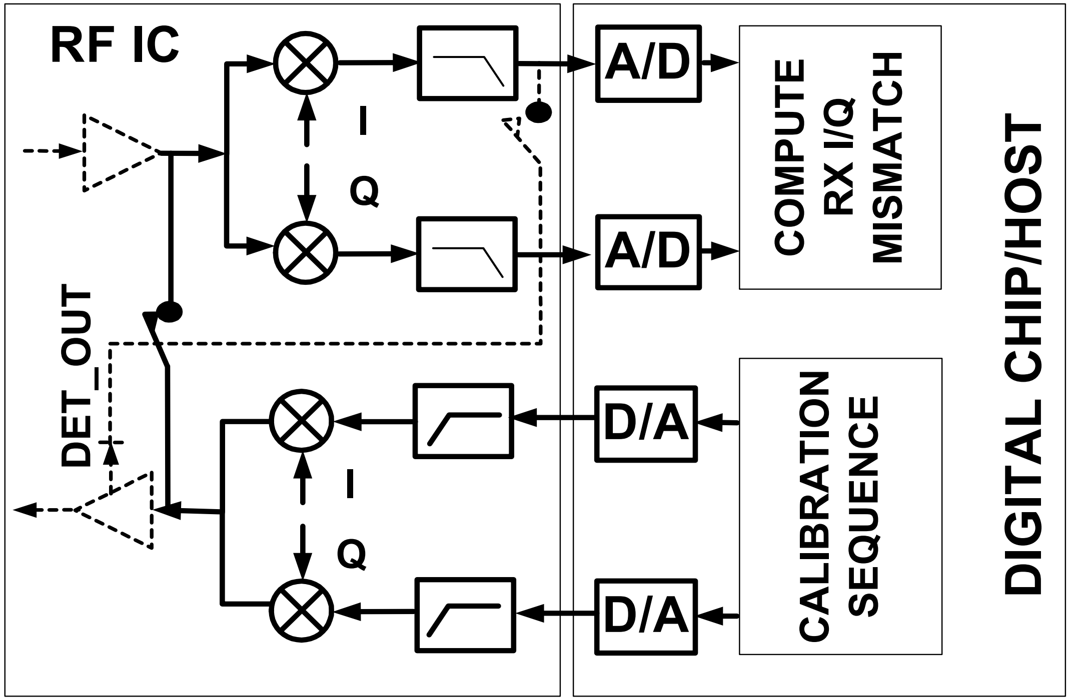


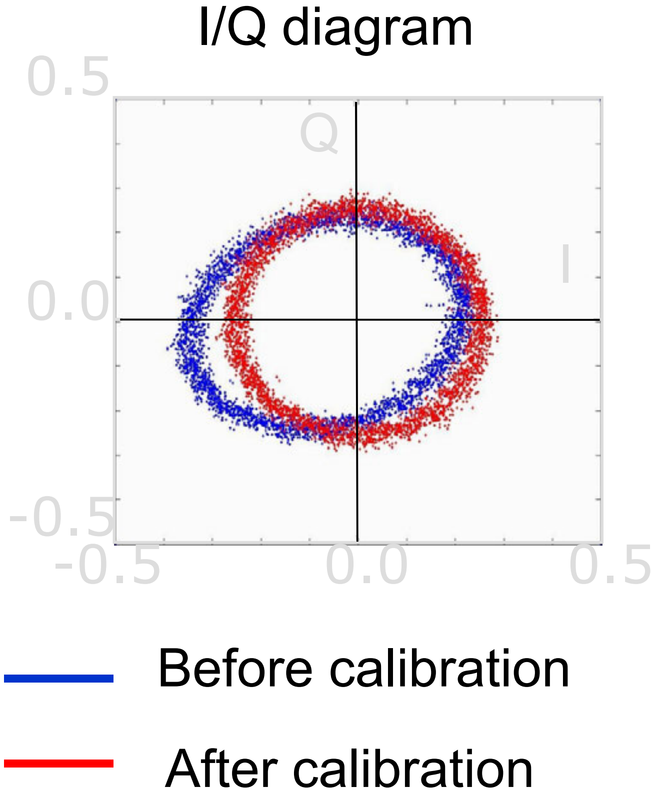
```
Ref 30 dBm          Atten 40 dB                       -41.094 dB
Samp
Log
10
dB/

LgAv
Center 5.240 000 GHz                                  Span 4 MHz
#Res BW 10 kHz            VBW 10 kHz      Sweep 152.5 ms (601 pts)
 Marker   Trace    Type         X Axis                 Amplitude
   1R      (1)     Freq    5.239 700 GHz               7.51 dBm
   1Δ      (1)     Freq          627 kHz             -54.20 dB
   2R      (1)     Freq    5.239 700 GHz               7.51 dBm
   2Δ      (1)     Freq          313 kHz             -41.09 dB
```

- Image Rejection -54 dBc
- LO Leakage -41 dBc
- Stable with temperature

[Vassiliou, JSSC 03]

RX IQ Mismatch Calibration

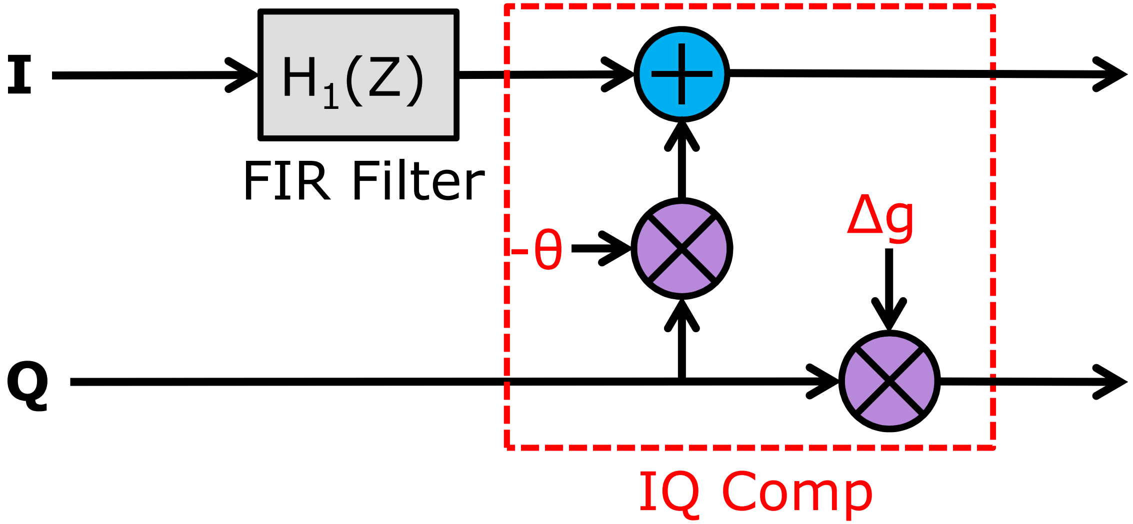




- Use corrected TX path
- Loopback to RX
- Compute RX I/Q mismatch (same principle as TX)
- Post-distort (digital or analog)

[Vassiliou, JSSC 03]

Frequency Dependent IQ CAL



- ☐ Apply different tone frequencies f_1 and f_2 for "frequency-dependent mismatch"
- ☐ Calculate IQ group delay mismatch Δt from difference of calibrated phase mismatches θ_1, θ_2
- ☐ Compensate delay with FIR filter

IQ Calibration Design Flow

1) Translate IRR spec to gain/phase mismatch

2) Determine required gain/phase tuning resolution

   0.5° phase step and 0.1 dB gain step -> 40 dB IRR

3) Determine max gain/phase mismatch range based on analog circuit block simulations

   +/- 5° phase mismatch and +/- 3 dB gain mismatch →

   20 phase steps and 60 gain steps required →

   Search matrix is ~ 1200

4) Design search algorithm (binary search, adaptive)

5) Design loopback path and envelope detector and provide gain tuning to guarantee linearity/noise

IQ Cal Practical Implementation

- ADC resolution and SNR are not critical
  - Averaging – trade time for SNR
- IQ mismatch changes during loopback
  - Different loading for LO circuitry
  - Supply voltage change
  - Coupling effects from loopback path
- Calibration time = $(t_s + (1/f_{bb})*N)*S*C$
  - Settling time of loopback path, t_s
  - Tone frequency, f_{bb}
  - # times of averaging, N
  - # searches required to reach target spec, S
  - # calibrations to perform, C (Tx/Rx, LOFT, gain/phase)

Outline

- Modern Transceiver Architecture
  - Direct Conversion Transceiver
  - Transceiver Building Blocks and their Impairments
  - Motivation for Digital Calibration

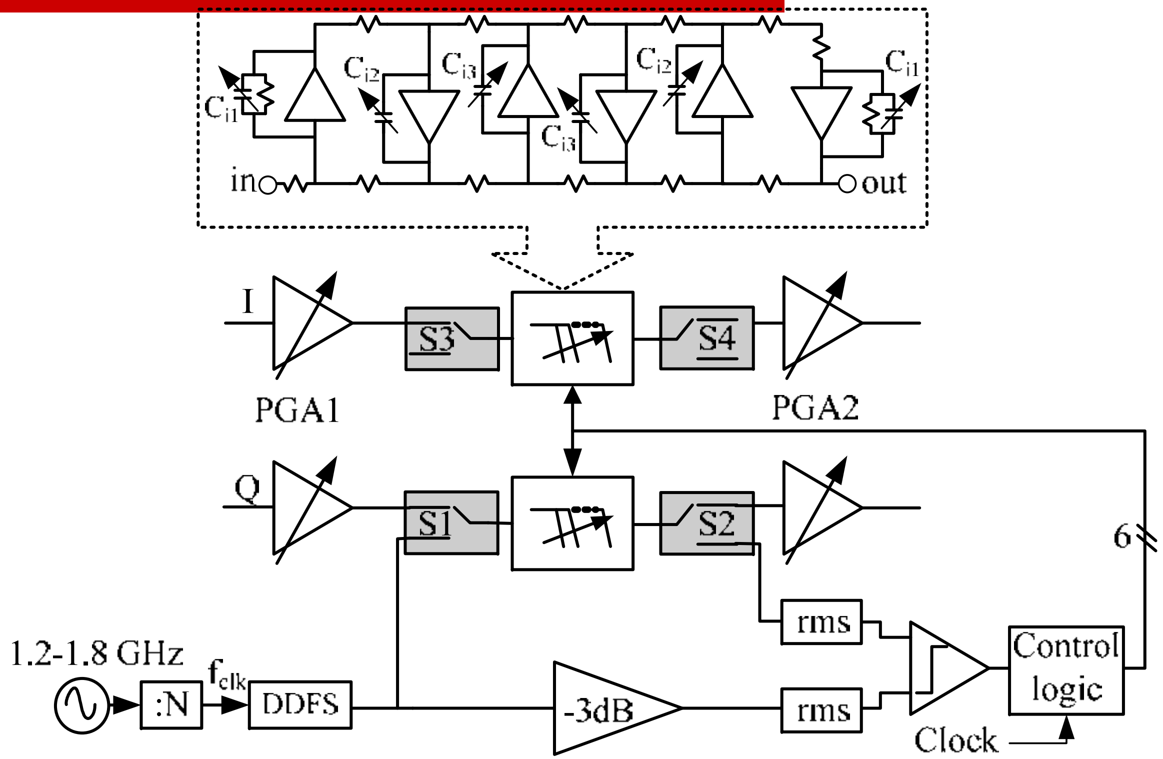
- Digital Calibration Techniques to Address
  - Device Mismatch
    - DC Offset Cancellation
    - IQ and LOFT Calibration
  - Process Variation
    - Analog Filter BW Calibration
    - Gain Calibration
    - VCO LC Tank Calibration
  - Non-Linearity
    - IMD Cancellation, PA Pre-Distortion

