RF impairments in MIMO OFDM:
impact and digital compensation

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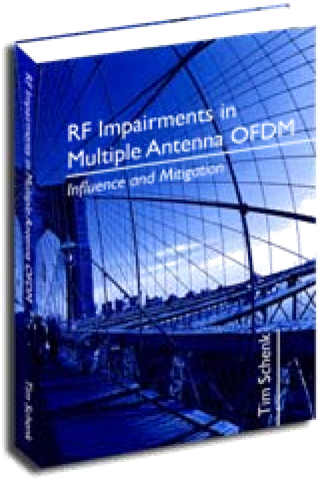
Friday June 8, 2007

Who’s Tim Schenk

- Research Scientist @ Philips Research, Eindhoven, Connectivity Systems and Networks – department
- 1996 – 2002: Ir./M.Sc. studies @TU/e
- 2002 – 2006: Ph.D. project @TU/e and Agere Systems

  Part of EZ-funded project BroadBand Radio@Hand: TU/e, KPN, TNO ICT, Philips Research and Agere Systems.
- Research interests:
  - Wireless communications
  - Crosslayer design
  - Analogue/digital signal processing
  - Low-power techniques
  - Sensor networks



Outline of this part of the course

- System introduction
  - Digital wireless communication system
  - Orthogonal frequency division multiplexing (OFDM)
  - Multiple antenna (MIMO) systems
  - System implementation
- RF imperfections
  - Influence of RF impairments
  - Digital compensation approaches
    - Carrier frequency offset
    - Phase Noise
    - IQ imbalance
    - Nonlinearities

Part I
11.00 – 12.00
Lunch
12.00 – 13.00
Part II
13.00 – 14.00
Instruction
14.00 – 14.30
Part III
14.30 – 15.30
Instruction
15.30 – 16.00

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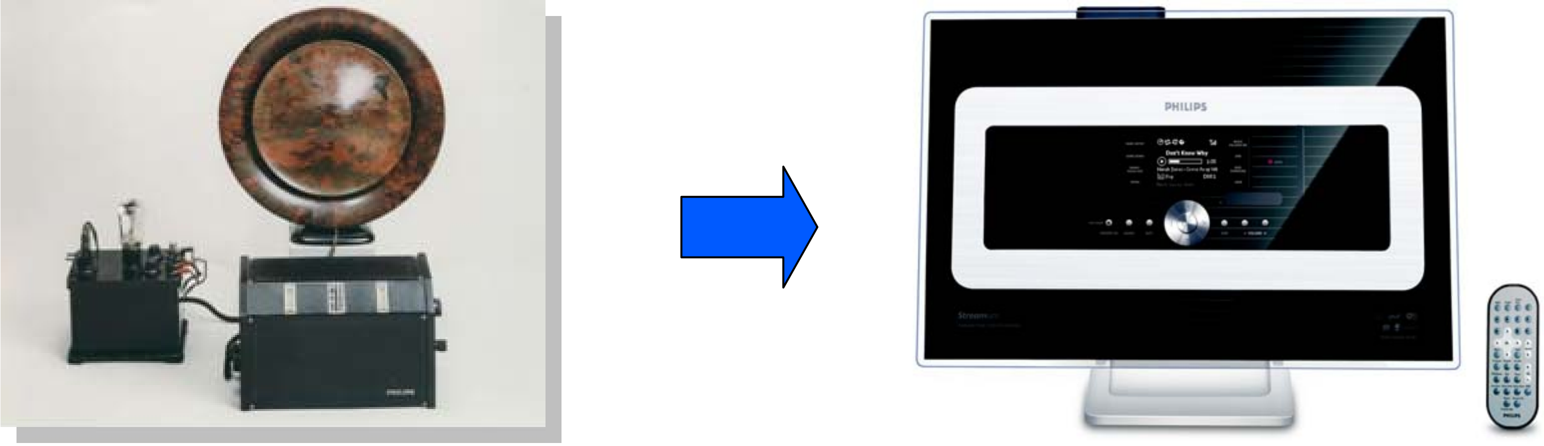
Part I: Channel, OFDM, MIMO and system

- Digital wireless communication system
- Orthogonal frequency division multiplexing (OFDM)
  - Review of the basics
  - System standards
  - Example TX/RX processing: IEEE 802.11a
- Multiple antenna (MIMO) systems
  - Review of the basics
  - Physical interpretation
- System implementation

Digital communication systems (I)



- Ever higher demand for speed in wireless communication systems
- Solution: increase in bandwidth f_s
  → $T = 1/f_\mathrm{s}$ → sample length decreases
- Increase in *spatial resolution* → More of the wireless channel is *observable*

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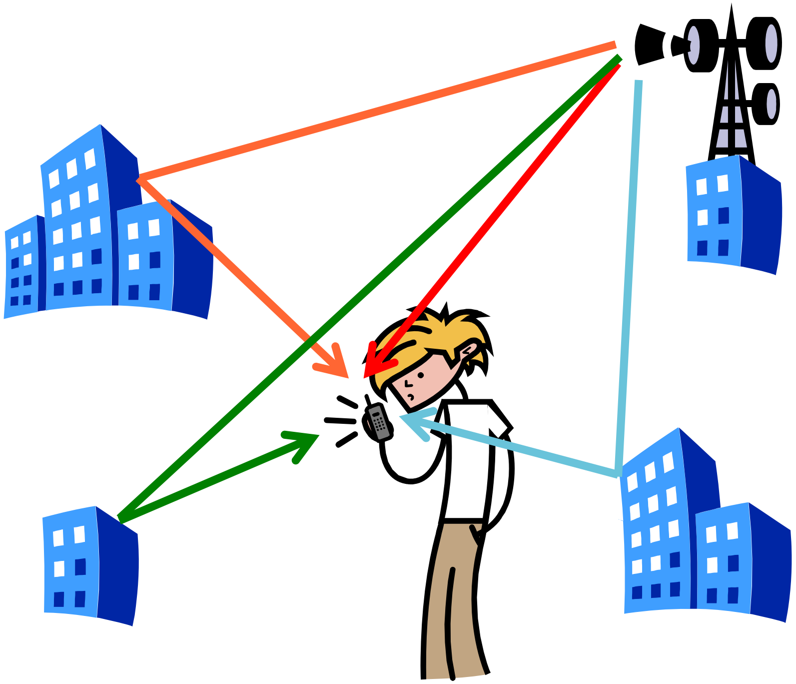
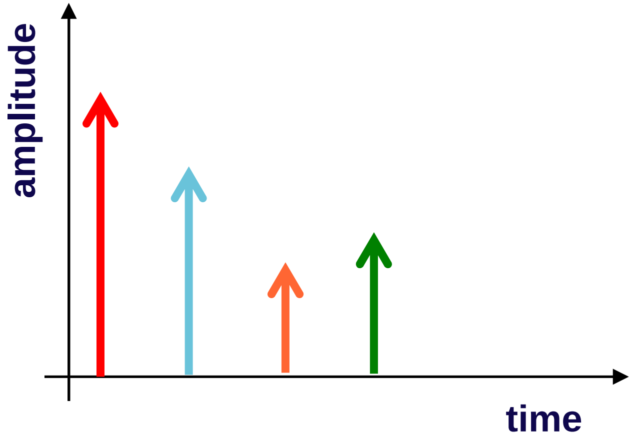
Digital communication systems (II)





Multipath Channel

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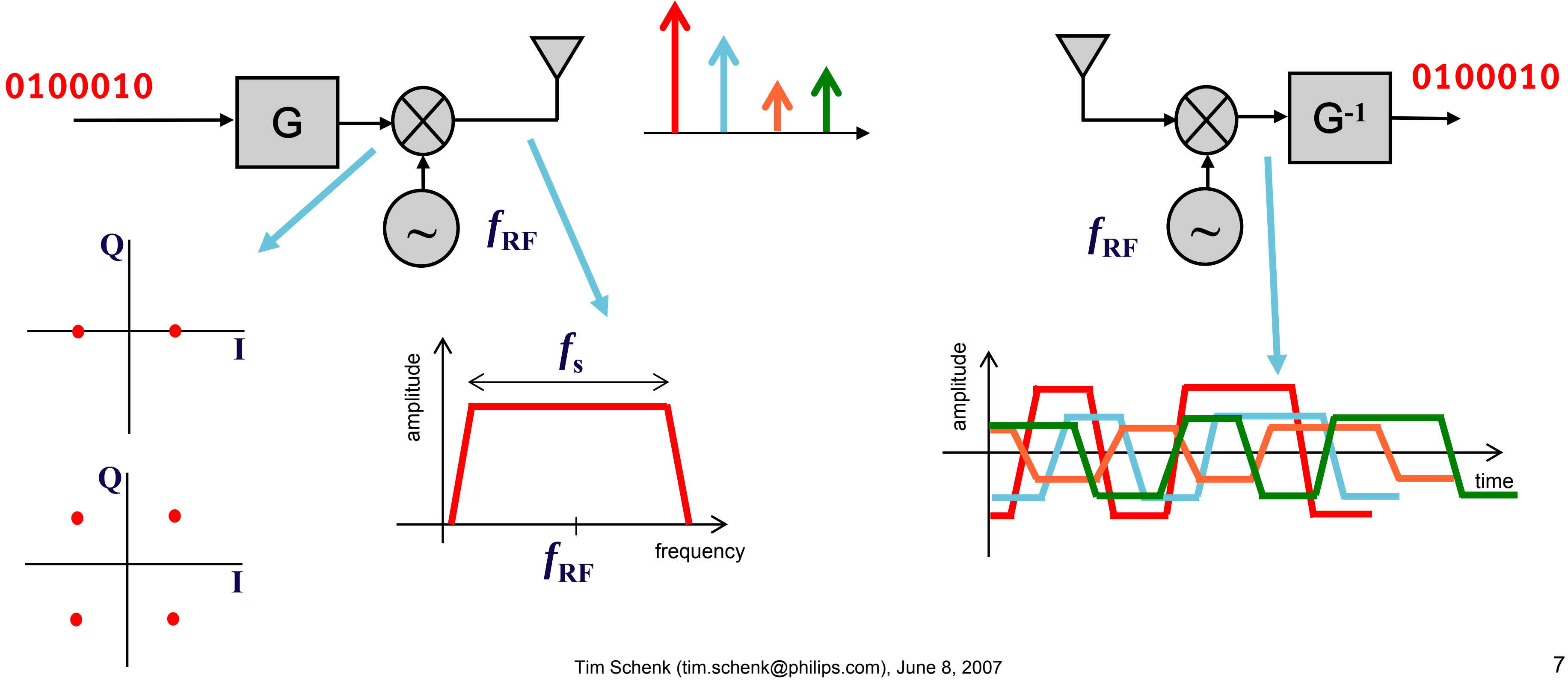
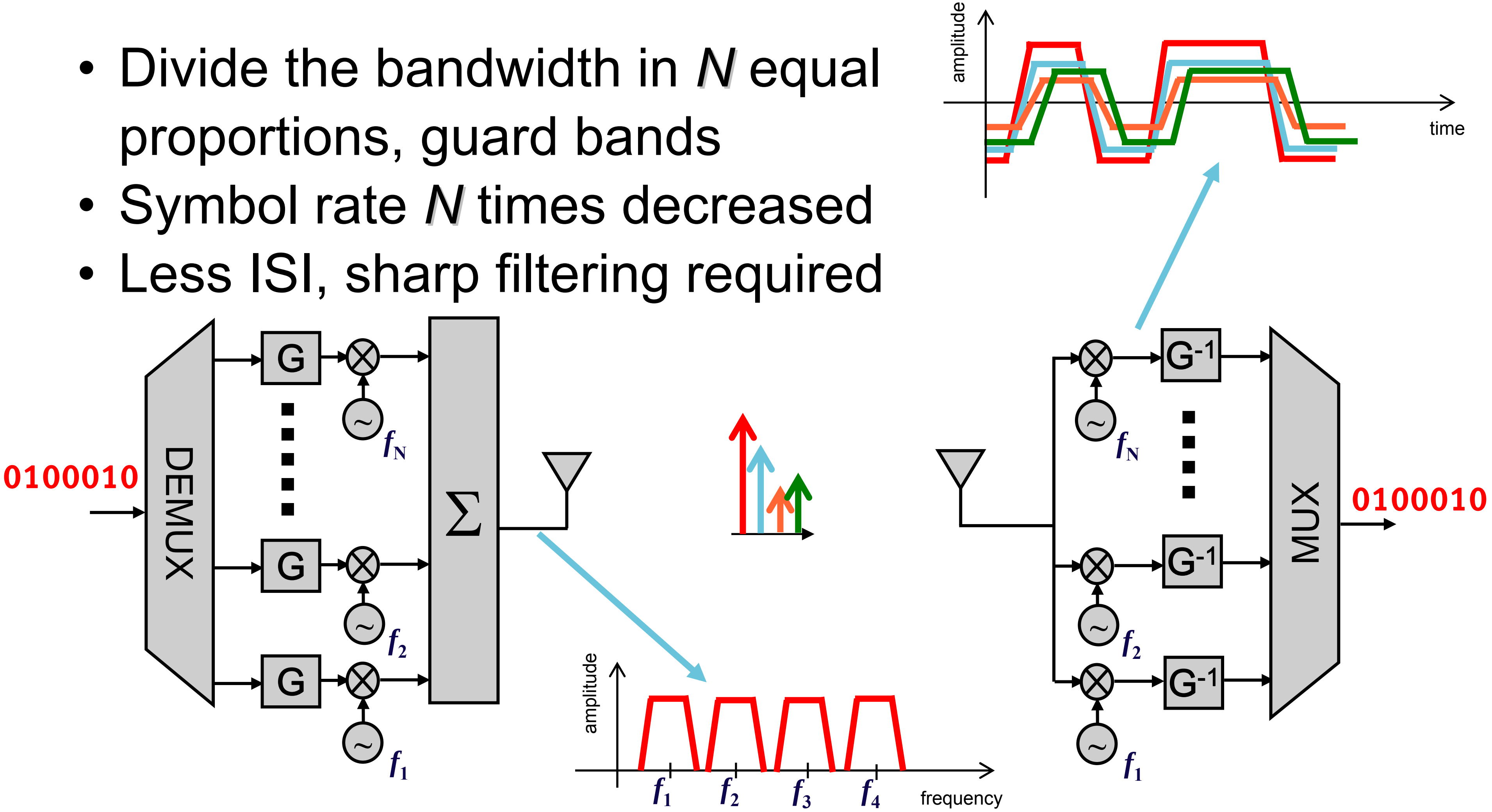
Digital communication systems (III)

- Datarate is proportional to bandwidth
- Linear decrease using complex modulation
- Multipath channel results in inter-symbol interference



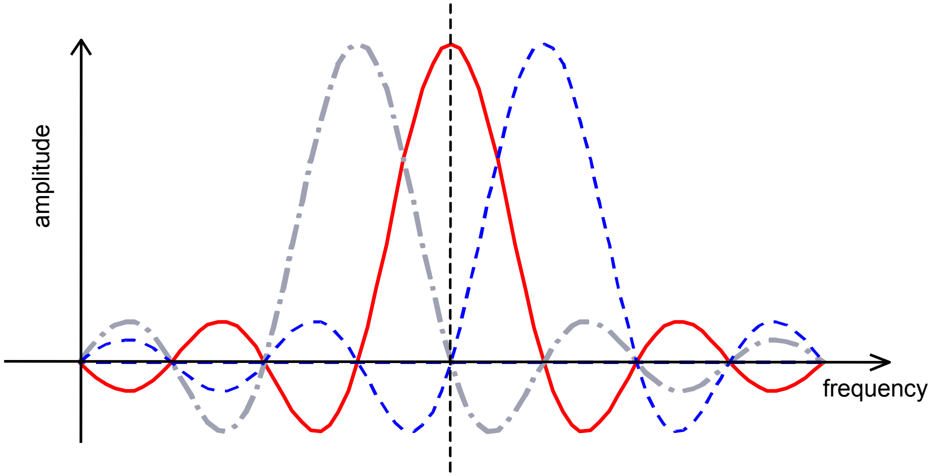
Multicarrier transmission

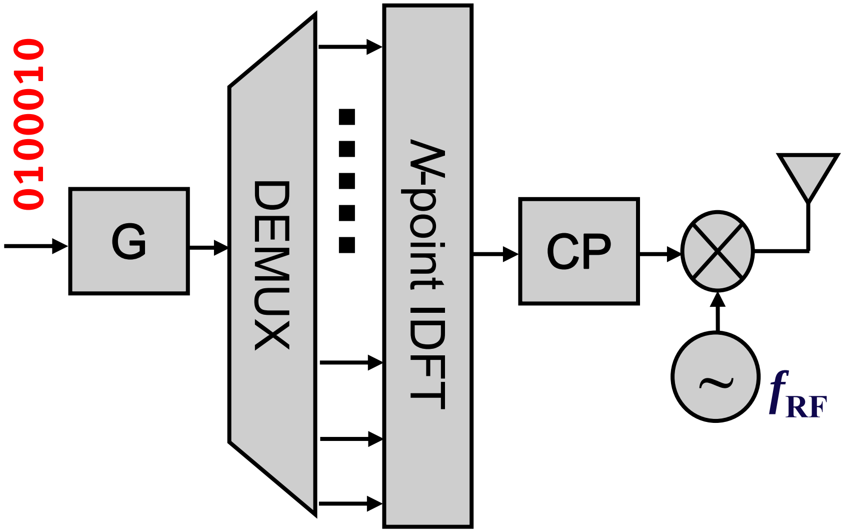
- Divide the bandwidth in *N* equal proportions, guard bands
- Symbol rate *N* times decreased
- Less ISI, sharp filtering required