Process Variation

IC Component	Process Variation (%)	Temperature Coefficient (%)
R	+/- 15	-/+ 5
C	+/- 20	-/+ 0.5
g_m	+/- 10	-/+ 15
V_t	+/- 10	-/+ 14
r_{out}	+/- 20	+/- 20

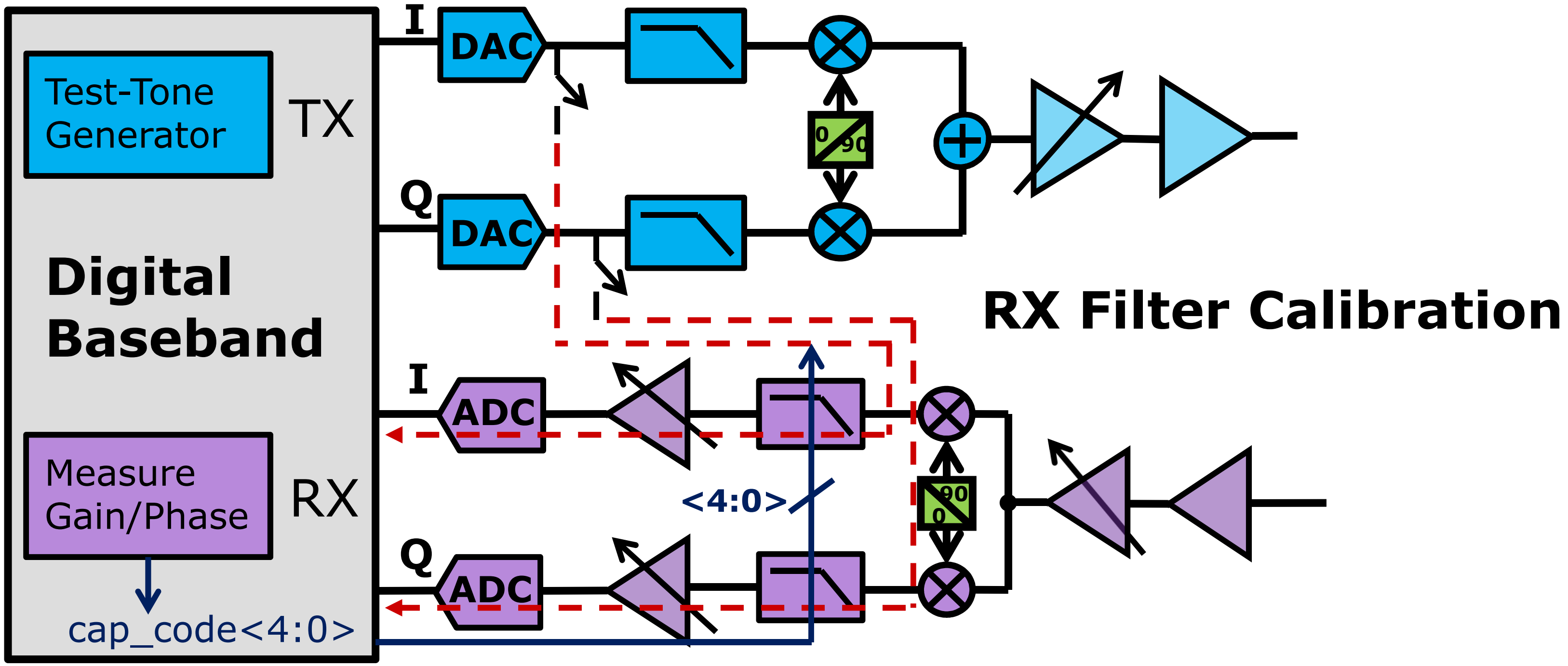
- RC Variation
  - Analog Filter Corner Frequency
- LC variation
  - Tuned amplifier gain (Inductor Q varies too!)
  - VCO Center Frequency
- g_m , r_{out}, V_t variation -> Gain/Linearity/Noise

Active RC Filter Calibration



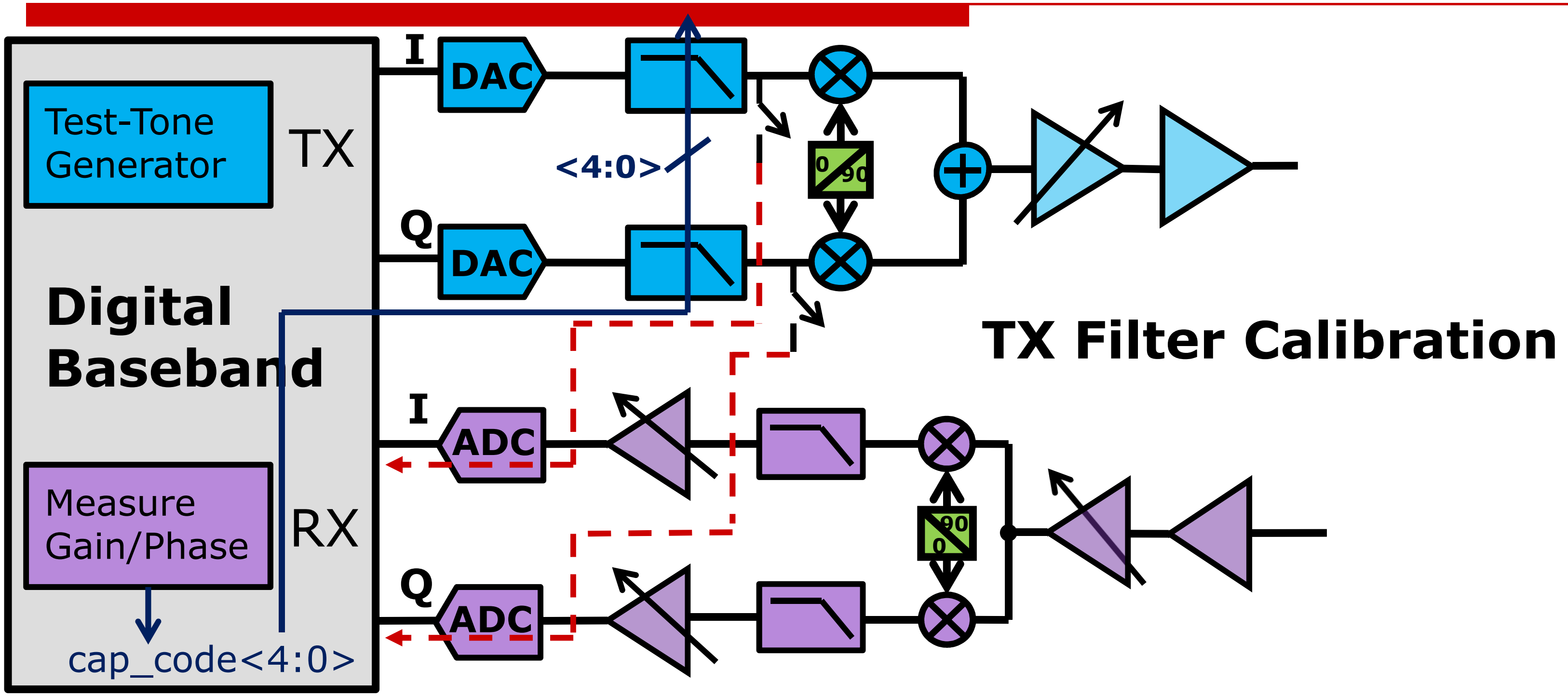
- Direct Tuning of -3 dB BW
  - Generate input tone at desired -3 dB cut-off frequency
  - Compare magnitude of tone at filter output with -3 dB scaled version of the original input tone
  - Tune R or C until the filter gain is -3 dB

Baseband Loopback Calibration



- ☐ Re-use signal path and DSP hardware on SoC
- ☐ Measure gain/phase response of baseband blocks through configurable loopback paths
  - Send signal from TX DAC
  - Read ADC, adjust RC code, and store in registers

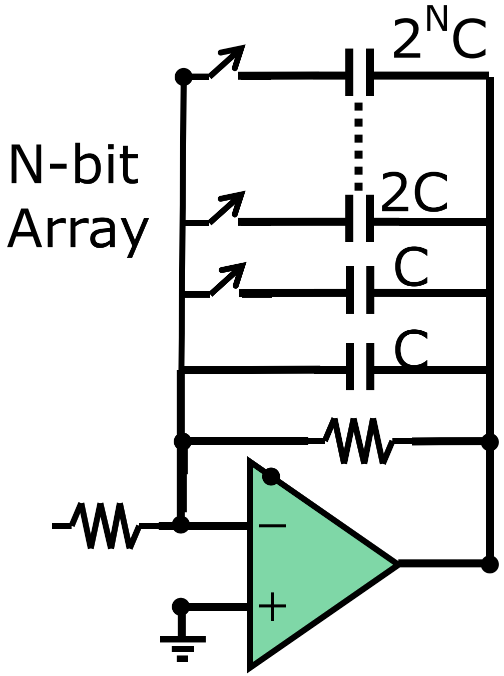
Baseband Loopback Calibration



- No matching issues since signal path is used
- Can send out multiple test tone frequencies
  - Measure gain, phase, group delay
  - Synchronize RX analysis with start of TX
- Calibrate I/Q filter paths separately

Capacitor Array Design

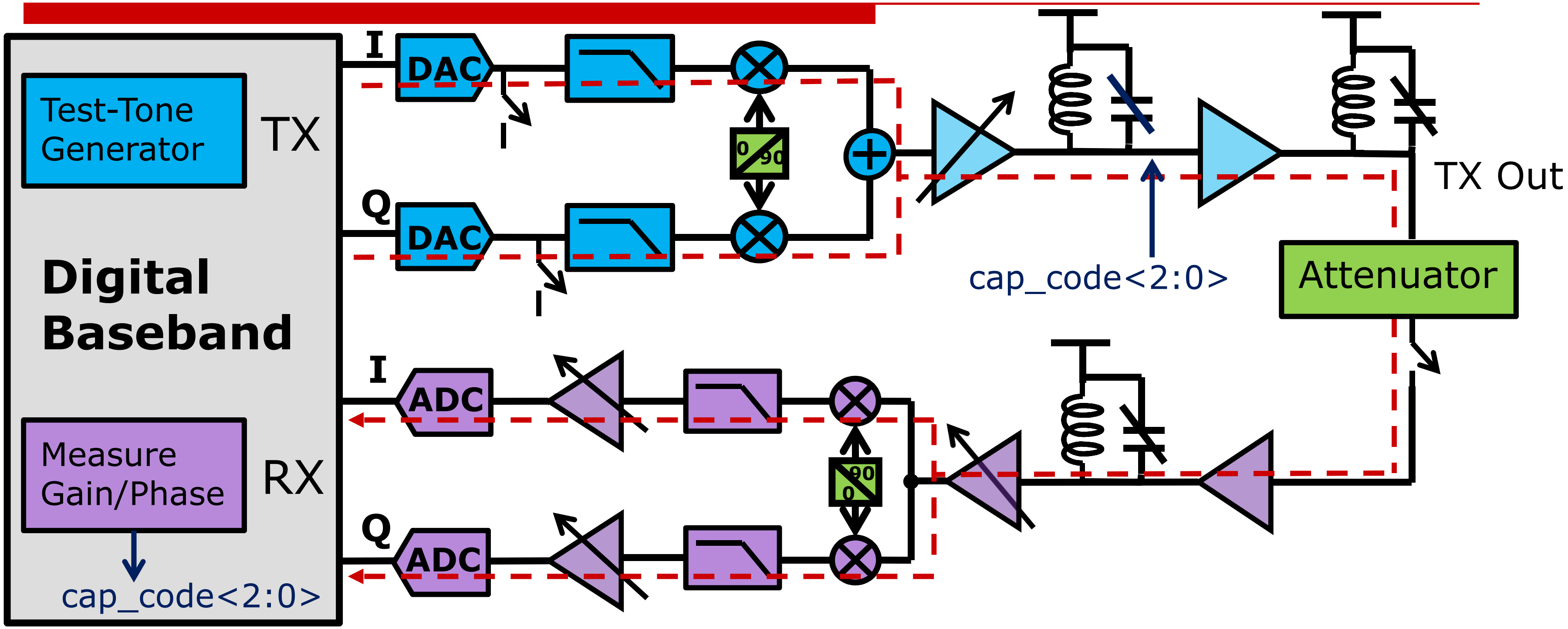
- Capacitor array should be designed to cover maximum process variation of RC
- For typical process variation, N = 5 or 6 to achieve < 1 % accuracy
- Switch should be placed on op-amp virtual ground side
  - Switch parasitics become part of filter design
  - Make sure it always turns on!