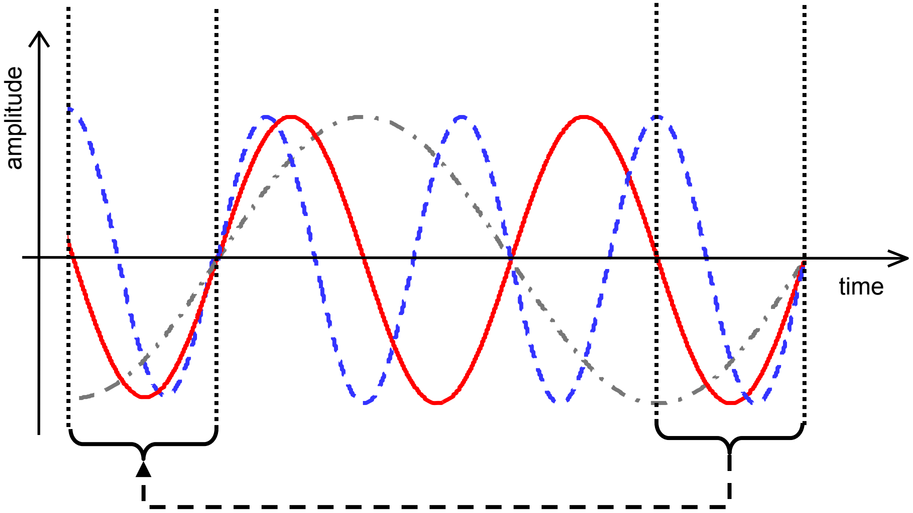


OFDM (I)

- *N* bands closer together due to orthogonality
  → symbol rate/*N*
- Addition of cyclic prefix







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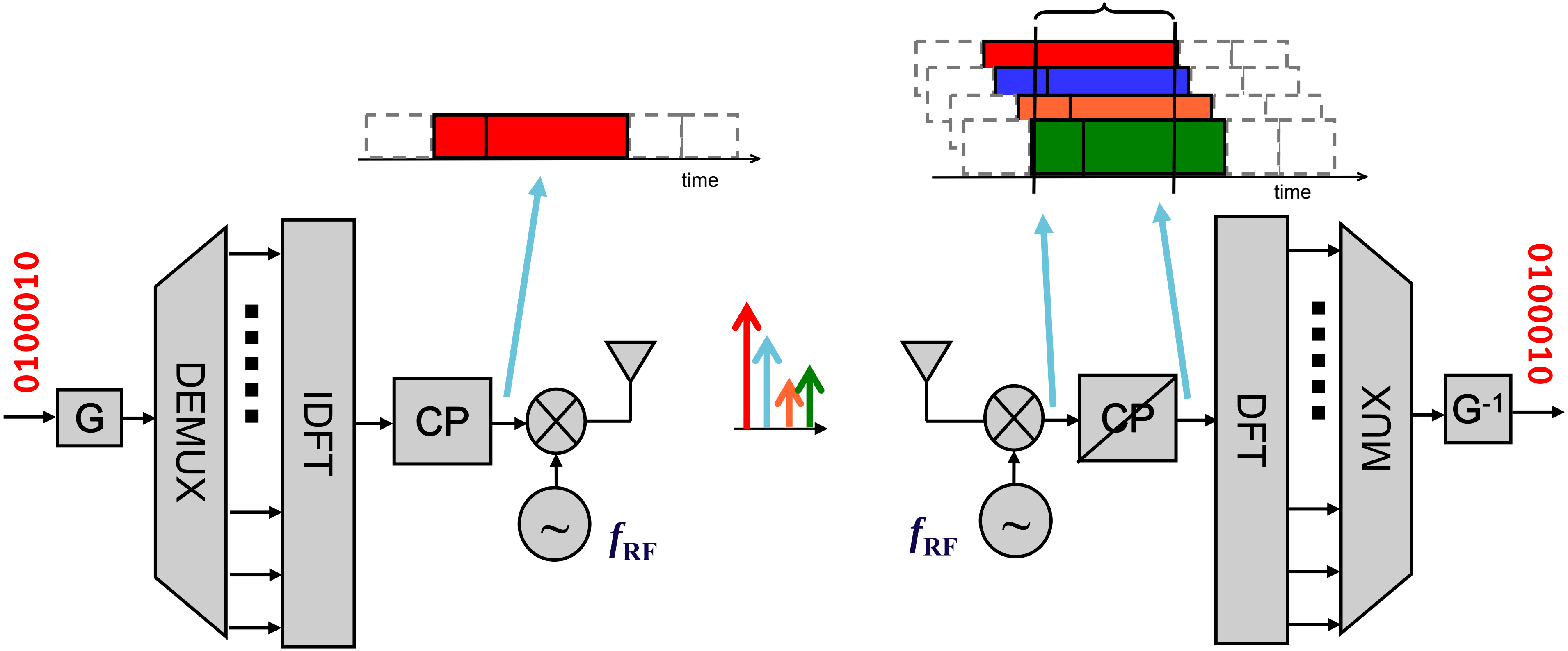
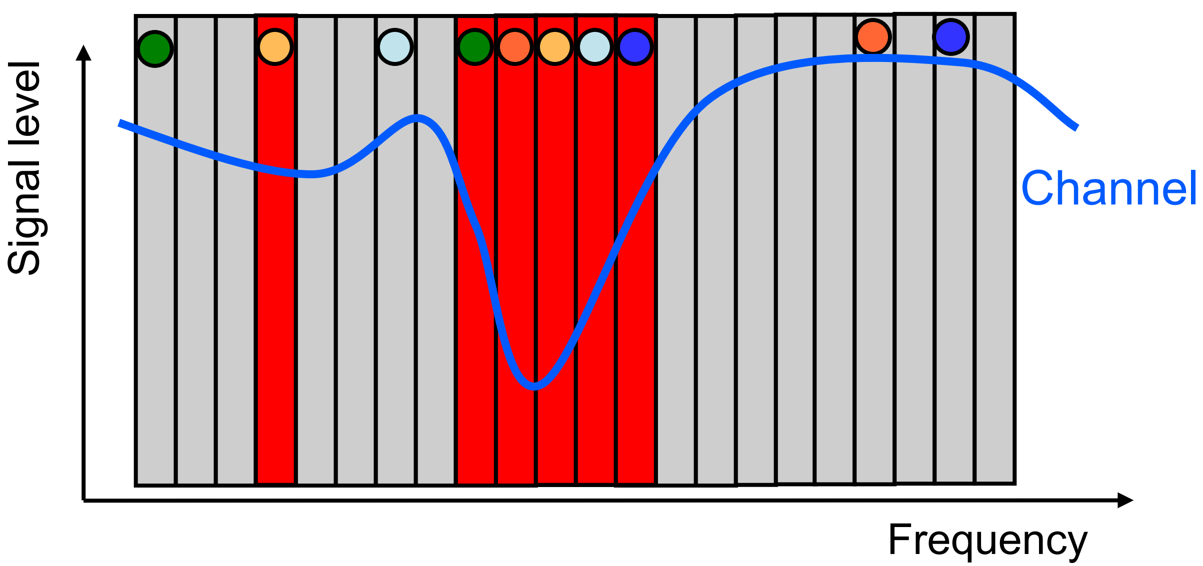
OFDM (II)

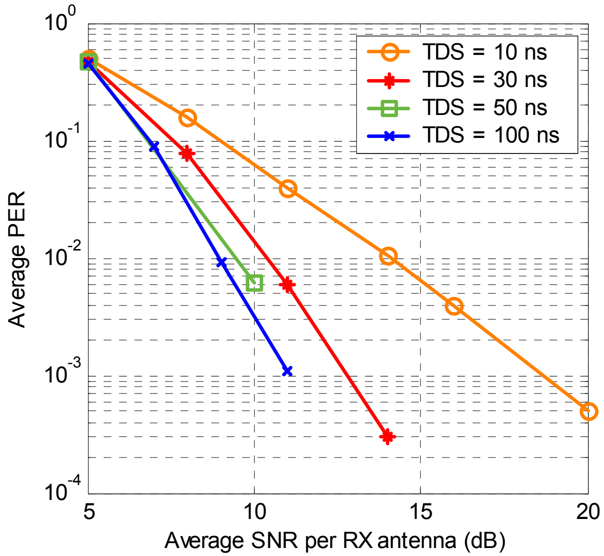
- Use of CP enables the removal of ISI at the receiver
- Increase of overhead due to CP



OFDM (III)

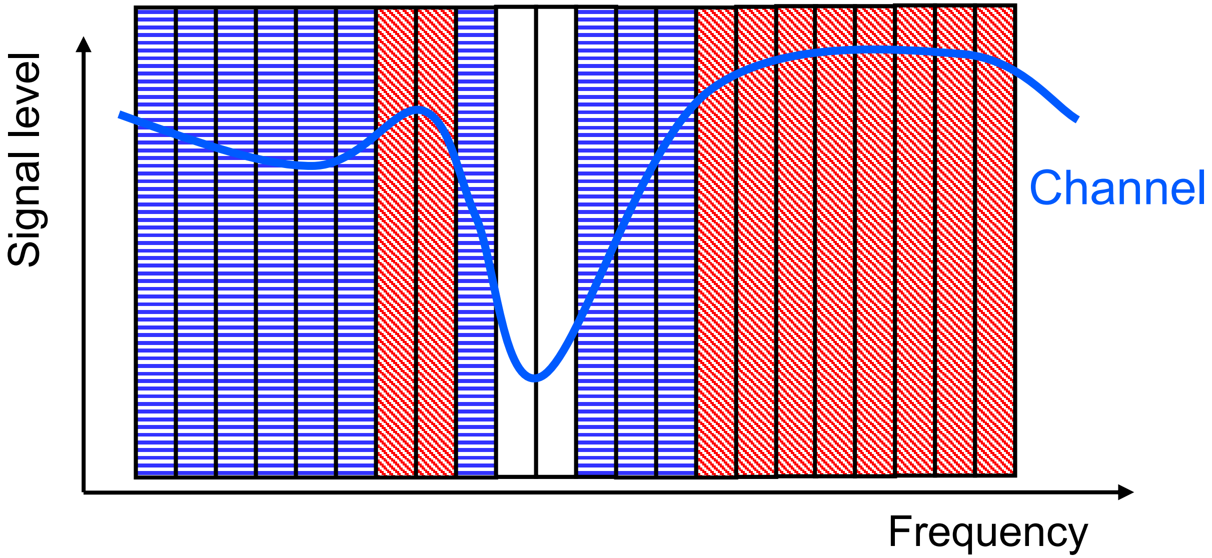
- Frequency selectivity grows with time delay spread (TDS), but flat channel per subcarrier.
- To get reliable wireless transmission *channel coding* is necessary. Coding/spreading over frequency increases diversity order.

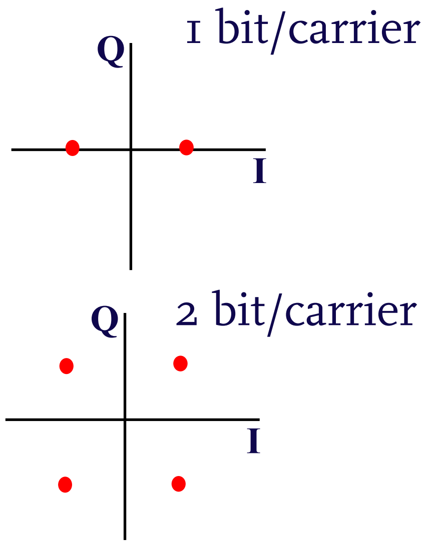




OFDM (IV)

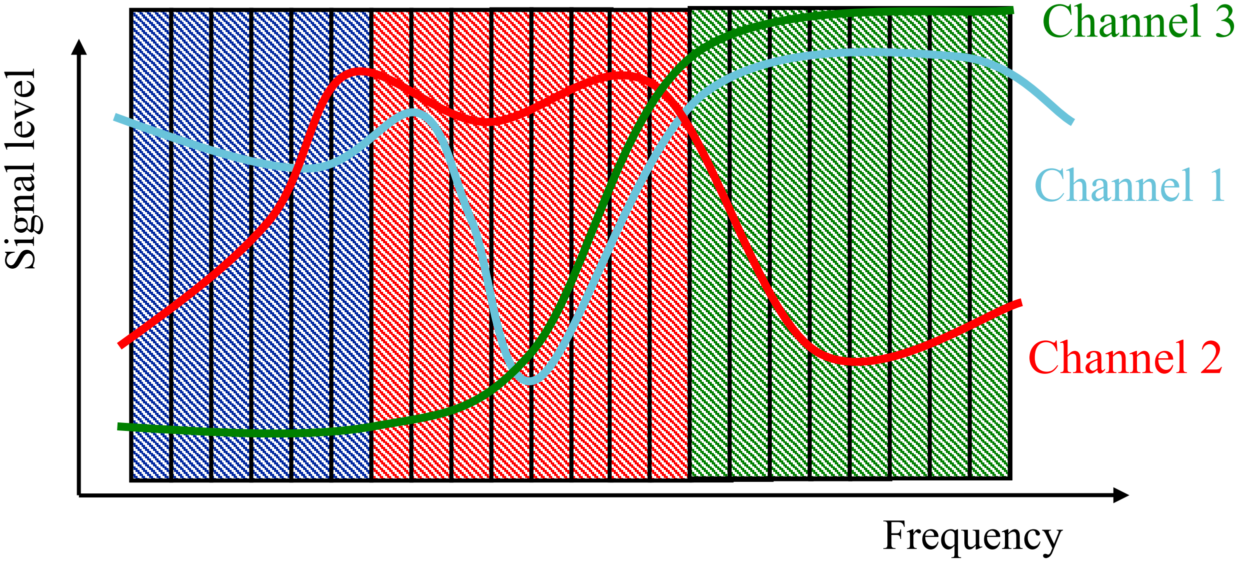
- To maximize the total throughput we can vary modulation depth on different carriers → *Adaptive modulation*
- High SNR → high number of bits,
  Low SNR → low number of bits.

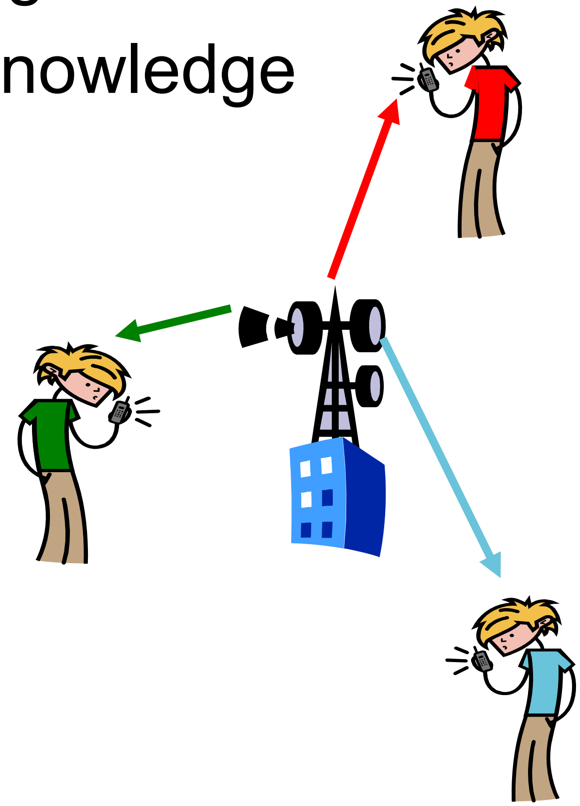




OFDM as multiple access technique → OFDMA

- Divide carriers between the different users
- Data rate differentiation between users
- Allocation can be based on channel knowledge



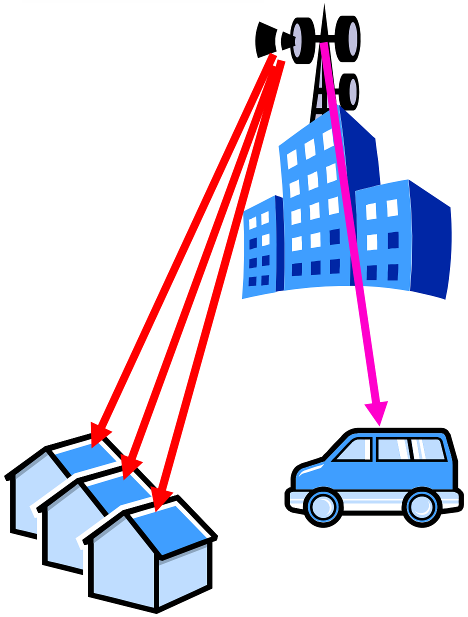


Wireless LAN

- Application: Office, home, Hotspots
- Standardized in IEEE 802.11a/g → WiFi (ETSI HiperLAN)
- Range: 10 – 100m
- 2.4 GHz and 5 GHz band
- Datarate: 6 – 54 Mbps
- Bandwidth = 20 MHz
- 64 subcarriers
- CP length = 16 samples → 800 ns

Wireless Local Loop / MAN

- Application: Last Mile (P2P), Mobile
- Standardized in IEEE 802.16 → WiMax (ETSI HiperMAN)
- Range: up to 50 km
- Flexible bandwidth: 1.5 - 20 MHz
- 16revD: 2-11 GHz, NLOS, up to 75 Mb/s
  OFDM, 256 subcarriers
  OFDMA, 2048 subcarriers
- 16revE: < 6 GHz, NLOS,
  5 MHz → 15 Mbit/s , Mobility



Digital Video Broadcasting

- Application: Replacement Traditional broadcasting, TV in car, TV everywhere
- DVB-T → Terrestrial broadcasting (European standard)
- Bandwidth: 8 MHz → 5 TV channels
- Variability: CP-length, modulation, coding rate
- Data rate: 3.7 - 31.7 Mbps
- 2k and 8k subcarriers
- 50 - 860 MHz