Tuning Accuracy (%) =

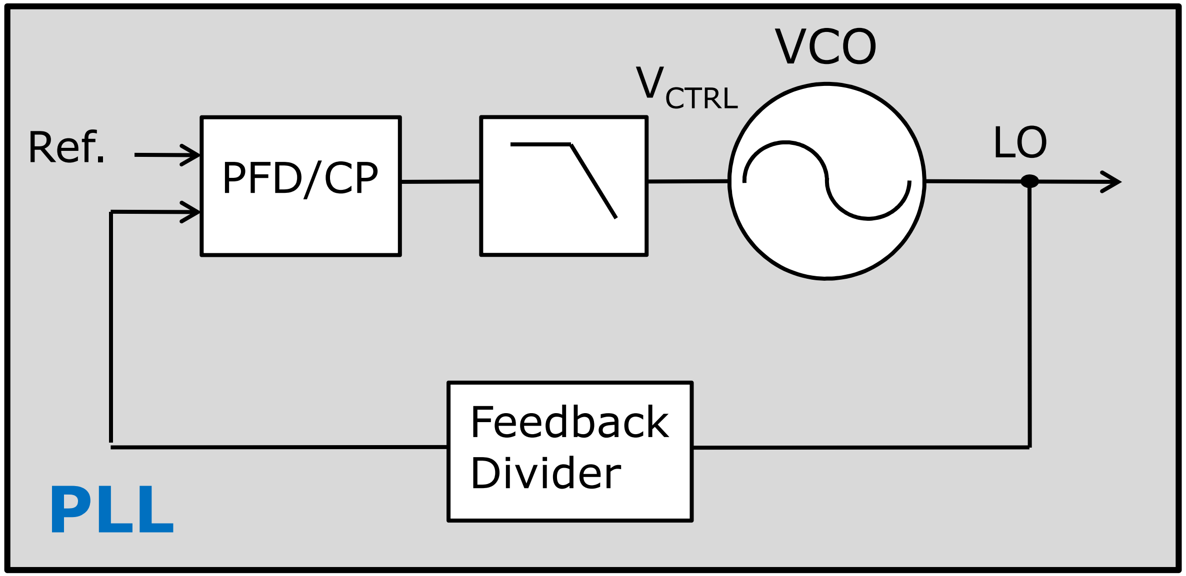
$$\left(\frac{C_{\max}-C_{\min}}{C_{target}}\right)\left(\frac{1}{2^{(N+1)}}\right)\times 100$$

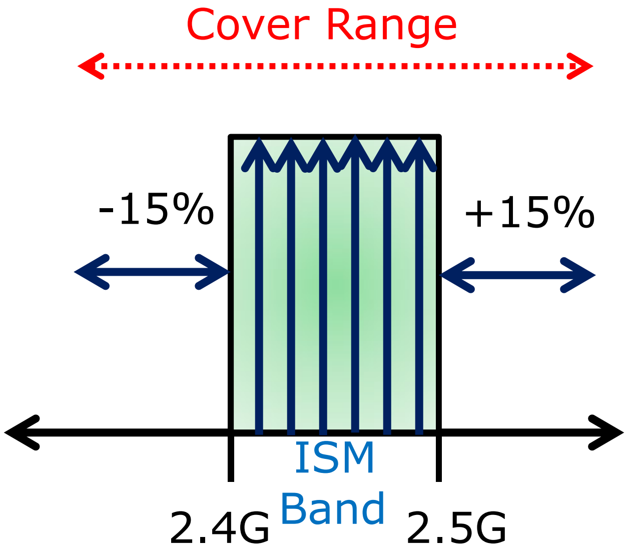
RF Loopback Calibration



- ☐ Global RF loopback path from DAC to ADC
- ☐ LC Tank Calibration
  - Send test tone from DAC and maximize ADC output
- ☐ TX/RX Gain Step Calibration

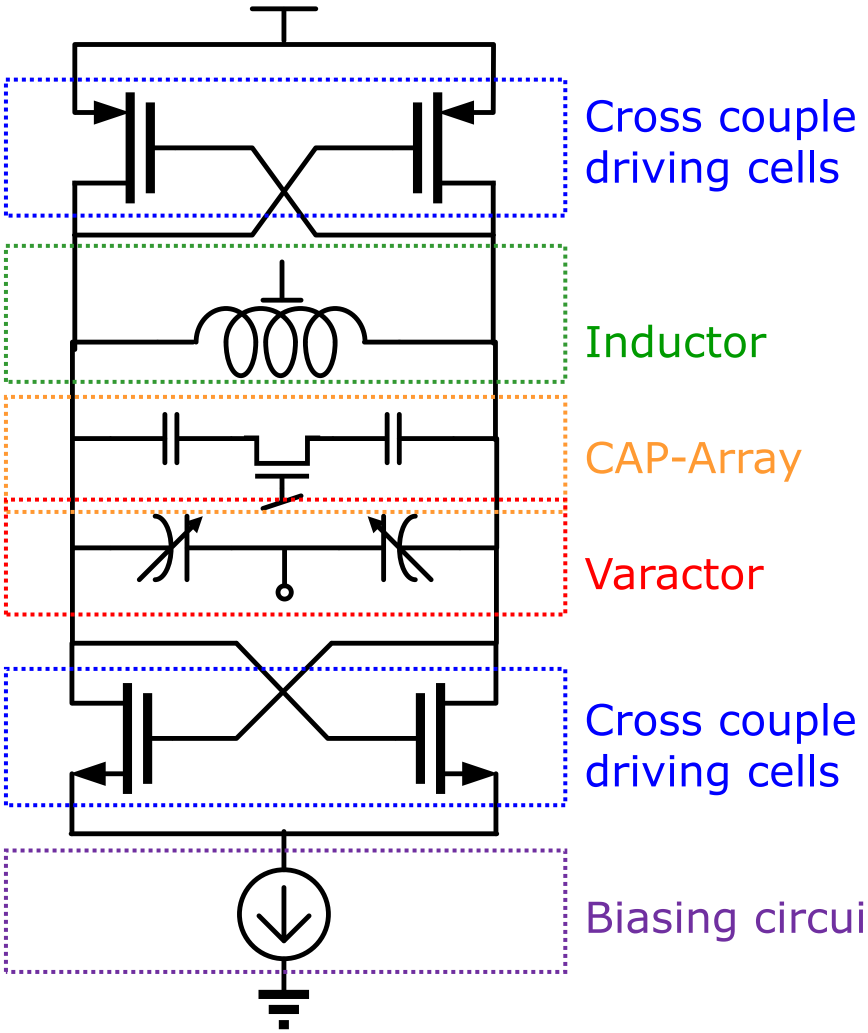
VCO LC Tank Variation





- VCO Center Frequency is function of L and C
- On-chip capacitance has ~ +/- 20% variation
- Need to calibrate capacitance in order to guarantee PLL lock over operating band

VCO Tuning Curves



Cross couple driving cells

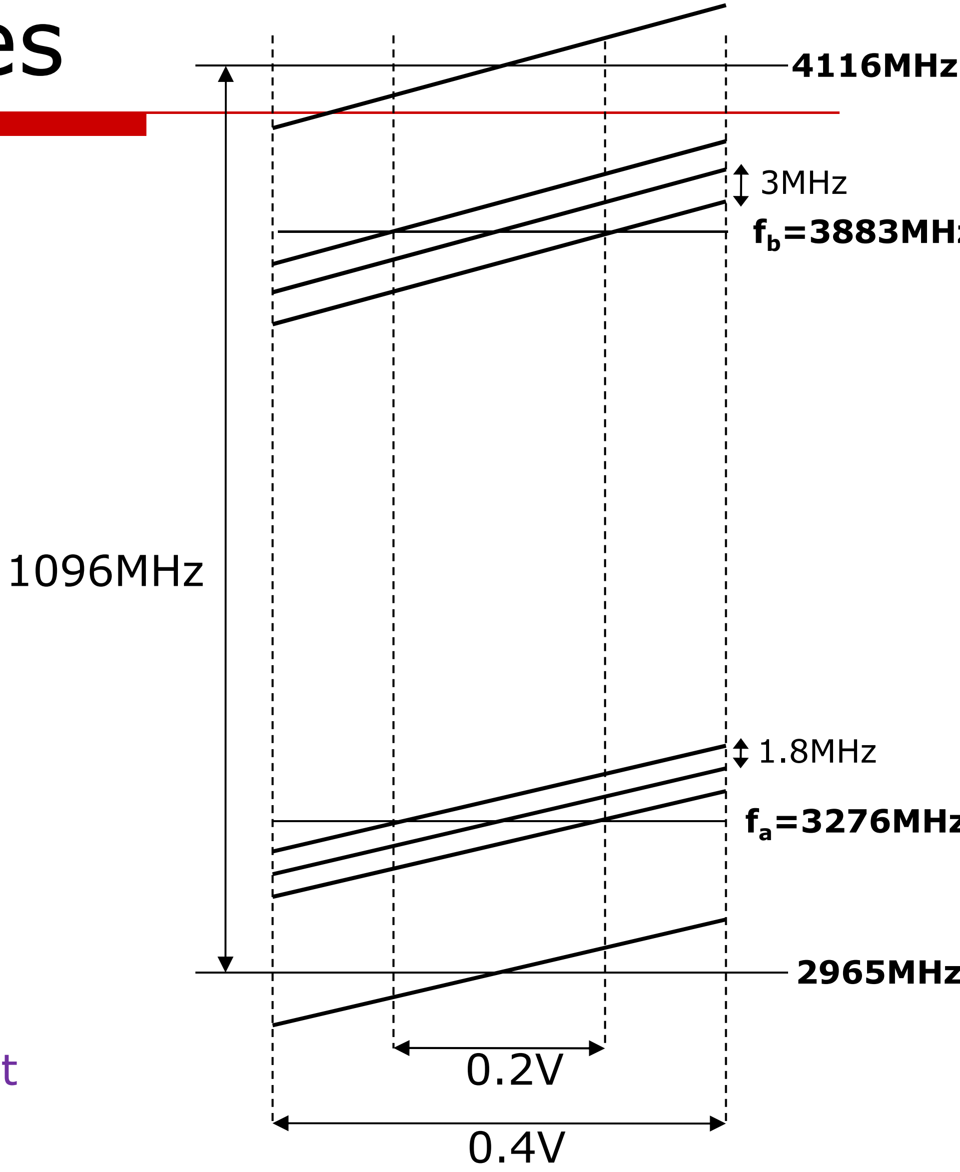
Inductor

CAP-Array

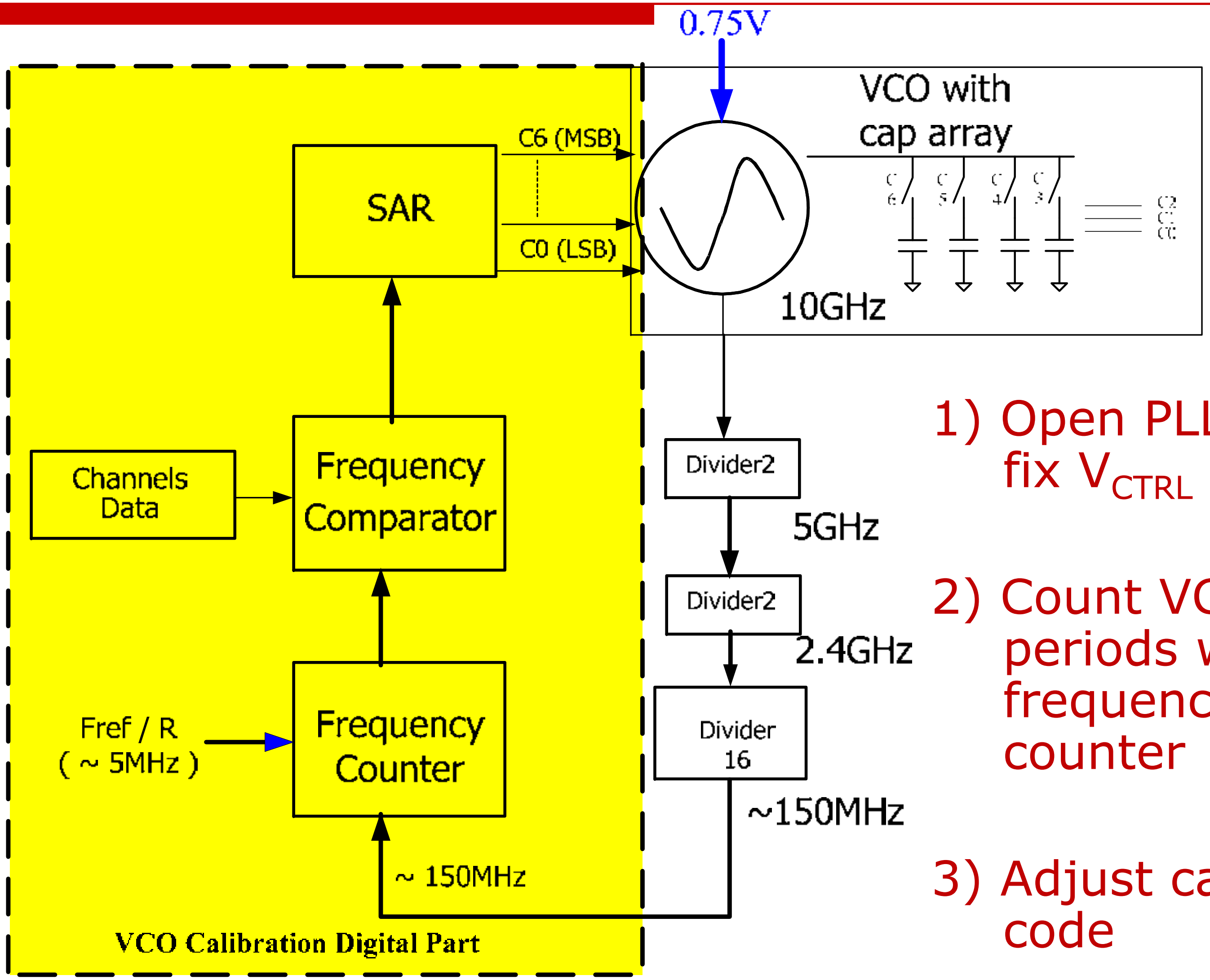
Varactor

Cross couple driving cells

Biasing circuit



VCO Calibration – Open Loop



1) Open PLL and fix V_{CTRL}

2) Count VCO periods with frequency counter

3) Adjust cap code

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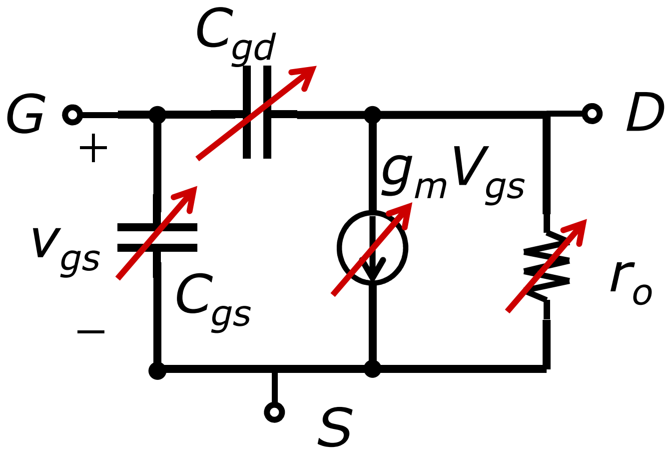
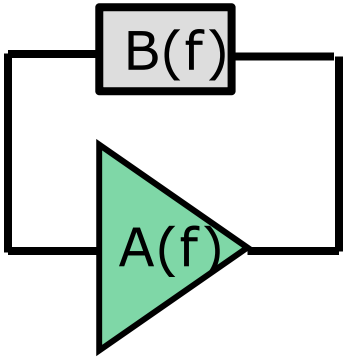
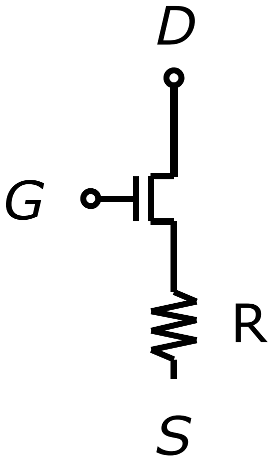
RF Circuit Non-Linearity



MOS Large-Signal Model



Linearized @ low freq.



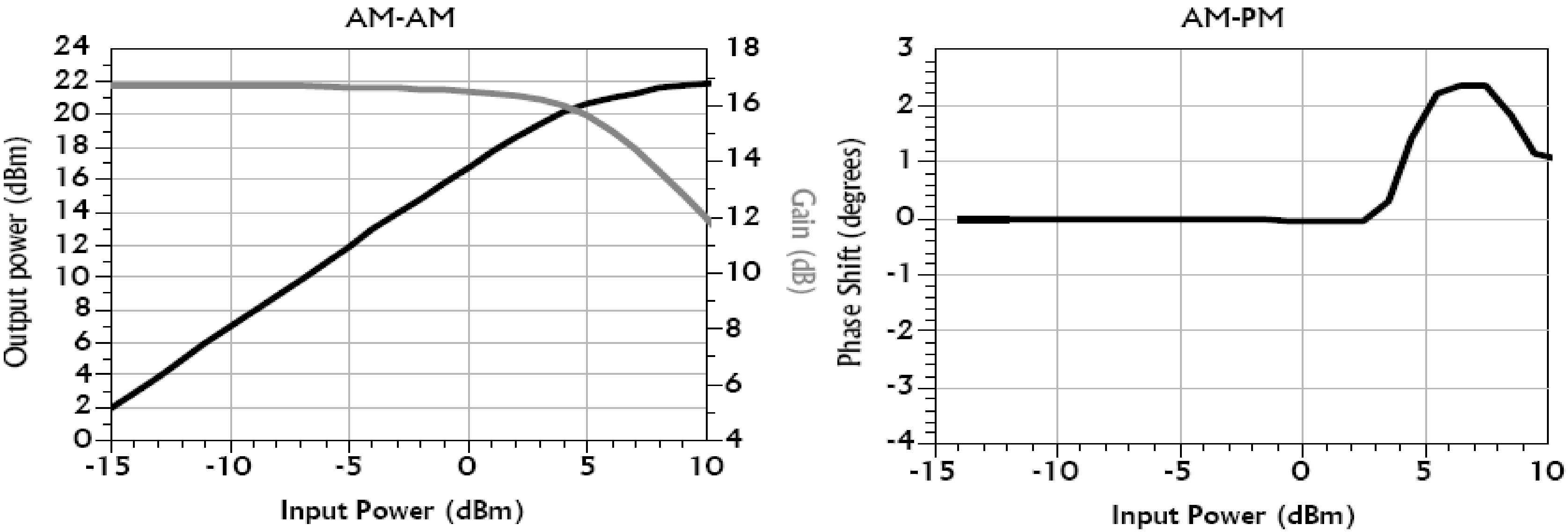
$g_m \approx 1/R$

- Non-linear g_m, C_{gs}, C_{gd}, r_o : $f(V_{gs}, V_{gd}, V_{ds})$
- Closed loop feedback at RF requires high gain-BW product
- Degeneration degrades noise
- Noise/Linearity Tradeoff

Non-Linearity Metrics (AM-AM/AM-PM)

- ☐ AM-AM, AM-PM can characterize linearity
- ☐ Can relate to EVM

$$v_{in}(t)=A(t)\cdot cos(\omega_c t+\phi(t))\Rightarrow v_{out}(t)=G[A(t)]\cdot cos(\omega_c t+\phi(t)+\Phi[A(t)])$$



Non-Linearity Metrics (IM3)

- Approximate input-output with polynomial

$$y = a_1 x + a_2 x^2 + a_3 x^3 + \ldots$$

- Let x be the sum of 2 sinusoids at ω_1, ω_2 and neglecting DC terms and harmonics

$$y(t) = \left(a_1 A + \frac{3a_3 A^3}{4} + \frac{3a_3 A^3}{2} \right) \cos(\omega_1 t) + \left(a_1 A + \frac{3a_3 A^3}{4} + \frac{3a_3 A^3}{2} \right) \cos(\omega_2 t)$$

$$+ \frac{3a_3 A^3}{4} \cos(2\omega_1 - \omega_2)t + \frac{3a_3 A^3}{4} \cos(2\omega_2 - \omega_1)t$$

- 3rd order inter-modulation (IM3) for weakly non-linear region

$$20 * \log 10 \left(\frac{3a_3 A^2}{4a_1} \right)$$

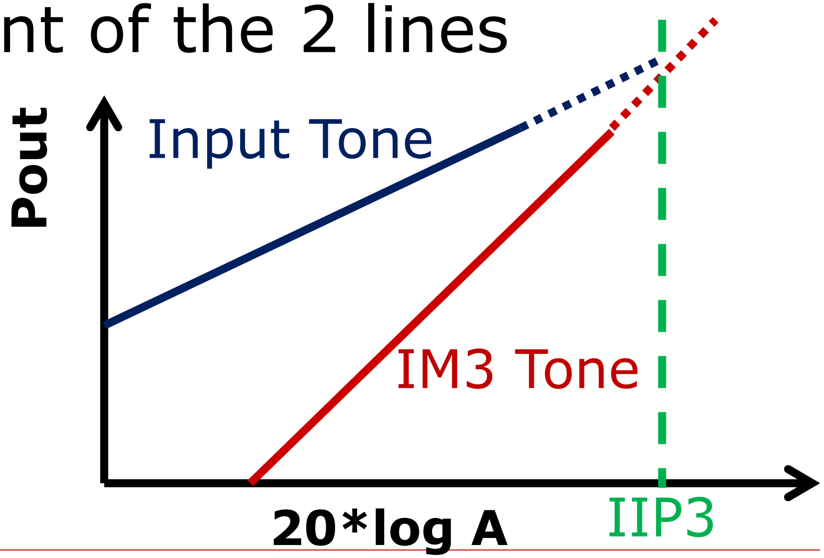
[Razavi]