Ultra Wide Band

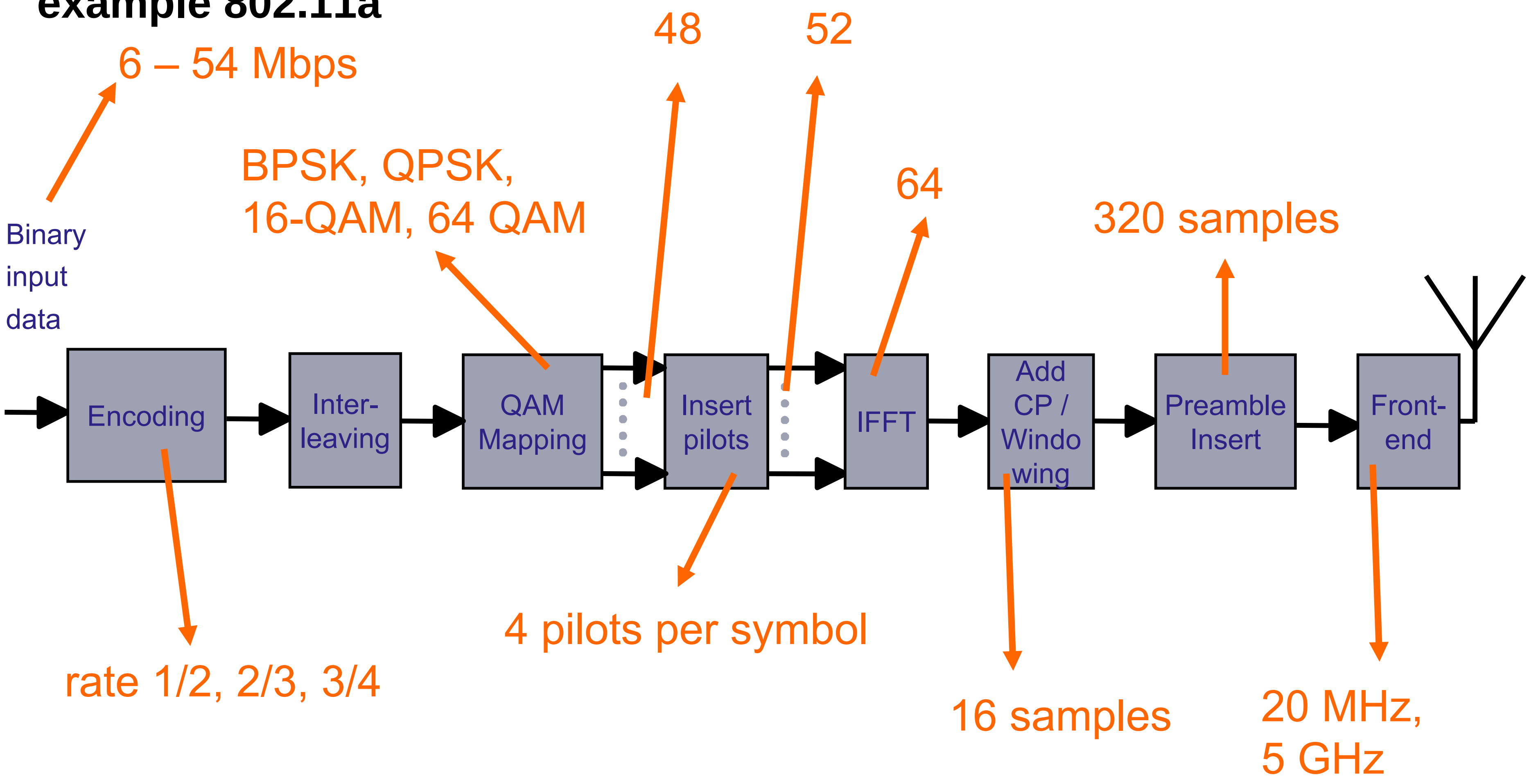
MBOA-SIG
MultiBand OFDM Alliance SIG

WiMedia ALLIANCE

- Application: Wireless PAN, cable-replacement
  → Wireless USB, Wireless FireWire, etc.
- One of last two proposals for IEEE 802.15.3a
- Range: up to 10m
- Initial: 3.1 GHz – 4.9 GHz
- Datarate: 53.3 – 480 Mbps
- Bandwidth = 528 MHz, 128 subcarriers
- QPSK modulation, variable coding rate

OFDM Transmitter (baseband)

example 802.11a



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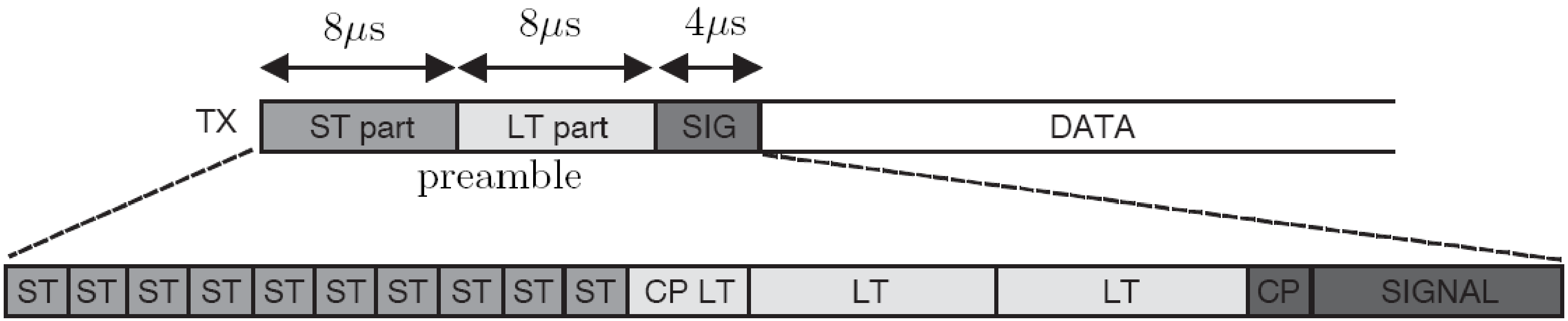
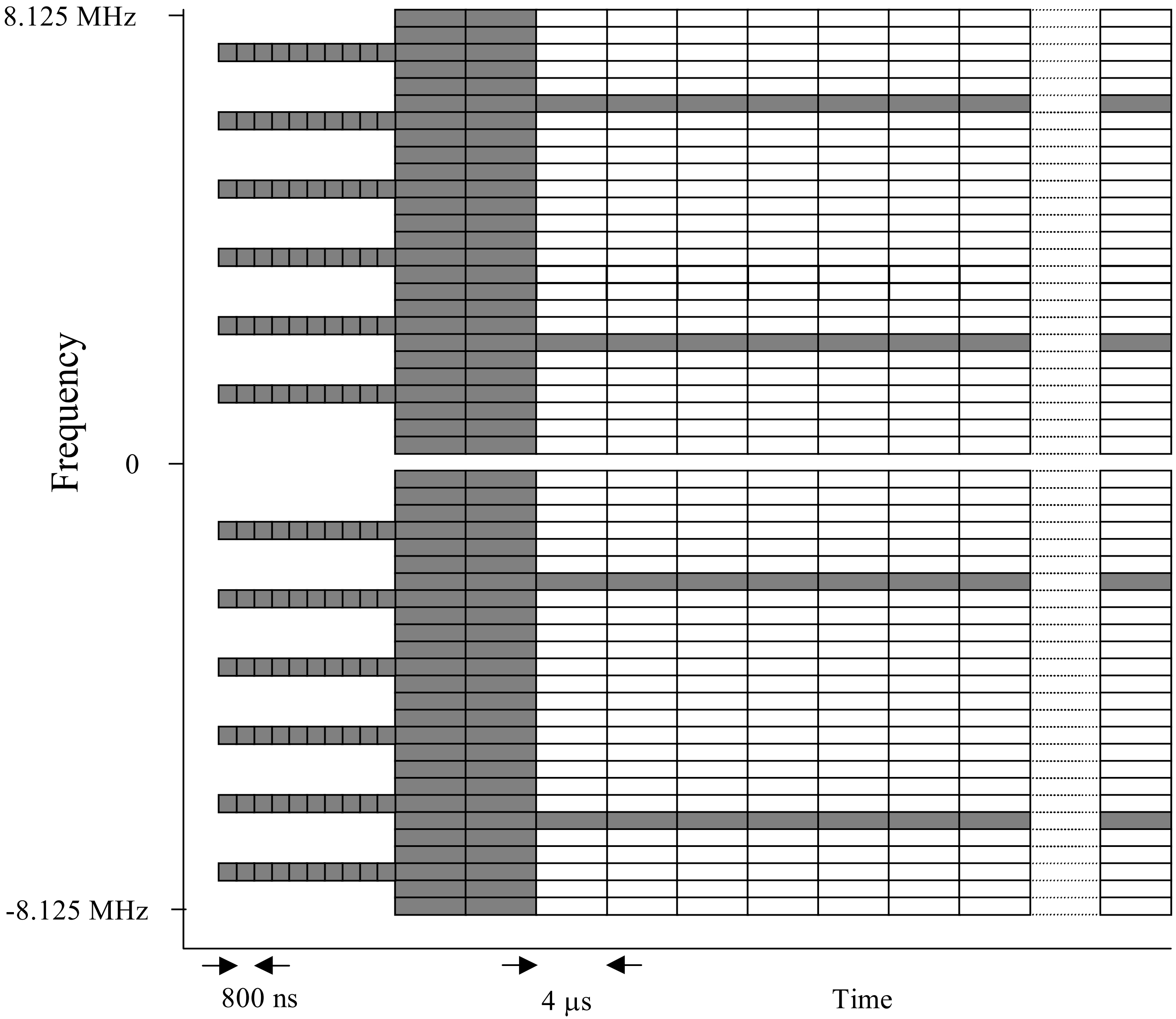
Preamble and pilot structure 802.11a (I)

Preamble



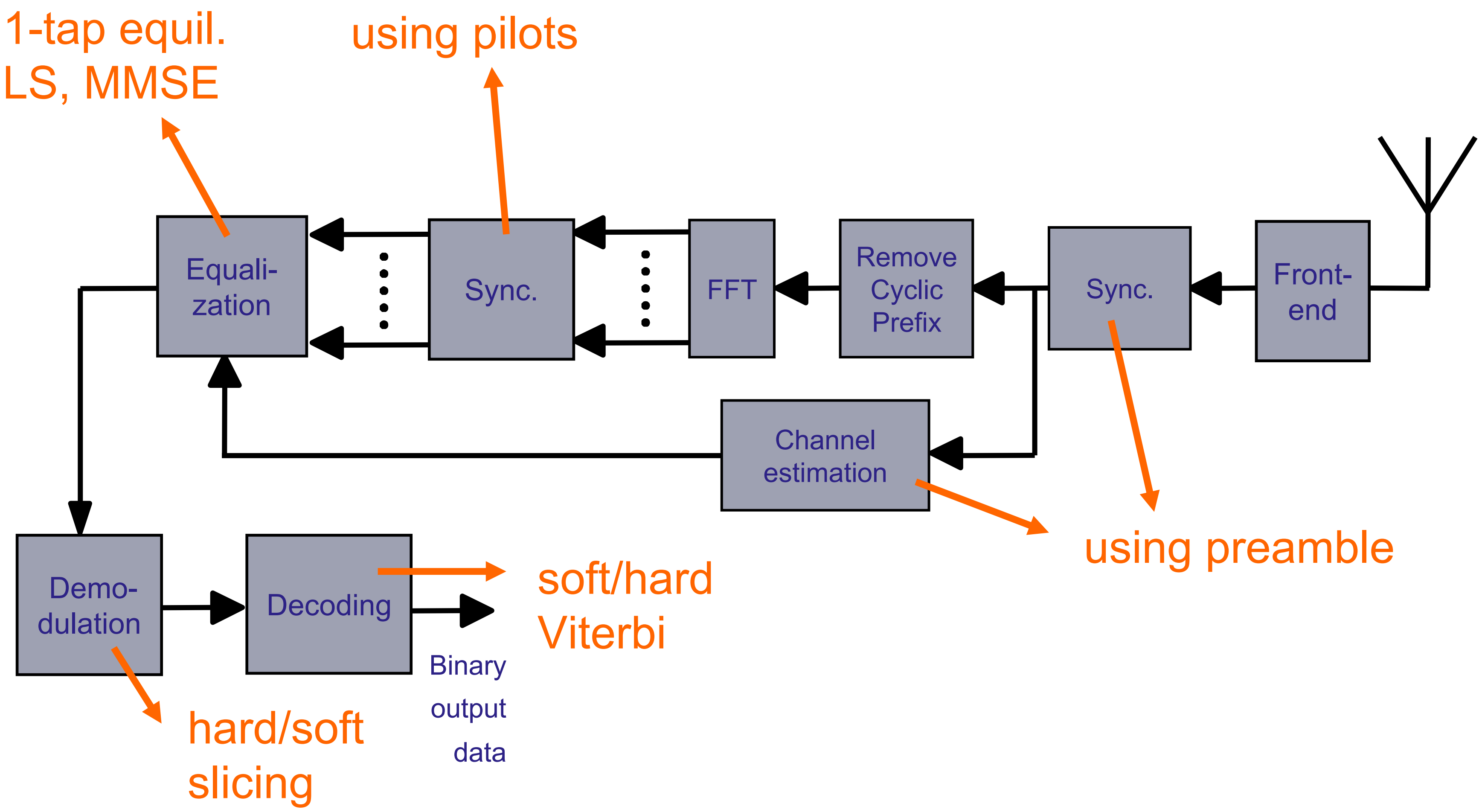
PHILIPS 3TU.

Preamble and pilot structure 802.11a (II)



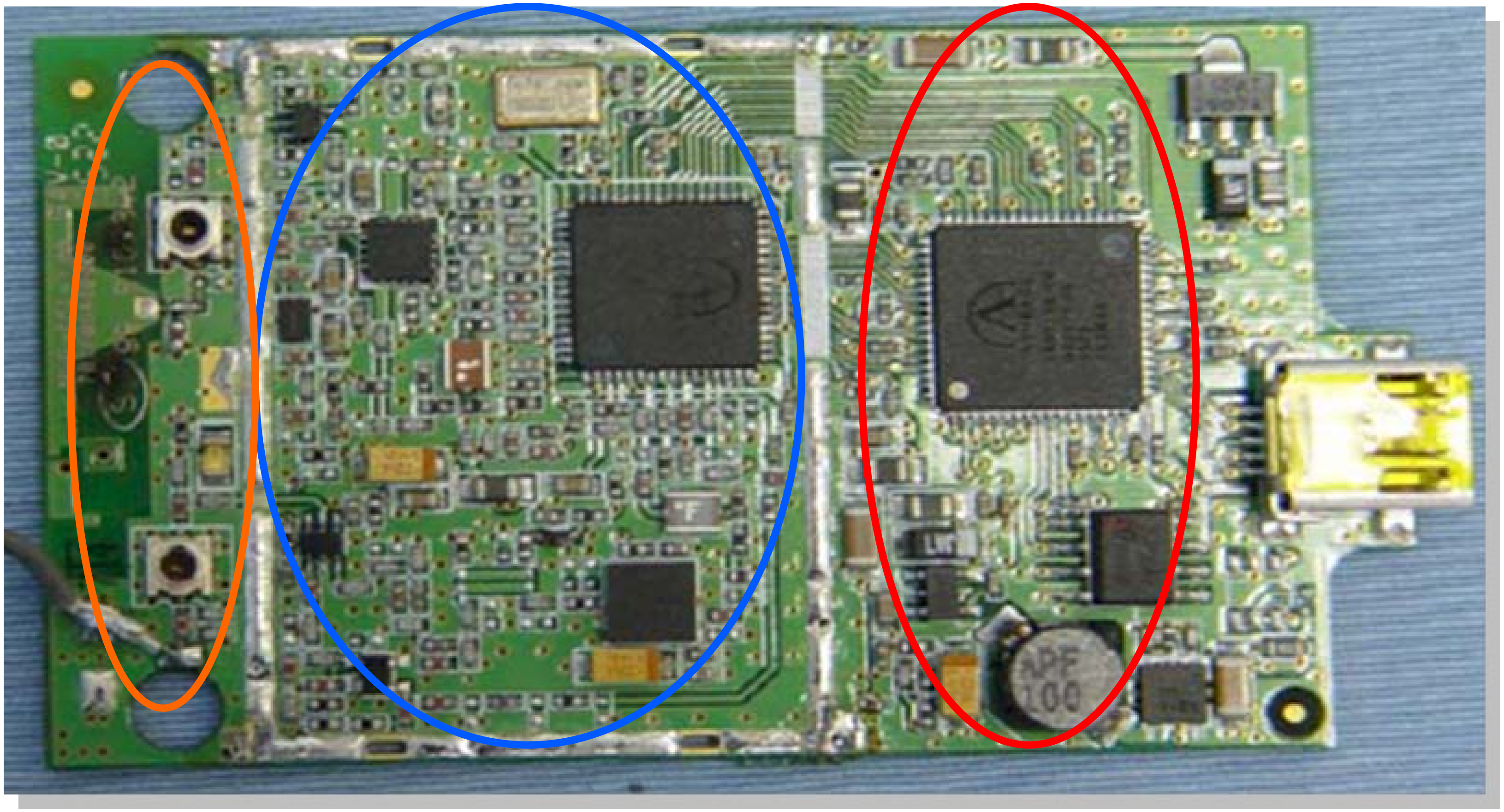
OFDM Receiver (baseband)