Non-Linearity Metrics (IP3)

- ☐ Define IP3 as performance metric for linearity using 2-tone test
- ☐ Plot input tones and IM3 products on log scale as function of input amplitude A
- ☐ IP3 is the intercept point of the 2 lines

$$|a_1| A_{IP3} = \frac{3}{4} |a_3| A_{IP3}^3$$

$$A_{IP3} = \sqrt{\frac{4|a_1|}{3|a_3|}}$$

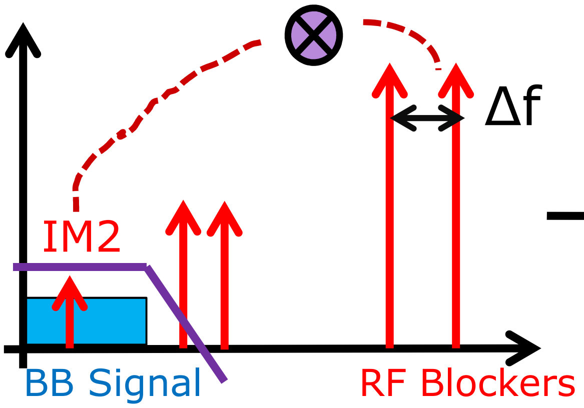


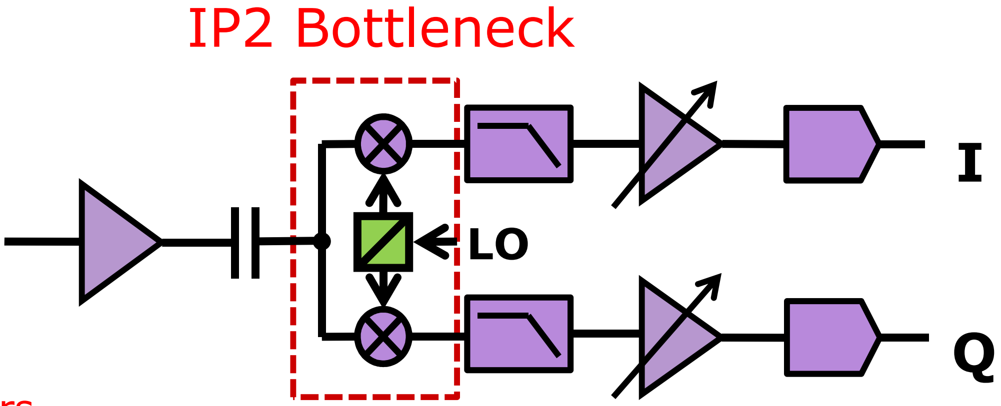
Non-Linearity Metrics (IM2, IP2)

- Other metrics can be similarly derived

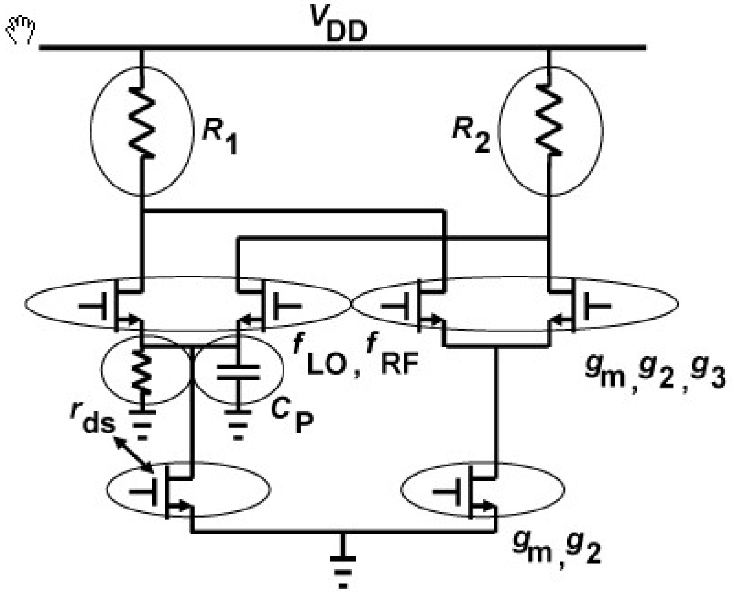
$$IM2 = 20 * \log 10\left(\frac{a_2 A}{a_1}\right) \qquad IP2 = \left(\frac{a_1}{a_2}\right)$$

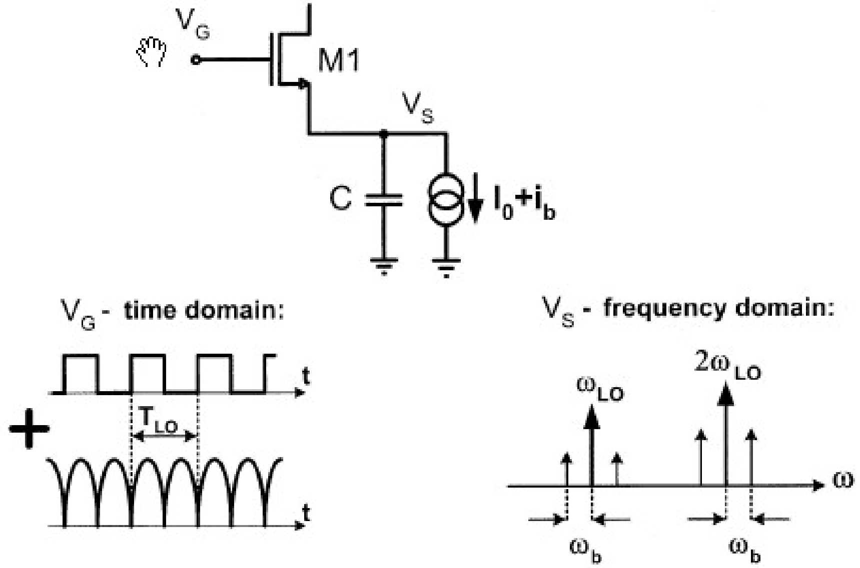
- In differential circuits, 2nd order distortion creates common-mode IM2. Circuit mismatch converts CM to DM signal path.





IM2 Distortion Mechanisms

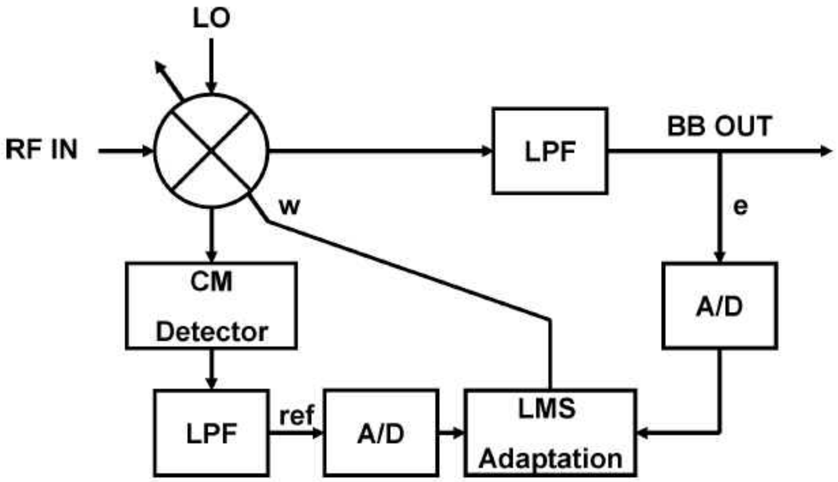


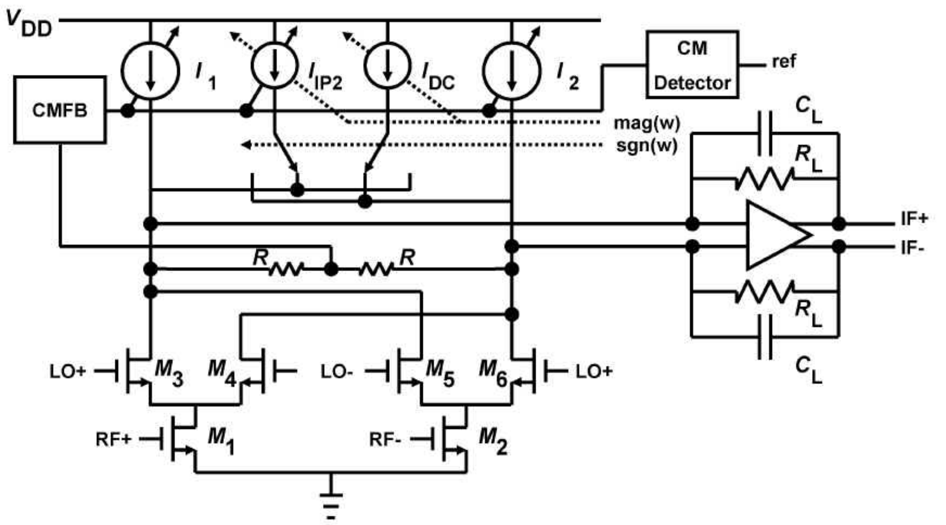


- Input g_m IM2 can be eliminated in principle (AC coupling to switches)
- Mixer switch non-linearity is intrinsic and determines the max achievable IIP2
  - Parasitic capacitance increases distortion current
  - Mismatch in switch and load impedance

[D. Manstretta, JSSC 2003]

IM2 Calibration Technique





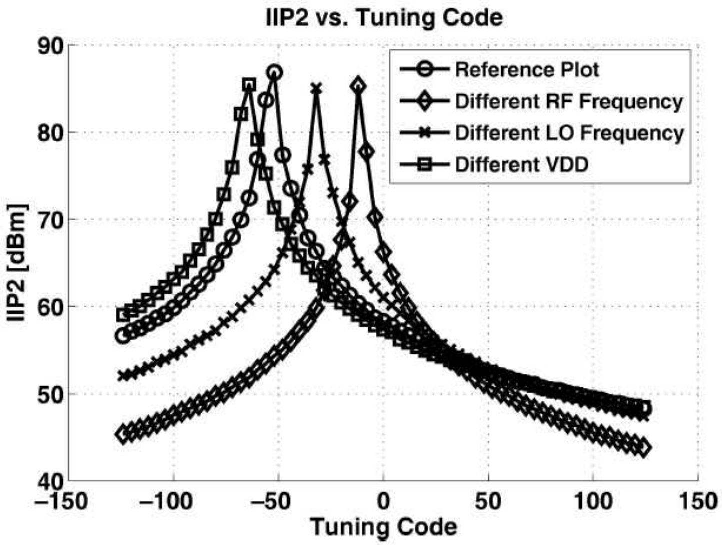
- Exploit existing common-mode distortion and tune circuit to convert part of CM to DM
- Define ratio of DM IM2 to CM IM2

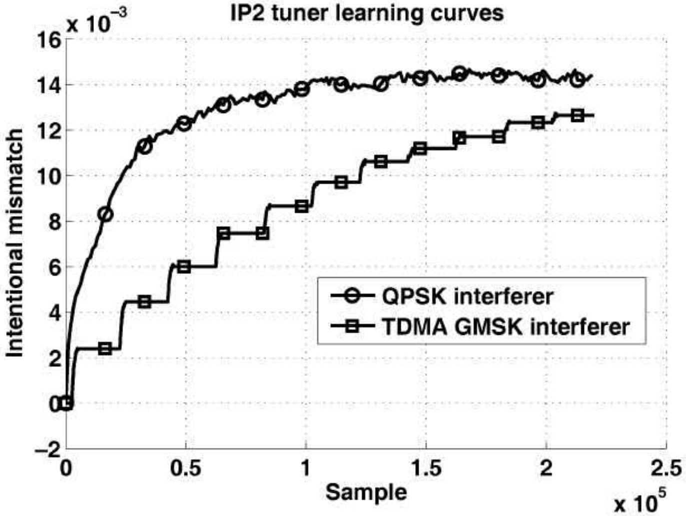
$$\Delta_{IMD2}^{effective} = \frac{i_{diff,imd2}}{i_{cm,imd2}}$$

- Detect 2nd order distortion in CM signal and correlate with ADC output to estimate DM IM2
- Can detect distortion hidden in noise

[K. Dufrene, JSSC 2008]

IP2 Tuner

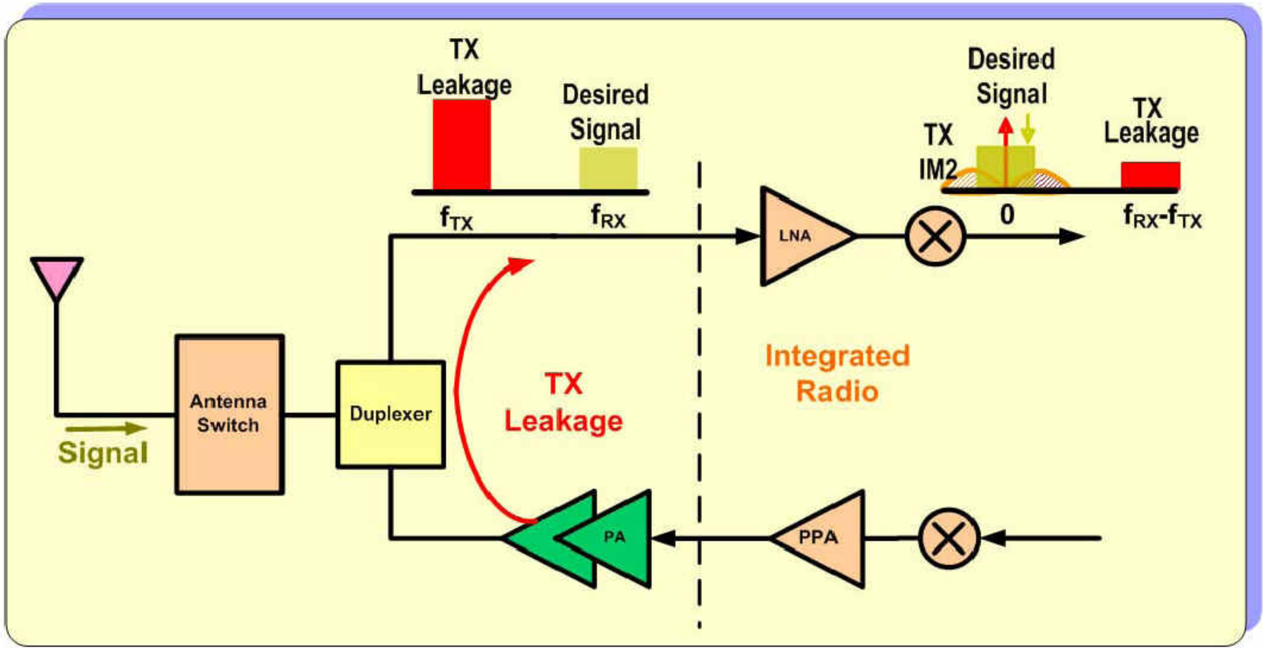




- Adaptive background calibration does not rely on test tones , for continuous reception
- Limitations
  - Tuning Code is a function of RF frequency
  - Convergence speed depends on input signal and cannot instantaneously track changes in interferers

[K. Dufrene, JSSC 2008]

IIP2 Calibration (SAW-less WCDMA)



- ☐ Leakage of TX modulated signal degrades NF
  - Envelope Demodulation due to RX IIP2
  - Inter-modulation (IM2) with RX blocker
- ☐ Eliminate SAW filter using IIP2 calibration

[Kahrizi, RWS 2008]

WCDMA IIP2 Requirement

- ☐ Maximum allowable noise power P_N

  P_N = Sensitivity + Gp - Eb/Nt

  = -117 dBm + 25 dB – 7 dB = - 99 dBm

- ☐ If IM2 must be 14 dB < P_N, then IM2 @ RX is

  $P_{IM2} = P_N - 14\ \text{dB} - IL_{Duplexer} = -99 -14 -2 = -115\ \text{dBm}$

- ☐ With output power P_{TX} and duplexer isolation

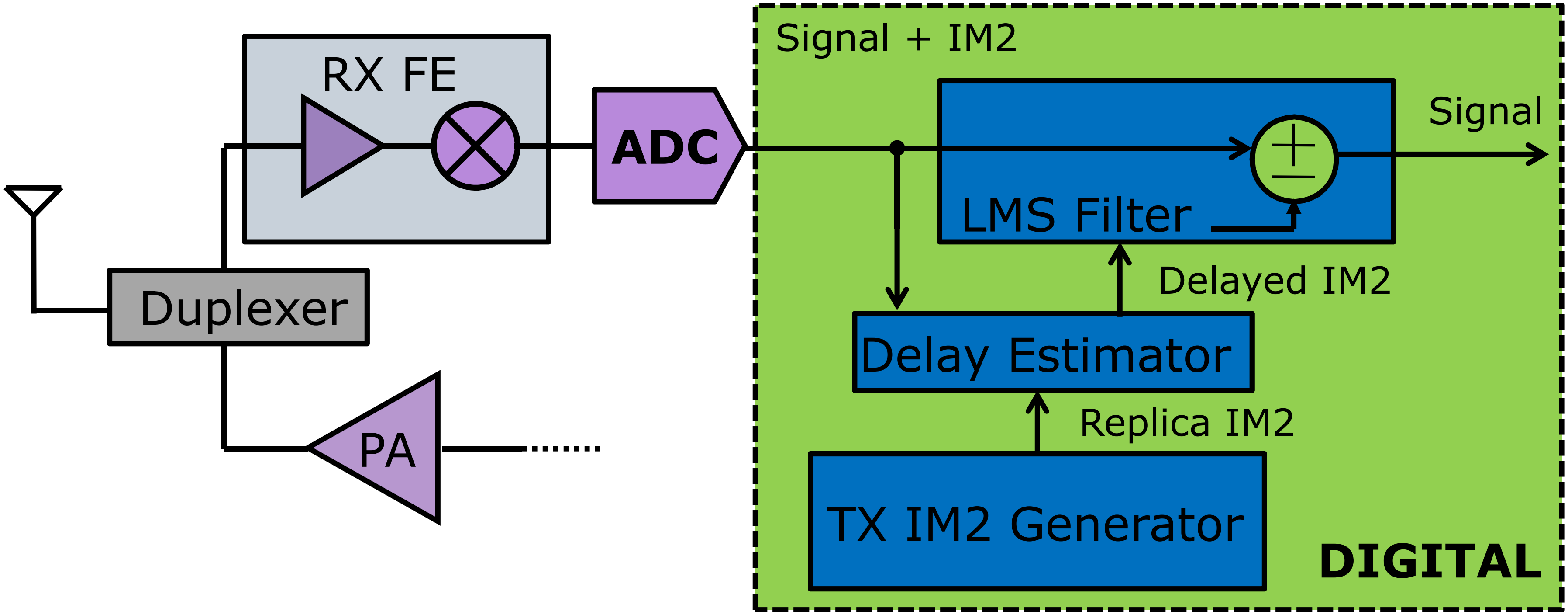
  $IIP2 = 2(P_{TX} - \text{Isolation}_{Duplexer}) - P_{IM2} - \text{Correction}$

  = 2(24 – 50) – (-115) – 13 = 50 dBm

Correction = (WCDMA IM2 in Signal BW)/(IM2 due to 2 tones)

[Kahrizi, RWS 2008]

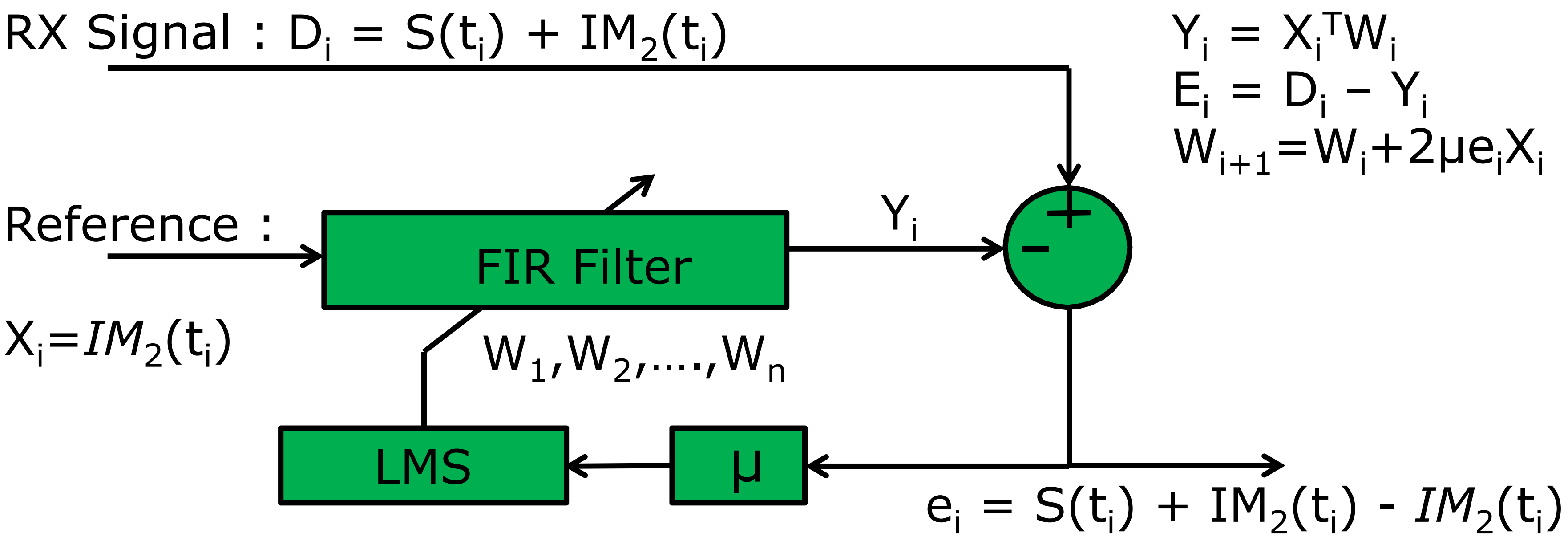
Adaptive Digital IIP2 Calibration



- ☐ Generate reference IM2 signal in digital
- ☐ Estimate delay between RX and TX with peak correlation
- ☐ LMS Filter adaptively subtracts IM2 from signal

[Kahrizi, RWS 2008]

IM2 Cancellation using LMS Filter



RX Signal : $D_i = S(t_i) + IM_2(t_i)$

$Y_i = X_i^T W_i$
$E_i = D_i - Y_i$
$W_{i+1} = W_i + 2\mu e_i X_i$

Reference :