example 802.11a



Implementation – IEEE802.11a/g

Atheros



Antenna connections

RF front-ends

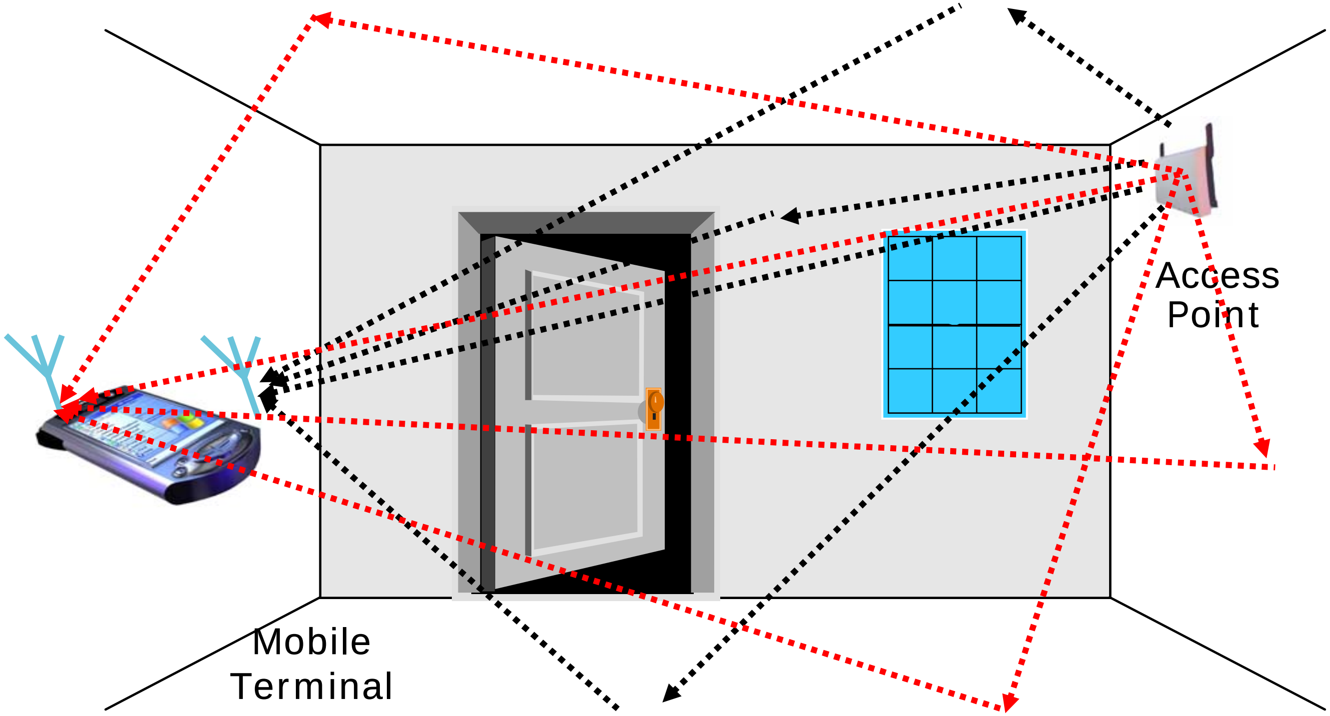
Baseband + MAC

Multiple-Input Multiple-Output

“The use of multiple antennas at both transmitter and receiver of a wireless communications system.”

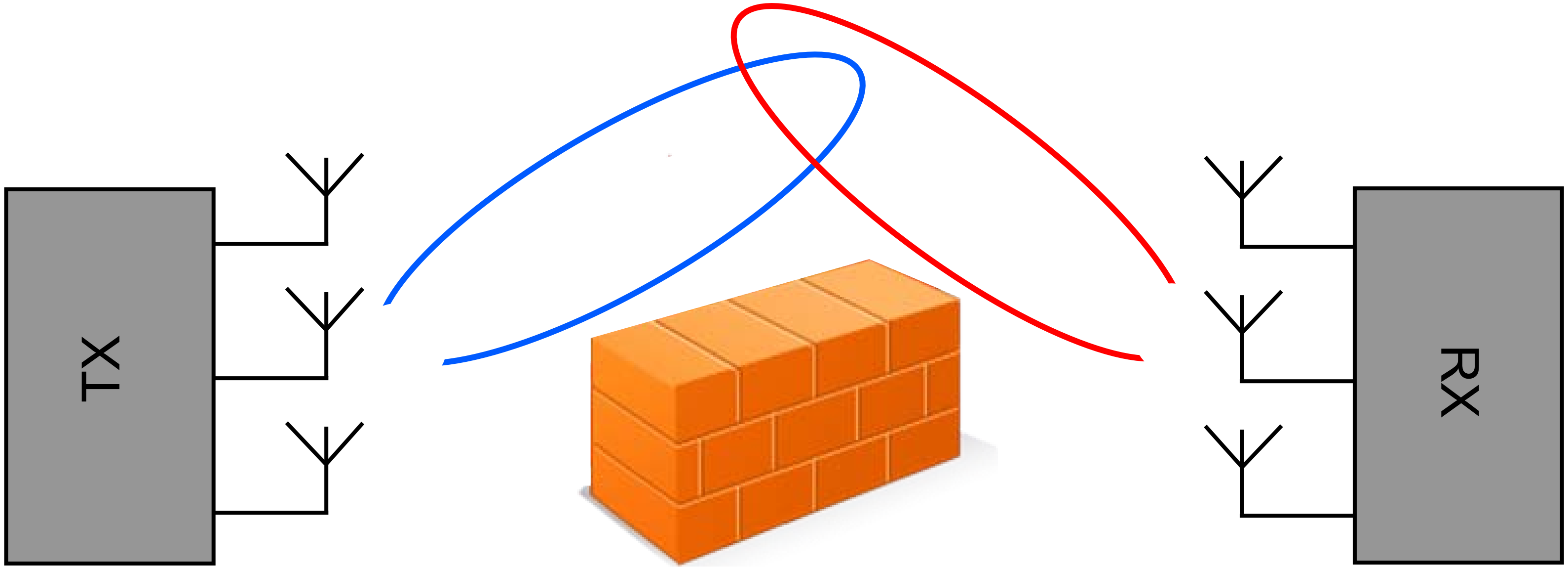
- Provides additional link throughput and/or range, without increase in bandwidth and transmit power.
- Three main flavours:
  - **Beamforming/Precoding**
  - **Diversity coding**
  - **Space division multiplexing**

MIMO Channel



MIMO Flavours: 1) Beam forming/precoding

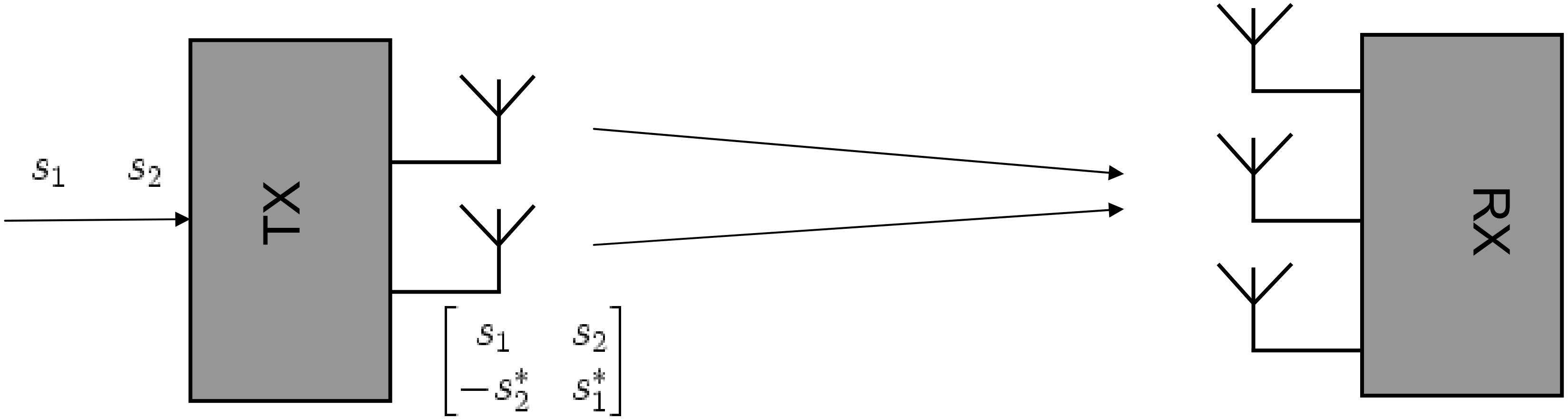
- Increase the reliability/received power of wireless link
- Transmission using the eigenmodes of the channel
- Requires channel state information at the TX
- Different transformations for the different antennas / subcarriers



MIMO Flavours: 2) Diversity coding

diversity scheme: “a method for improving the reliability of a message signal by utilizing two or more communication channels with different characteristics, in order to combat fading and interference”

- No channel state information available at the TX
- Coding over space-time-frequency (block/trellis)
- Orthogonal/non-orthogonal codes
- Most well-known: Alamouti, Tarokh



$$\begin{bmatrix} s_1 & s_2 \\ -s_2^* & s_1^* \end{bmatrix}$$

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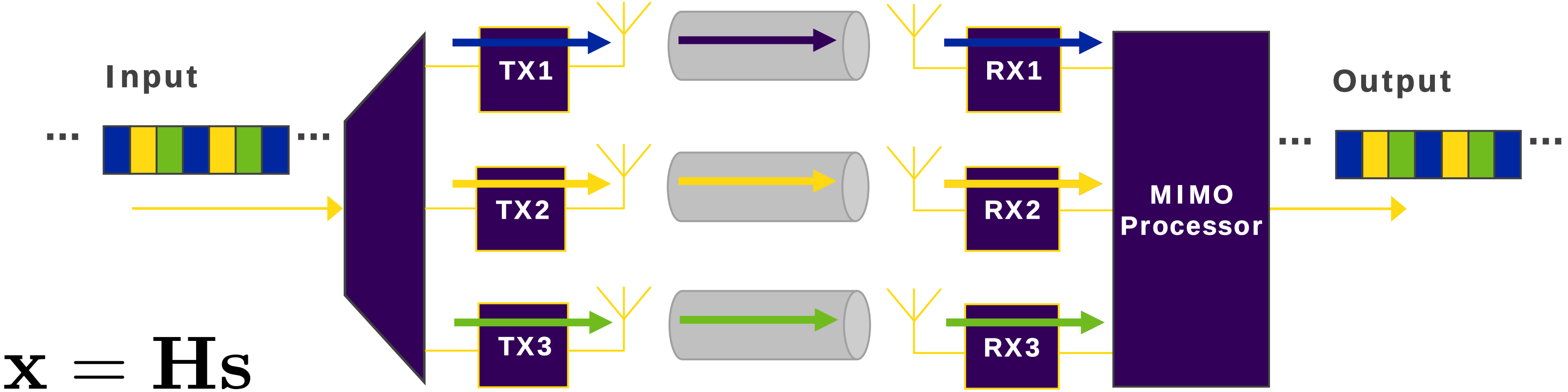
MIMO Flavours: 3) SDM

- High speed data stream demultiplexed into lower rate parallel streams
- Complexity in TX: to estimate the TX data
- No knowledge of wireless channel at TX required



$$\mathbf{x} = \mathbf{H}\mathbf{s}$$

$$\mathbf{s} = [s_1 s_2 \ldots s_{N_\mathrm{t}}]^T \quad \mathbf{H} = \begin{bmatrix} H_{11} & \cdots & H_{1N_\mathrm{r}} \\ \vdots & \ddots & \vdots \\ H_{N_\mathrm{t}1} & \cdots & H_{N_\mathrm{t}N_\mathrm{r}} \end{bmatrix} \quad \mathbf{x} = [x_1 x_2 \ldots x_{N_\mathrm{t}}]^T$$

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MIMO Flavours: 3) SDM (cont'd)

$$\mathbf{x} = \mathbf{H}\mathbf{s}$$

- Different kind of RX processing:
  - Linear:
    - Zero-Forcing: $\tilde{\mathbf{s}} = \mathbf{H}^{\dagger}\mathbf{x} = (\mathbf{H}^{H}\mathbf{H})^{-1}\mathbf{H}^{H}\mathbf{x}$
    - MMSE: $\tilde{\mathbf{s}} = (\frac{1}{\mathrm{SNR}}\mathbf{I} + \mathbf{H}^{H}\mathbf{H})^{-1}\mathbf{H}^{H}\mathbf{x}$
  - Iterative:
    - BLAST: 1) Detect s_n with the highest SNR
      2) Substract contributes of s_n of $\mathbf{x}$
      3) return to 1) until all streams are detected
  - Full search:
    - MLD: $\tilde{\mathbf{s}} = \arg\min_{\mathbf{s}} ||\mathbf{x} - \mathbf{H}\mathbf{s}||$

MIMO Flavours: Combinations

- MIMO transmission schemes is an very active research field: #1 topic on IEEE communications conferences for last few years.
- Tradeoffs between diversity and data rate is main topic → combination between 2) and 3)
  - 2 TX-based schemes (spatial spreading) with 3 TX branches
- Schemes exploiting partial feedback of CSI → combination between 1) and 3)
  - Precoding / pre-equilization at the transmitter
  - MU MIMO
- Holy Grail: Differential MIMO

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Physical Interpretation (I)

- Assuming narrowband communication and using the matrix notation, the system equation can be written as (omitting noise)

$$\mathbf{x} = \mathbf{H}\mathbf{s}$$

- If **H** is square and invertible, the most simple solution (in math) is called zero-forcing and given by

$$\tilde{\mathbf{s}} = \mathbf{H}^{\dagger}\mathbf{x} = \mathbf{H}^{\dagger}\mathbf{H}\mathbf{s} = \mathbf{s}$$

- The rows of $\mathbf{H}^{\dagger}$are the so called weight vectors, denoted by $\mathbf{w}$. For a 2 x 2 system this results in

$$\tilde{\mathbf{s}}_1 = \mathbf{w}^1\mathbf{x}$$

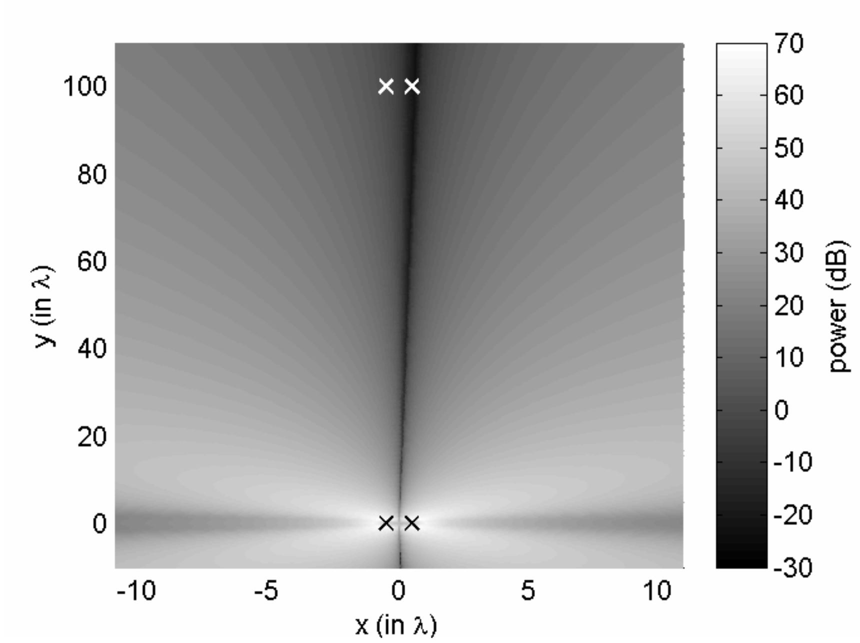
$$\tilde{\mathbf{s}}_2 = \mathbf{w}^2\mathbf{x}$$

Slides: Allert van Zelst, Qualcomm

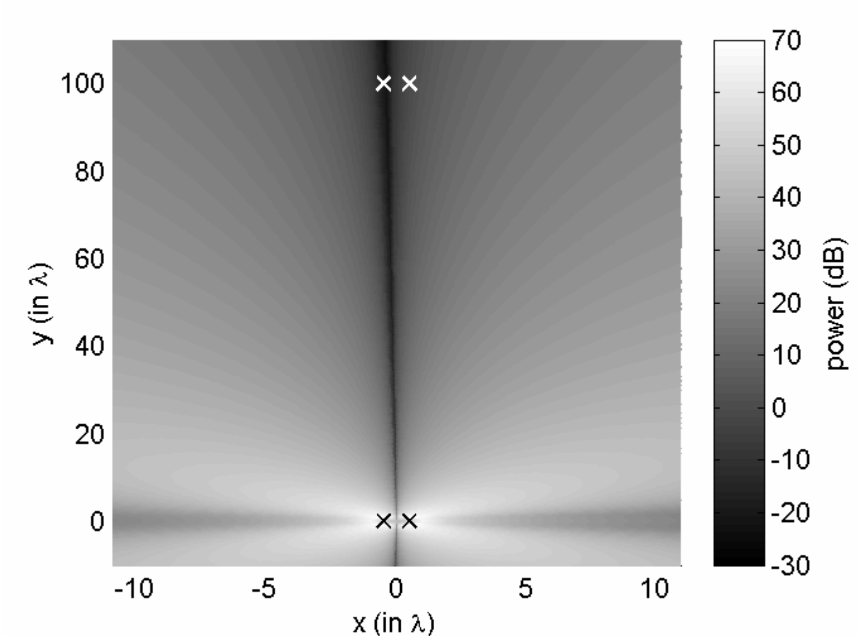
Physical Interpretation (II)

- Applying the correct weights in line-of-sight with free space path loss, we get:

Applying Weight Vector $\mathbf{w}^1$



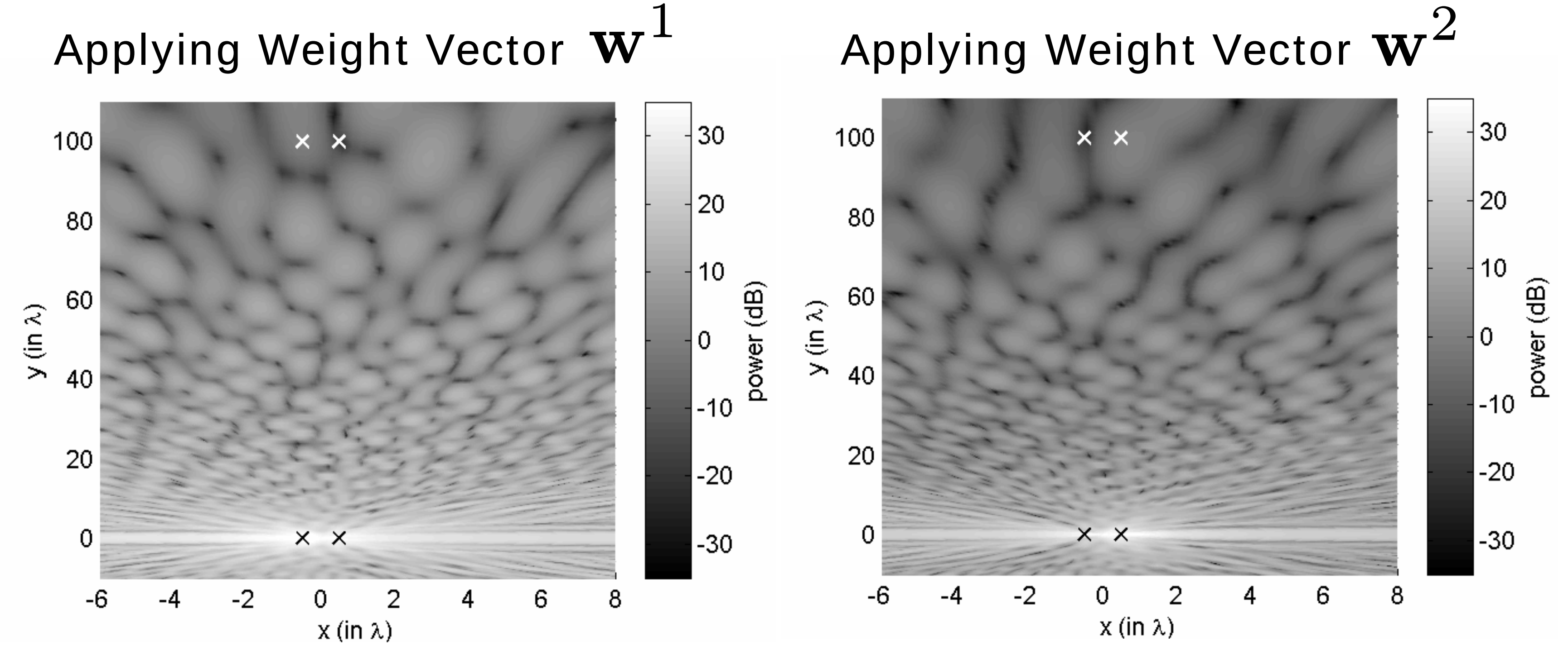
Applying Weight Vector $\mathbf{w}^2$



Slides: Allert van Zelst, Qualcomm

Physical Interpretation (III)

- In an environment with reflection planes at x = -6λ and x = 8λ, taking up to two bounces into account, we get:

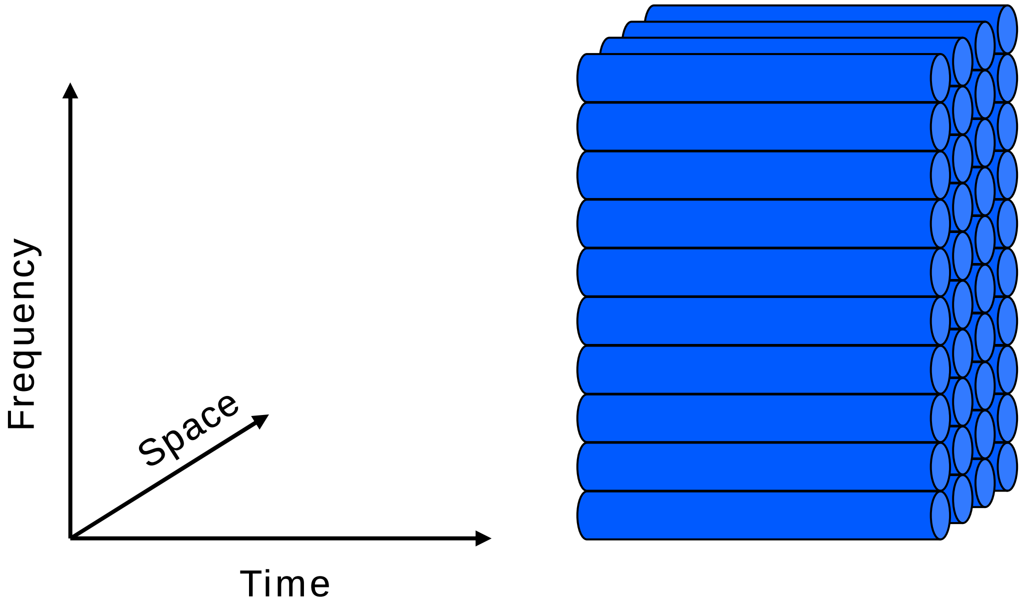


Slides: Allert van Zelst, Qualcomm

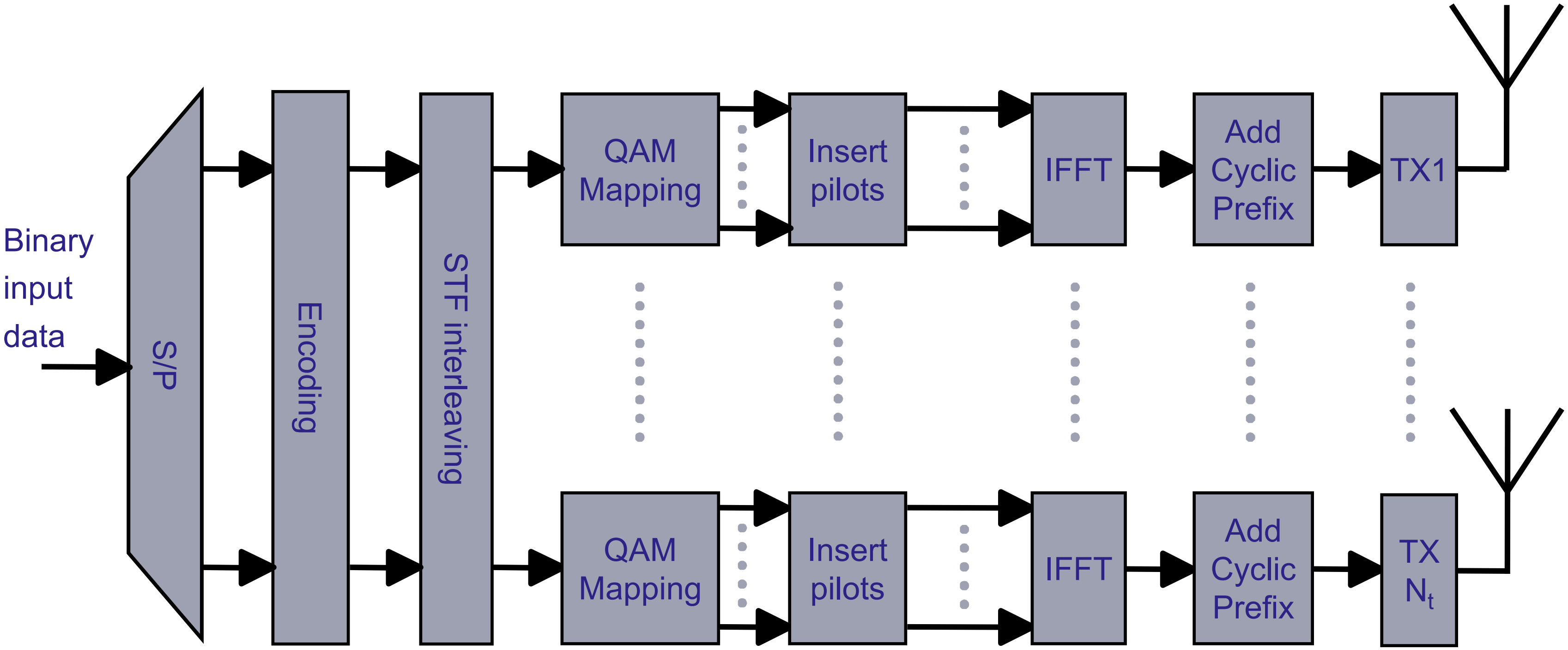
MIMO Applied to OFDM

- MIMO is a narrow band technique
- Combining MIMO and OFDM results in a MIMO transmission and detection per subcarrier

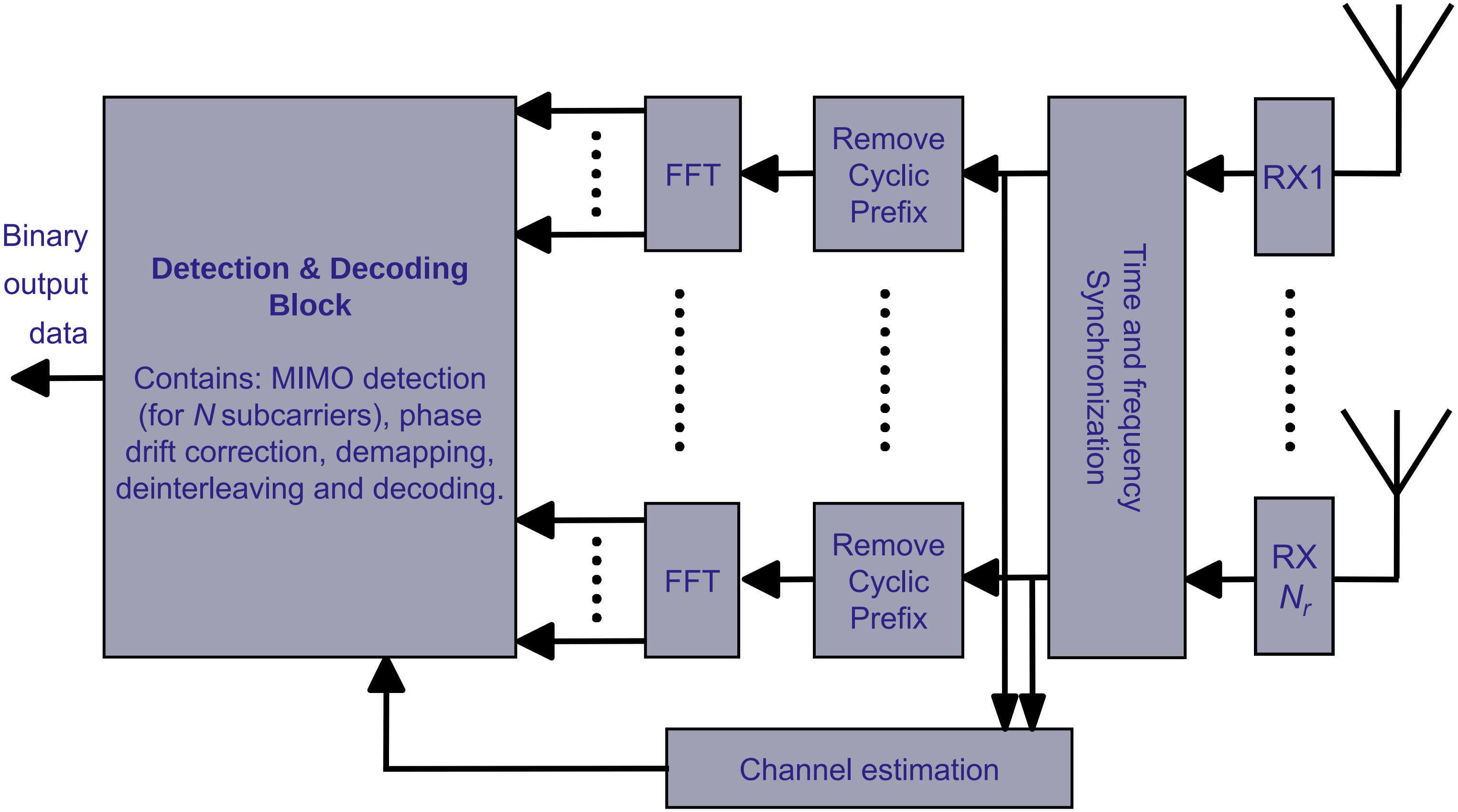
MIMO OFDM



MIMO OFDM Transmitter



MIMO OFDM Receiver


Binary output data

Detection & Decoding Block

Contains: MIMO detection (for *N* subcarriers), phase drift correction, demapping, deinterleaving and decoding.

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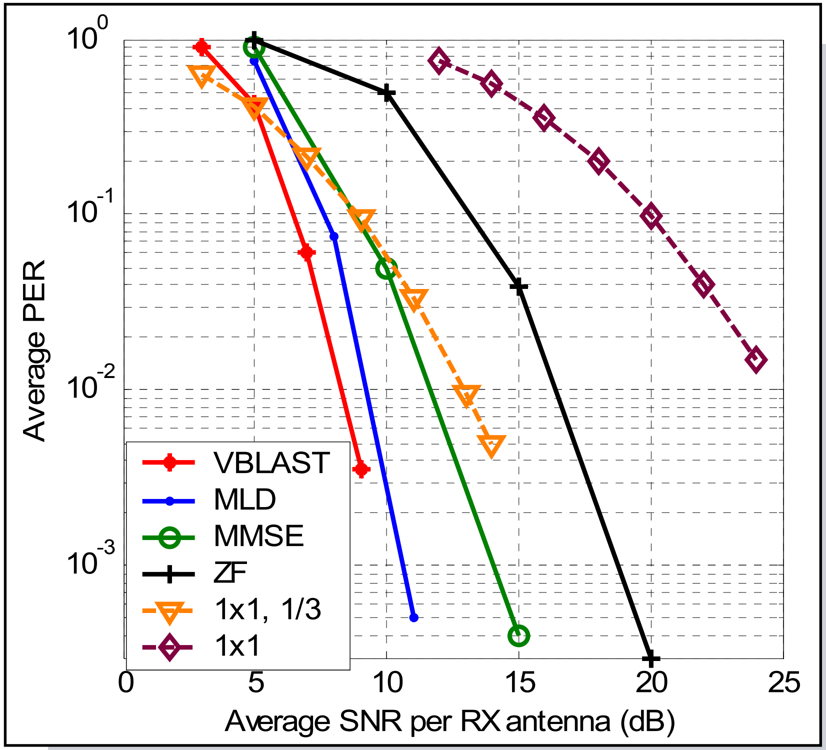
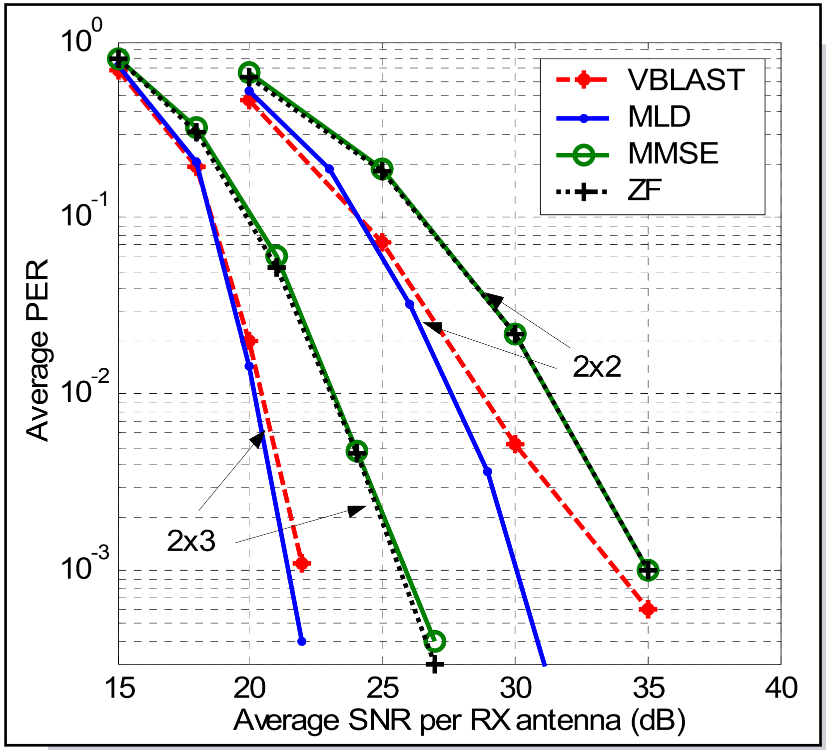
Performance of MIMO OFDM (I)



3x3 configuration for QPSK, rate ½ conv. coding (36 Mbps), 1000 byte packets, rms delay spread = 50 ns

- Results for 3x3 system, different detection methods. Corresponding curves for SISO systems with equal branch rate (1x1, 1/3) and equal data rate (1x1).
- Perfect sync. and channel knowledge at RX
- VBLAST is less complex than MLD but performance is comparable

Performance of MIMO OFDM (II)



- 2x3 linear detector schemes, ZF/MMSE, outperform 2x2 computational complex detectors
- When extra receiver antenna is affordable, linear detection can be applied to achieve good performance.