$X_i = \mathit{IM}_2(t_i)$

FIR Filter

Y_i

$+$
$-$

$W_1, W_2, \ldots, W_n$

LMS

μ

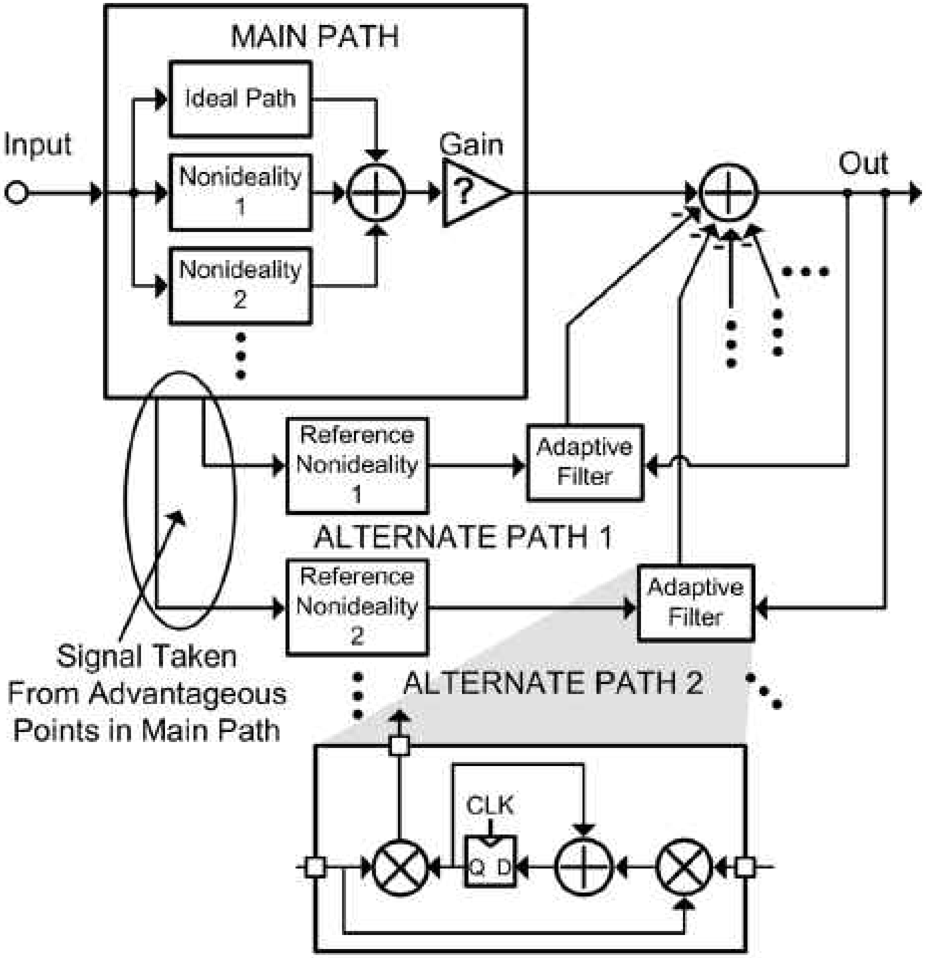
$e_i = S(t_i) + IM_2(t_i) - \mathit{IM}_2(t_i)$

- FIR coefficients are adjusted based on correlation between reference signal and RX signal
- Filter taps converge to yield minimum mean-squared error at output

[Kahrizi, RWS 2008]

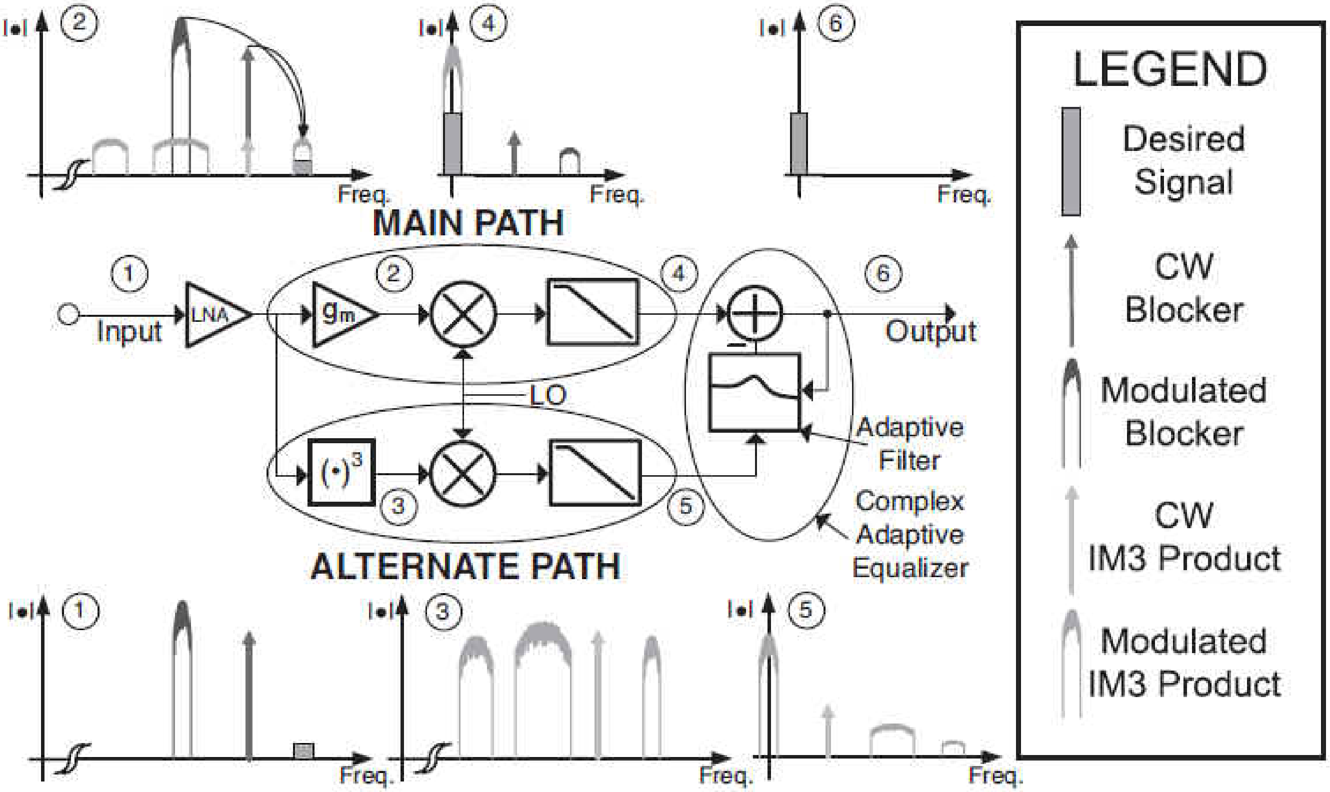
Adaptive Feedforward Cancellation

- ☐ General technique to generate and cancel non-linearity using LMS adaptive filter
- ☐ By cancelling errors instead of error-producers, alternate path can be narrowband
- ☐ Multiple LMS loops cancel interferers in parallel



[Keehr, ISSCC 2008]

IM3 Cancellation Receiver



- ☐ Generate cubic non-linearity to create reference IM3 signal
- ☐ After down-conversion and filtering, cancel IM3

[Keehr, ISSCC 2008]

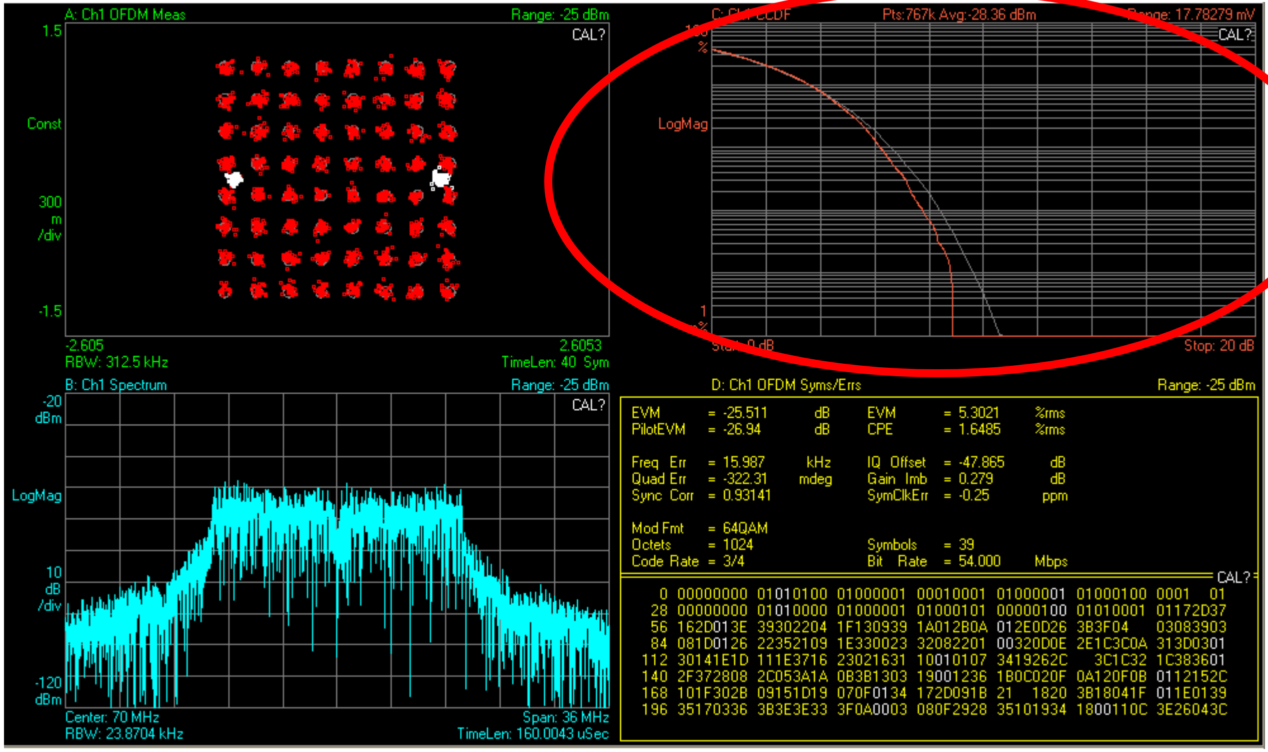
Peak to Average (PAR) Issue

- OFDM has high peak to average ratio
  - A is maximum amplitude for each sub-carrier
  - 52 sub-carriers average power : $P_{ave}=52*(A^2/2)$
  - Sub-carriers in-phase : $P_{peak}=(52*A)^2/2$

$$PAR=10*\log 10\left(\frac{P_{peak}}{P_{ave}}\right)=10*\log 10(52)=17dB$$

- To linearly amplify the peak signal swings
  - Transmit average power at 17 dB back-off from P_{1dB}
- EVM degradation depends on waveform statistics
- In practice, can tolerate clipping at 7-8 dB for -25 dB TX EVM

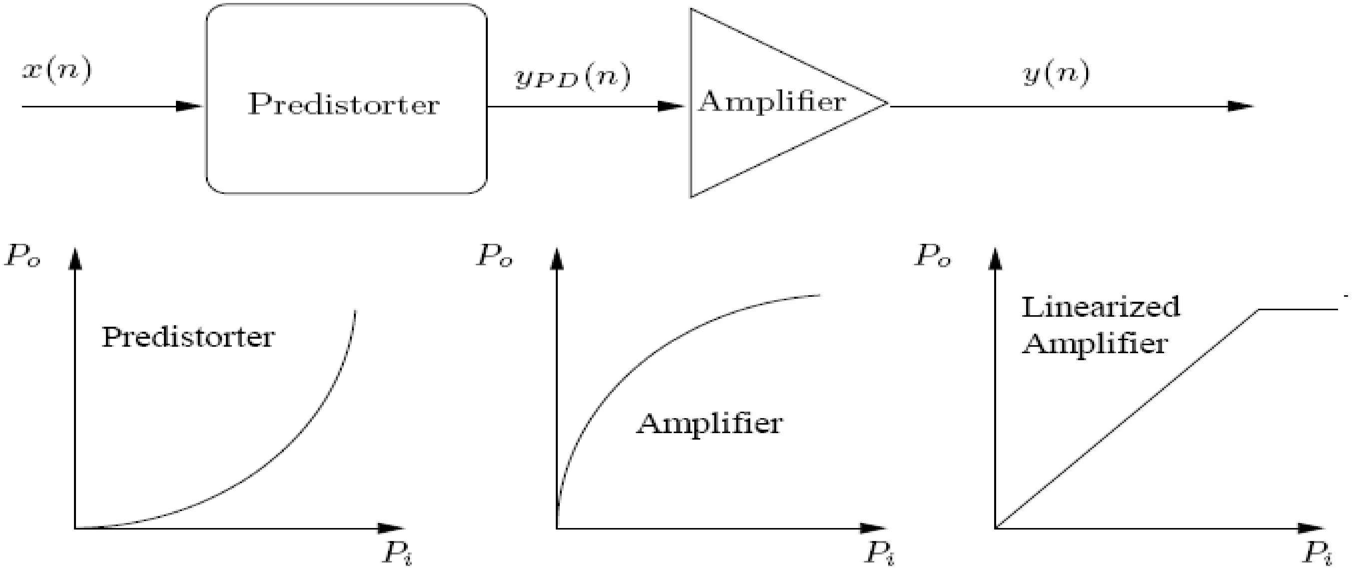
TX EVM Degradation (High PAR)



```
EVM        = -25.511    dB     EVM        = 5.3021    %rms
PilotEVM   = -26.94     dB     CPE        = 1.6485    %rms

Freq  Err  = 15.987     kHz    IQ  Offset = -47.865     dB
Quad Err   = -322.31    mdeg   Gain  Imb  = 0.279       dB
Sync  Corr = 0.93141           SymClkErr  = -0.25       ppm

Mod Fmt    = 64QAM
Octets     = 1024              Symbols    = 39
Code Rate  = 3/4               Bit   Rate = 54.000     Mbps

  0 00000000 01010100 01000001 00010001 01000001 01000100 0001   01
 28 00000000 01010000 01000001 01000101 00000100 01010001 01172D37
 56 162D013E 39302204 1F130939 1A012B0A 012E0D26 3B3F04   03083903
 84 081D0126 22352109 1E330023 32082201 00320D0E 2E1C3C0A 313D0301
112 30141E1D 111E3716 23021631 10010107 3419262C   3C1C32 1C383601
140 2F372808 2C053A1A 0B3B1303 19001236 1B0C020F 0A120F0B 0112152C
168 101F302B 09151D19 070F0134 172D091B 21   1820 3B18041F 011E0139
196 35170336 3B3E3E33 3F0A0003 080F2928 35101934 1800110C 3E26043C
```

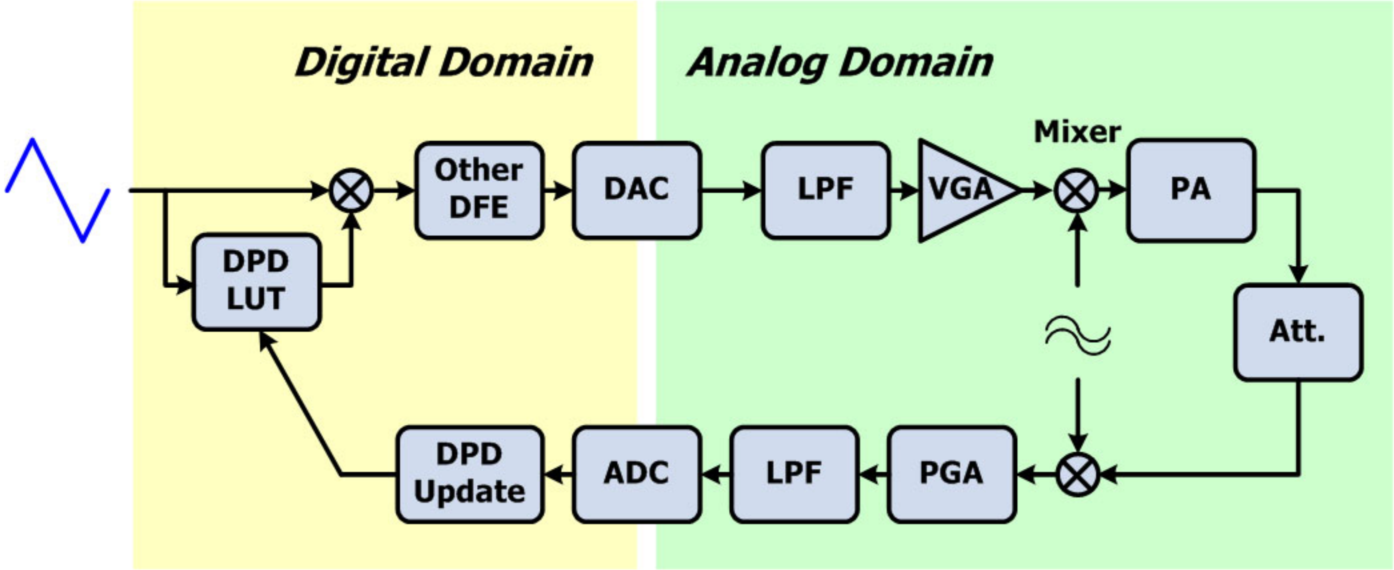
- CCDF shows probability of amplitude levels
- If large number of samples are clipped
  - EVM Degrades

Digital Pre-Distortion (DPD)



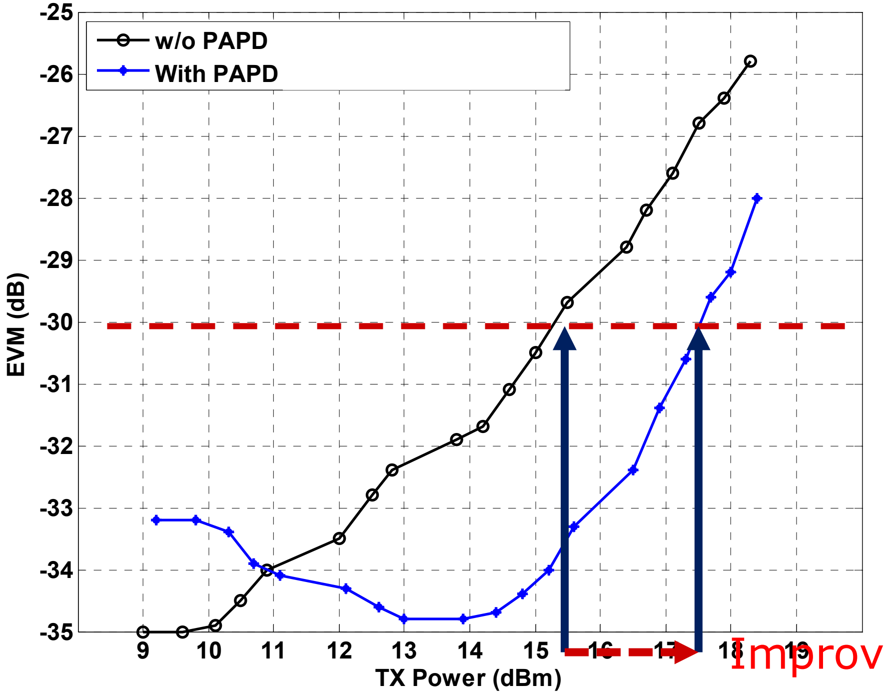
- Look-up table (LUT) Based DPD
  - The inverse PA AM/AM & AM/PM curve are quantized and stored in memory
  - The input signal is then pre-distorted by multiplying the AM/AM gain and rotating the AM/PM phase
  - Memory-less pre-distortion

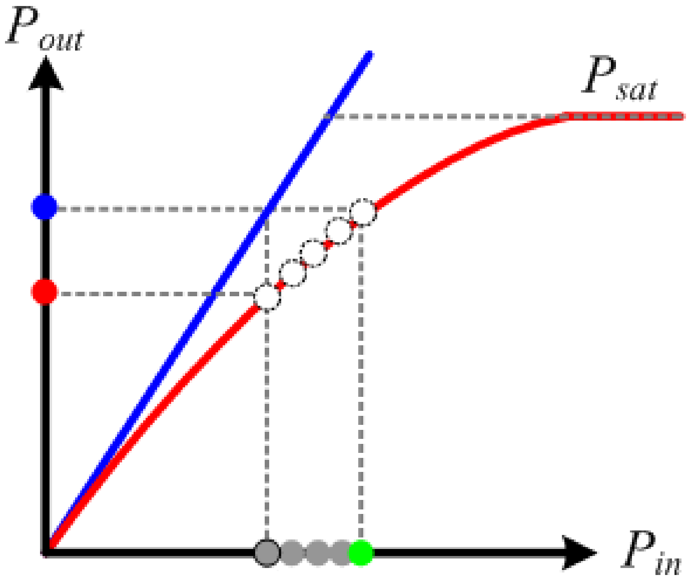
DPD Loopback Path



- ☐ Internal loop-back path from PA output to Rx path is designed for DPD calibration
- ☐ Training signal is passed through Tx chain, loops back through RX, and is re-sampled by Rx ADC.
- ☐ The re-sampled signal is used to construct AM/AM & AM/PM

Benefits of DPD





Improve from 15 dBm -> 17.5 dBm for -30 dB

- Improve TX EVM or increase Pout @ EVM spec
- Reduce required back-off from Psat
  - Reduce PA bias current for given Pout/EVM
- Improve spectral mask

Practical Issues

- When to run calibration?
  - Power-Up
  - System Idle Periods, Between Tx<->Rx, Background?
  - How long will calibration take? Need to re-calibrate?
- Will calibration parameter change?
  - Temperature, Voltage
  - Antenna variation
  - AGC Gain Change, LNA Gain Change
- What limits the calibration "noise floor"
  - Step size of the adjustment element (DAC, Capacitor)
  - Resolution of the measurement (bit length, noise)
  - Non-idealities in the calibration path

Further Topics for Study

- RF Built-in Self Test (RF BIST)
  - Digital BB + TX = Signal Generator
  - RX + Digital BB = Spectrum Analyzer
  - Loopback Measurements (EVM, IM3, Gain steps, etc.)
  - Save test cost (equipment and time)
- Polar Transmitters – synchronize AM, PM paths
- Closed-loop Power Control
- Envelope Tracking
- Co-existence of wireless standards

Conclusion

- Digital Calibration can correct fundamental device and circuit limitations such as mismatch, variation, and non-linearity to optimize key radio parameters such as EVM and IP2/IP3. This results in improved wireless system performance and enables cost/area reduction of integrated radios.
- Trends in device scaling and higher RF EVM for increased data rates will require continued advances in digital calibration

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