2x2 and 2x3 configuration for 64QAM, rate 3/4 conv. coding (108 Mbps), 64 byte packets, rms delay spread = 30 ns.

MIMO OFDM in Wireless LANs

- In the IEEE 802.11a/g wireless LAN standards, we have 48 subchannels available for data communication
  - Each subchannel has a maximum air-throughput of 1.5 Mbits/s
- For MIMO OFDM with three transmit antennas, this would lead to a maximum throughput of 3*48*1,5 Mbits/s = 216 Mbits/s
- To have a more robustness link, coding is used in IEEE 802.11a/g. The maximum coding rate is ¾
- This leads to a net throughput of ¾*216 = 162 Mbits/s

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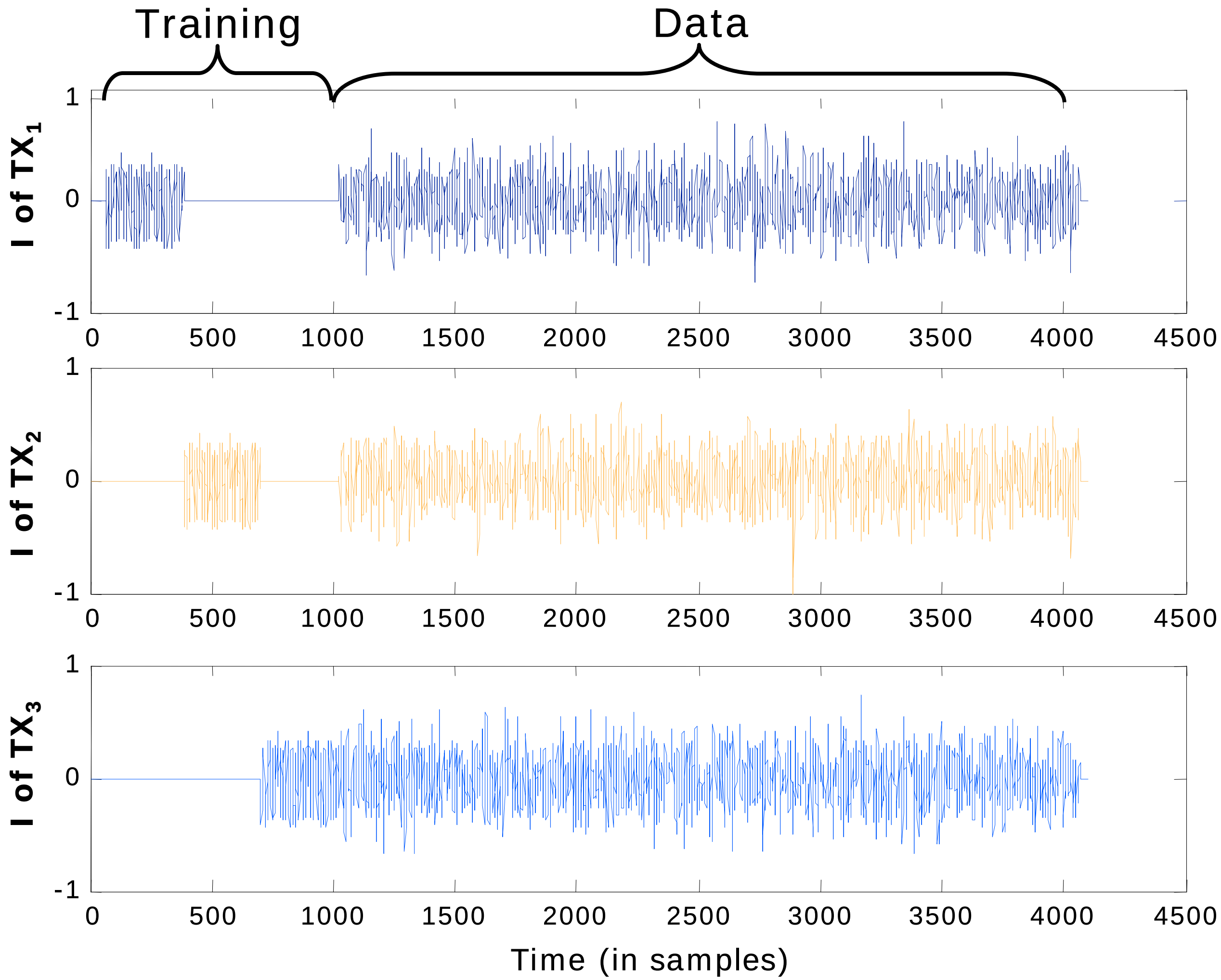
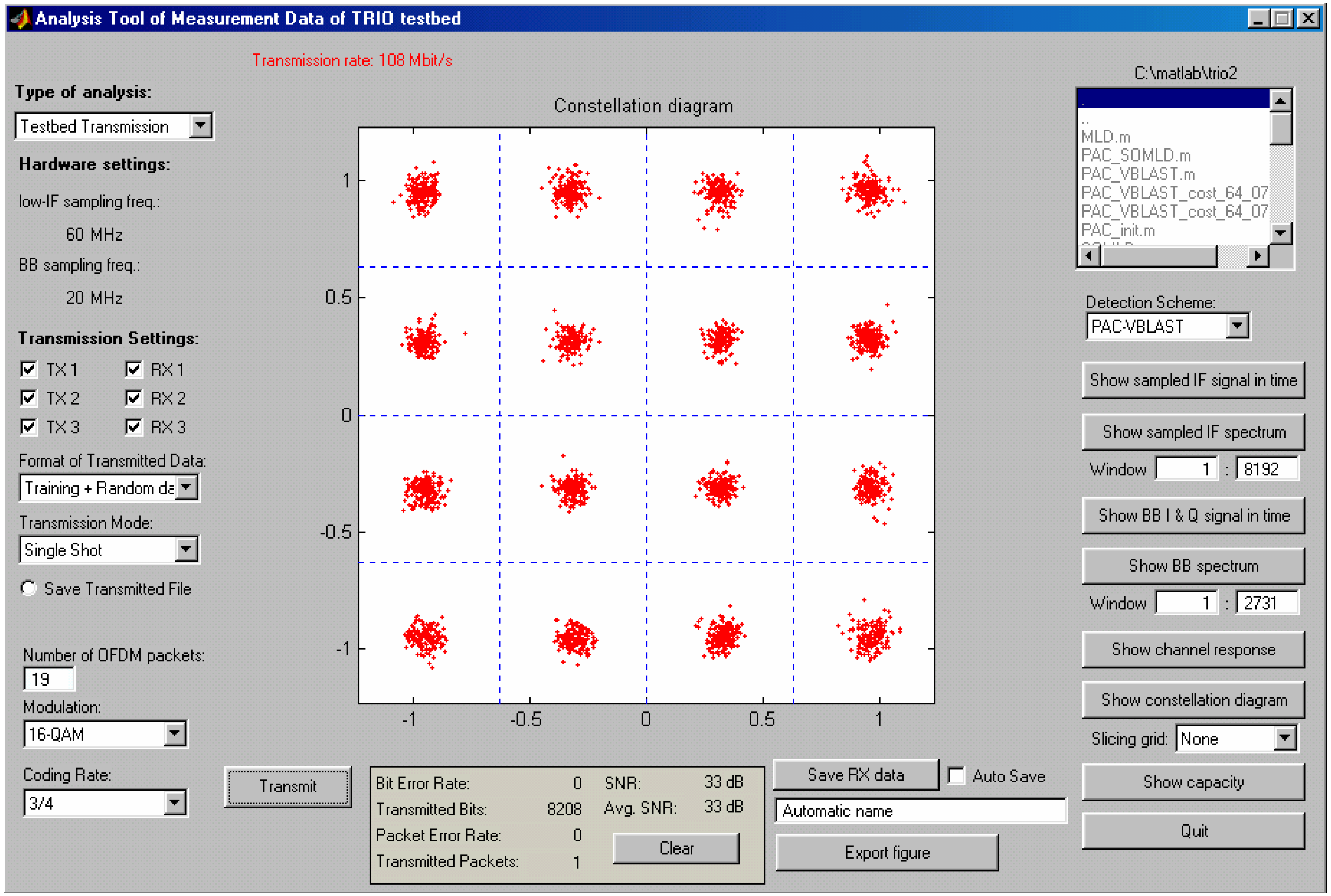
TRIO (Triple Input Output)

- In 2002 we have demonstrated 162 Mbits/s with Agere’s TRIO test system
- It transmits on a standard IEEE 802.11a channel in the 5 GHz ISM band with about 50 mW per transmit antenna
- It is based on in-house developed RF, IF and baseband boards
- For RF, a 5 GHz GaAs chip is used
- Non-real time: off-line MIMO OFDM processing

The Transmit Signals



108 Mbits/s Transmission



Analysis Tool of Measurement Data of TRIO testbed

Transmission rate: 108 Mbit/s

Type of analysis: Testbed Transmission

Hardware settings:

low-IF sampling freq.: 60 MHz

BB sampling freq.: 20 MHz

Transmission Settings:

☑ TX 1 ☑ RX 1
☑ TX 2 ☑ RX 2
☑ TX 3 ☑ RX 3

Format of Transmitted Data: Training + Random da

Transmission Mode: Single Shot

○ Save Transmitted File

Number of OFDM packets: 19

Modulation: 16-QAM

Coding Rate: 3/4

Bit Error Rate:	0	SNR:	33 dB
Transmitted Bits:	8208	Avg. SNR:	33 dB
Packet Error Rate:	0		
Transmitted Packets:	1		

☐ Auto Save

Automatic name

C:\matlab\trio2

- ..
- MLD.m
- PAC_SOMLD.m
- PAC_VBLAST.m
- PAC_VBLAST_cost_64_07
- PAC_VBLAST_cost_64_07
- PAC_init.m

Detection Scheme: PAC-VBLAST

Window 1 : 8192

Window 1 : 2731

Slicing grid: None

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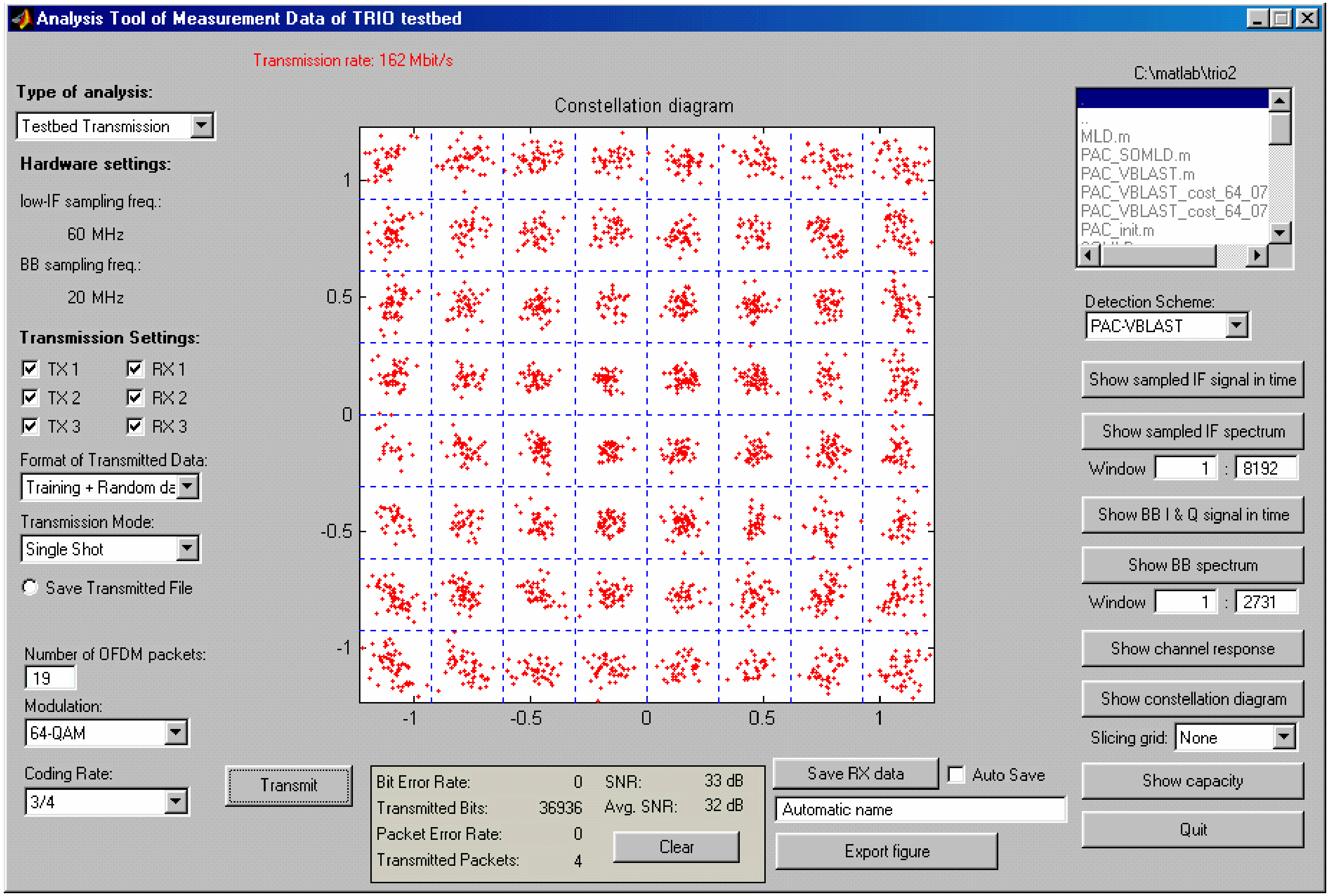
162 Mbits/s Transmission



Analysis Tool of Measurement Data of TRIO testbed

Transmission rate: 162 Mbit/s

Type of analysis: Testbed Transmission

Hardware settings:

low-IF sampling freq.: 60 MHz

BB sampling freq.: 20 MHz

Transmission Settings:

☑ TX 1 ☑ RX 1
☑ TX 2 ☑ RX 2
☑ TX 3 ☑ RX 3

Format of Transmitted Data: Training + Random da

Transmission Mode: Single Shot

○ Save Transmitted File

Number of OFDM packets: 19

Modulation: 64-QAM

Coding Rate: 3/4

Bit Error Rate:	0	SNR:	33 dB
Transmitted Bits:	36936	Avg. SNR:	32 dB
Packet Error Rate:	0		
Transmitted Packets:	4		

☐ Auto Save

Automatic name

C:\matlab\trio2

..
MLD.m
PAC_SOMLD.m
PAC_VBLAST.m
PAC_VBLAST_cost_64_07
PAC_VBLAST_cost_64_07
PAC_init.m

Detection Scheme: PAC-VBLAST

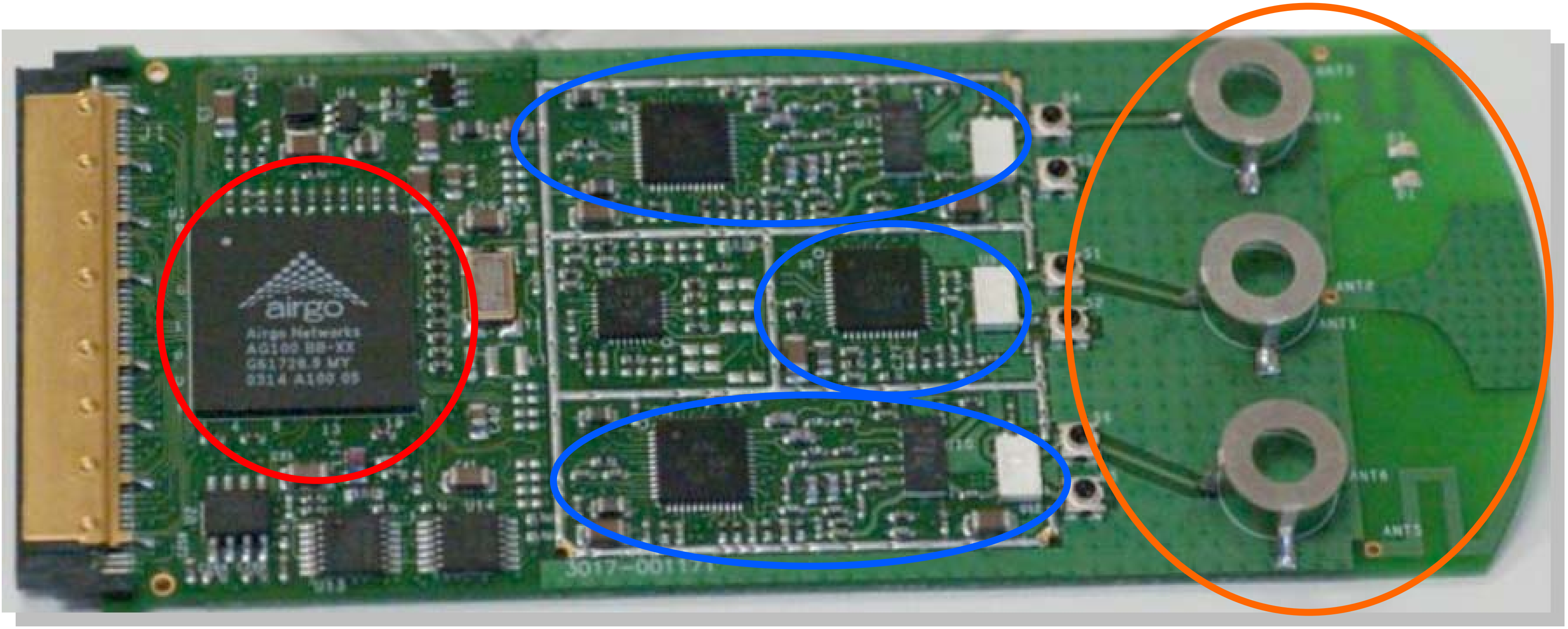
Window 1 : 8192

Window 1 : 2731

Slicing grid: None

Implementation – MIMO 802.11a/g

Airgo Networks (now Qualcomm)

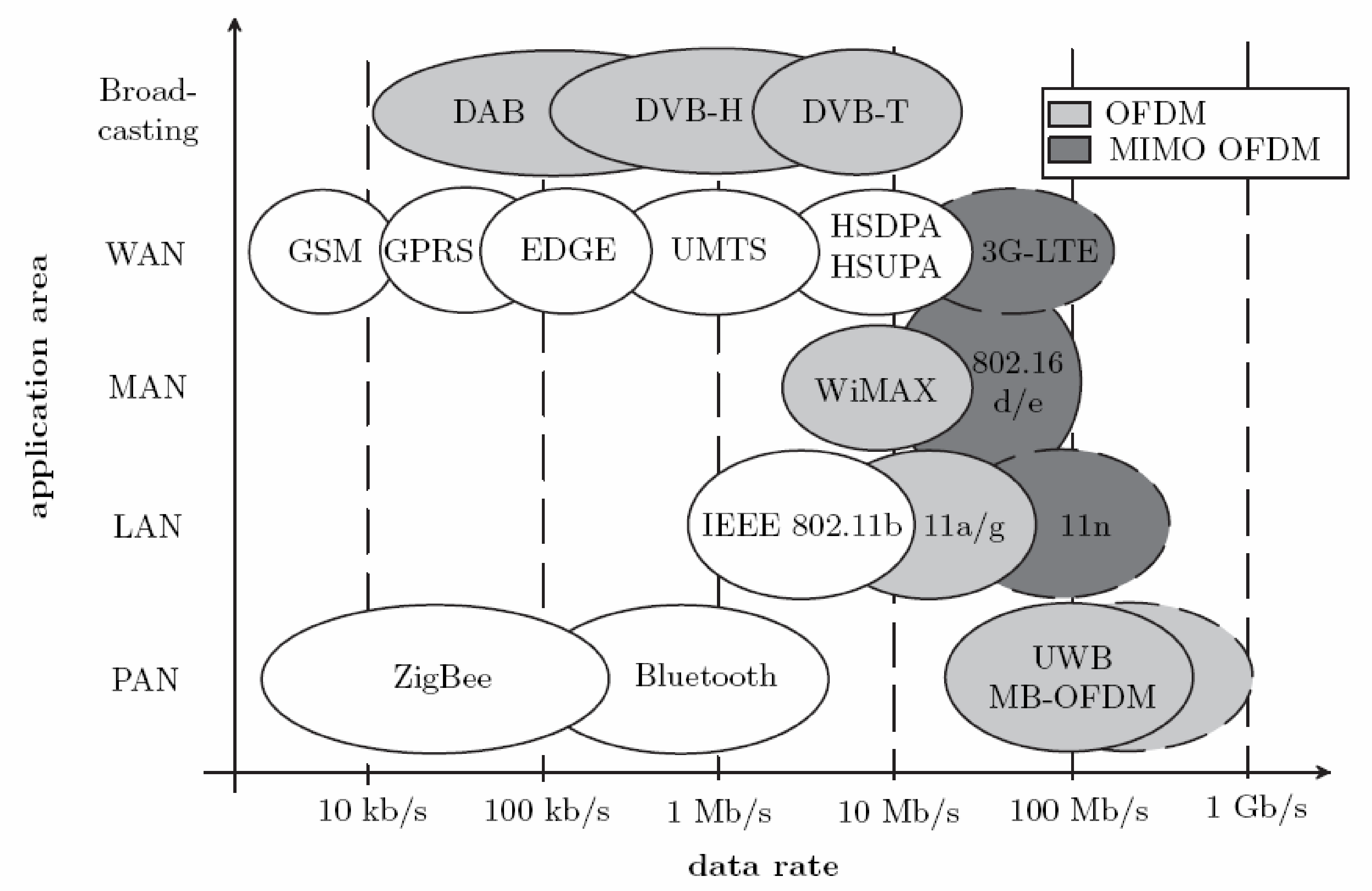


RF front-ends

Antennas

Baseband + MAC

Wireless standards + MIMO/OFDM



Summary – Part I

- Combination of **MIMO** and **OFDM** for high speed wireless systems
  - ☺ Works well in multipath environments
  - ☺ High spectral efficiency
  - ☺ Allows block processing and simple equalization (FFT)

  - ☹ Requires multiple RF front-ends
  - ☹ Performance is jeopardized by analogue radio front-end impairments (Part II&III)

MIMO/OFDM – Additional reading

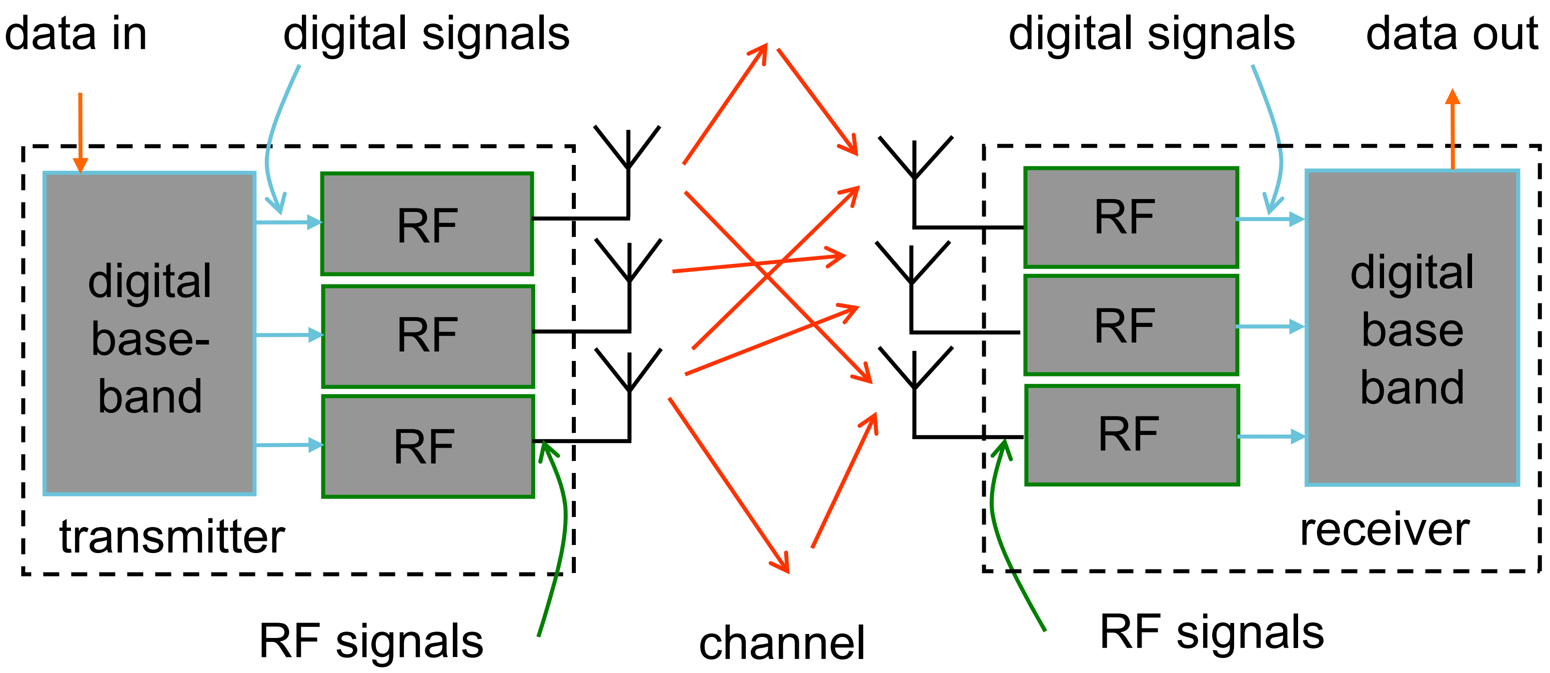
- D. Gesbert, et al., “From theory to practice: an overview of MIMO space-time coded wireless systems”, IEEE JSAC, April 2003.
- S. Alamouti, “A simple transmit diversity technique for wireless communications,” IEEE JSAC, 1998.
- G. Foschini, “Layered space-time architecture for wireless communication in a fading environment when using multi-element antennas,” Bell Labs Technical Journal, 1996.
- Stüber, et al., “Broadband MIMO-OFDM Wireless Communications,”, Proc. IEEE, Feb. 2004.
- Van Zelst and Schenk, “Implementation of a MIMO OFDM-based wireless LAN system,” IEEE Trans. Sign. Proc., Feb. 2004.
- R. van Nee and R. Prasad, “OFDM for Wireless Multimedia Communications”, Book, Artech House.

Part II: Digital comp. of RF impairments (I)

- Why digital compensation?
- Carrier frequency offset
  - What is the influence?
  - How to treat it?
- Phase Noise
  - What is the influence?
  - How to treat it?

MIMO OFDM system

$\mathbf{s}_k \sim N_\mathrm{t} \times 1$ $\mathbf{H}_k \sim N_\mathrm{r} \times N_\mathrm{t}$ $\mathbf{x}_k \sim N_\mathrm{r} \times 1$



data in

digital signals

digital signals

data out

digital base-band

RF

RF

RF

transmitter

RF

RF

RF

digital base band

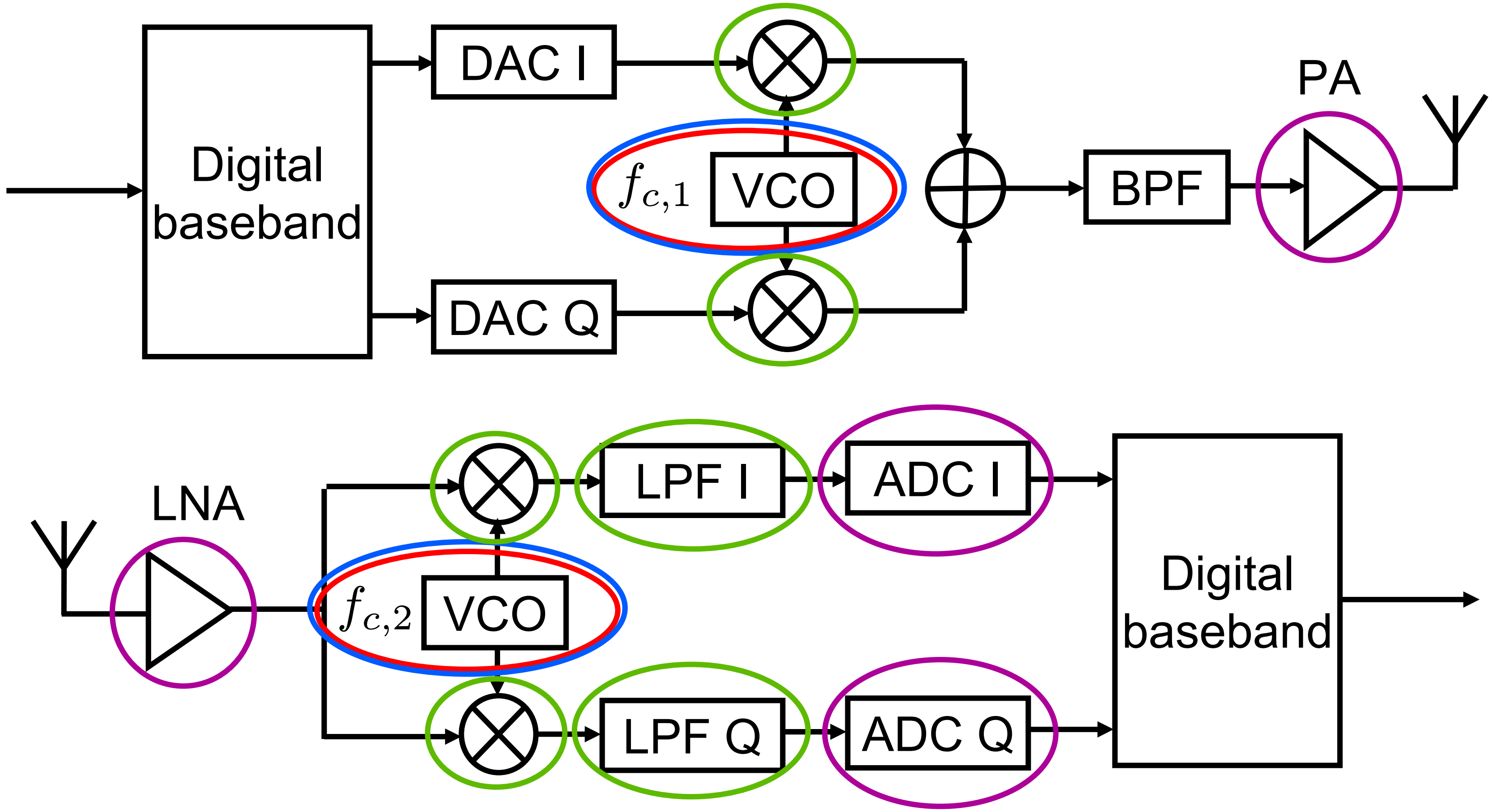
receiver

RF signals

channel

RF signals

Typical transceiver structure - impairments



carrier frequency offset

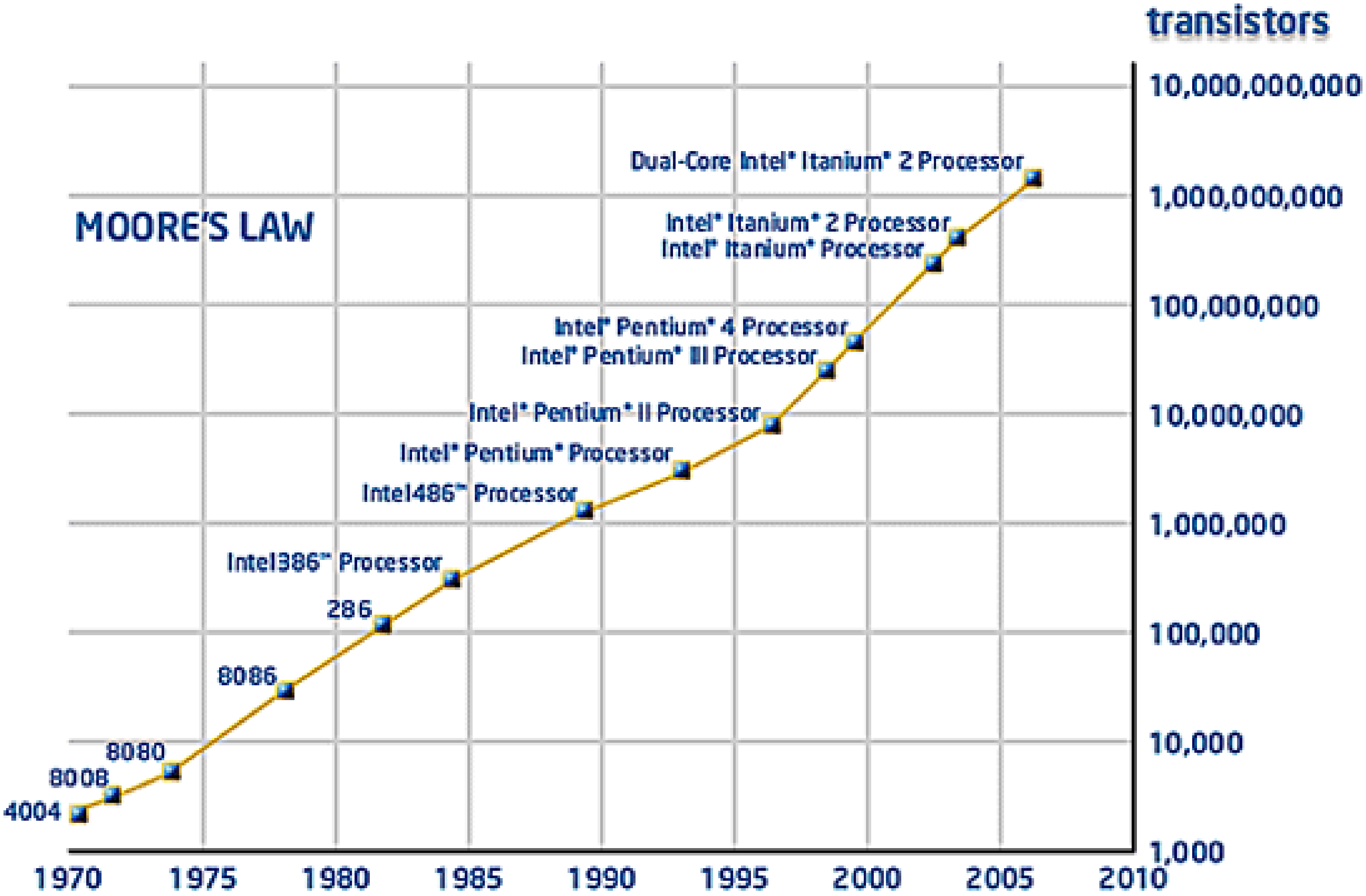
IQ imbalance

phase noise

nonlinearities

Tim Schenk (tim.schenk@philips.com), June 8, 2007

Moore’s Law


transistors
MOORE'S LAW
Dual-Core Intel® Itanium® 2 Processor
Intel® Itanium® 2 Processor
Intel® Itanium® Processor
Intel® Pentium® 4 Processor
Intel® Pentium® III Processor
Intel® Pentium® II Processor
Intel® Pentium® Processor
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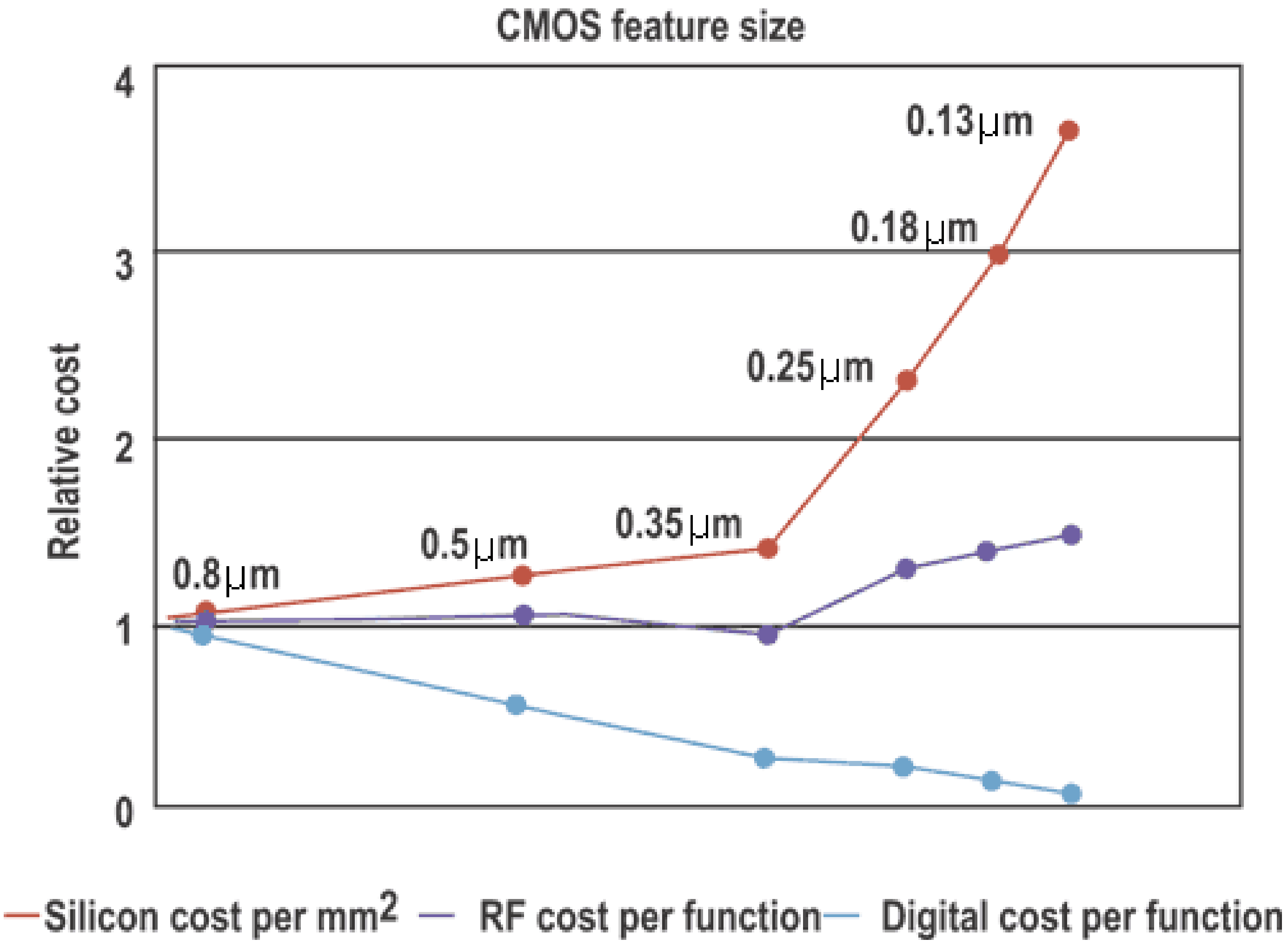
Moore’s Law helps digital



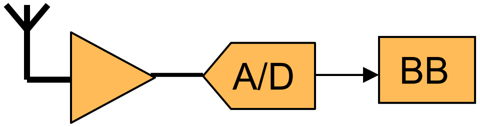
Towards co-design of baseband & RF

- Co-design of RF front-end and baseband part will result in most efficient solution
- Mean idea: perform task in part where it is most optimal for system performance, cost and power consumption.
- Problem: systems become increasingly complex: good understanding of front-end and baseband nessecarry.
- First step: Digital estimation and compensation of front-end impairments

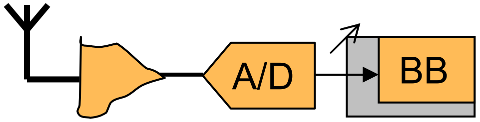
Tim Schenk (tim.schenk@philips.com), June 8, 2007

Technical innovations: roadmap of paradigms

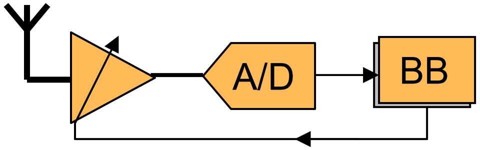
- Optimise and improve RF to meet specs



- Accept that RF has limitations
  - Baseband compensates RF imperfections

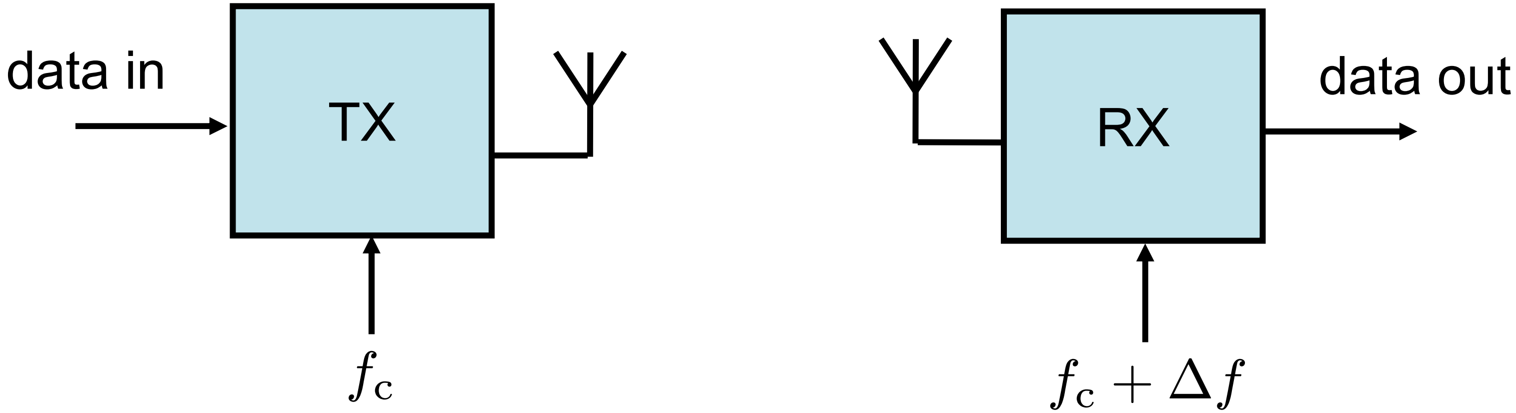


- Exploit the adaptively of RF
  - Control, Calibration and Compensation (C3): adaptively sets the parameters of RF
  - Adaptive RF controlled by BB to optimize RF settings



Slide: Jean-Paul Linnartz, Philips Research

Carrier Frequency Offset



- Difference in reference frequency between transmitter and receiver
- Present in all pratical radio systems
- Relatively easy to understand
- Most commonly studied and compensated front-end impairment: nice study case for today.

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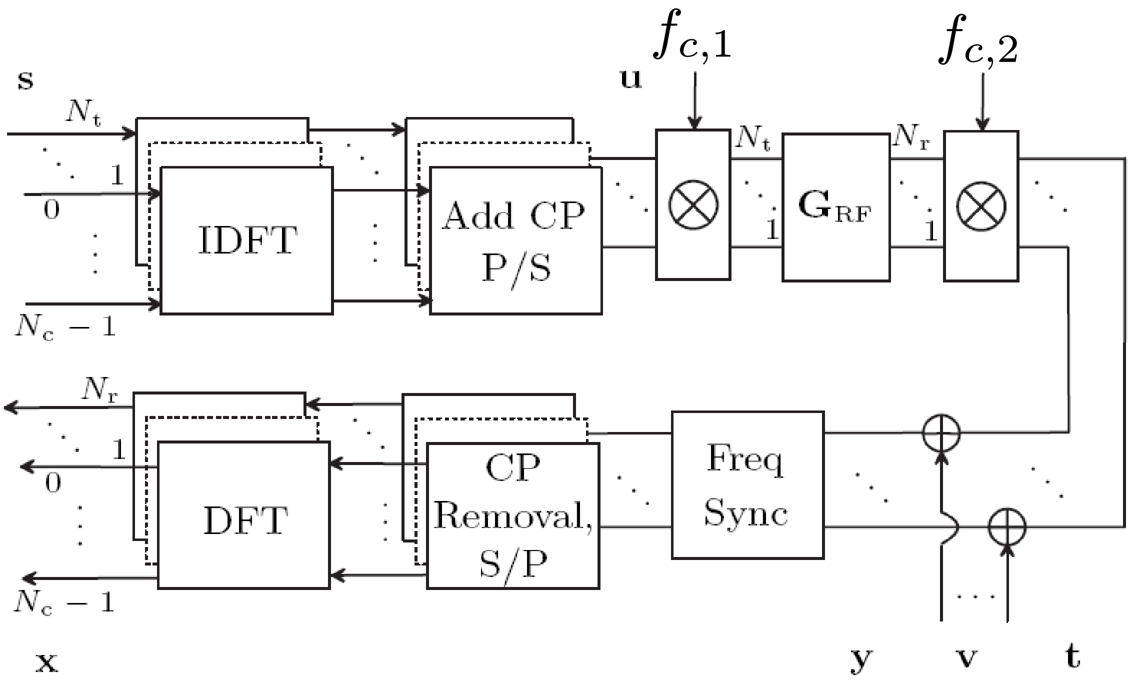
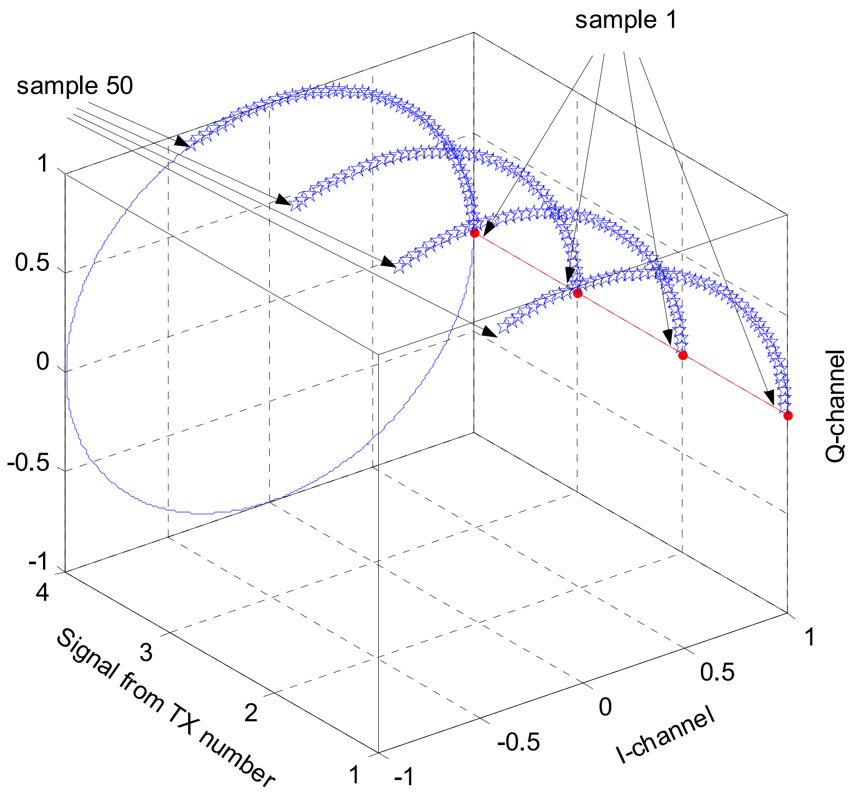
Carrier Frequency Offset – system model

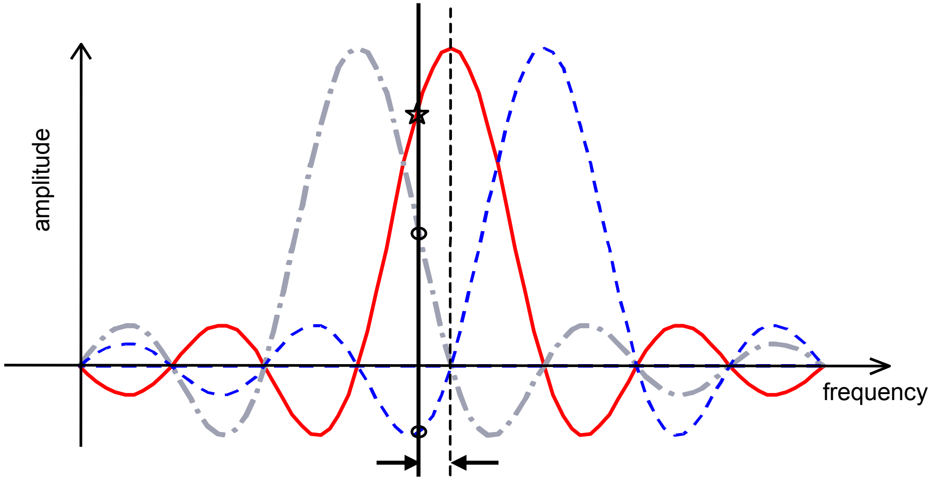


carrier frequency offset (CFO): $\Delta f = f_{c,2} - f_{c,1}$

received time-domain signal: $\hat{t}_m(n) = t_m(n) \exp\left(j2\pi \Delta f T_s (mN_s + n)\right)$

CFO – received signal





Time domain

Frequency domain

CFO – influence (I)

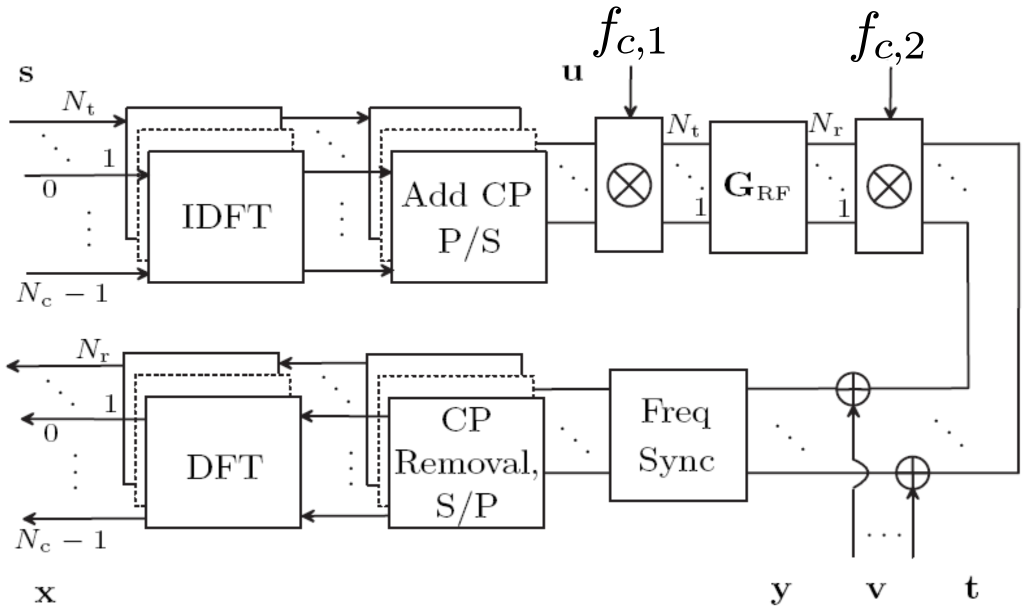
DFT CP removal phase shifts

$$\begin{aligned}\mathbf{x}_m &= (\mathbf{F}\boldsymbol{\Upsilon} \otimes \mathbf{I}_{N_\mathrm{r}})\mathbf{E}_m \mathbf{y}_m \\ &= (\mathbf{F}\boldsymbol{\Upsilon} \otimes \mathbf{I}_{N_\mathrm{r}})\mathbf{E}_m \breve{\mathbf{G}}(\boldsymbol{\Theta}\mathbf{F}^{-1} \otimes \mathbf{I}_{N_\mathrm{t}})\mathbf{s}_m + \mathbf{n}_m \\ &= (\boldsymbol{\mathcal{G}}_m \otimes \mathbf{I}_{N_\mathrm{r}})\,\mathbf{H}\mathbf{s}_m + \mathbf{n}_m \quad ,\end{aligned}$$

CP add

$$\mathbf{E}_m = \mathrm{diag}(e_m(0), e_m(1), \ldots, e_m(N_\mathrm{s}-1)) \otimes \mathbf{I}_{N_\mathrm{r}}$$

$$e_m(n) = \exp\left(j2\pi\Delta f T_\mathrm{s}(mN_\mathrm{s}+n)\right)$$



CFO – influence (II)

- Influence of CFO in (MIMO) OFDM systems
  - rotation of the received constellation point after the DFT
  - inter-carrier interference (ICI)

$$\begin{aligned}\mathbf{x}_m &= \left(\boldsymbol{\mathcal{G}}_m \otimes \mathbf{I}_{N_\mathrm{r}}\right)\mathbf{H}\mathbf{s}_m + \mathbf{n}_m \\ &= \left(\gamma_0 \mathbf{I}_{N_\mathrm{C} N_\mathrm{r}}\right)\mathbf{H}\mathbf{s}_m + \left(\boldsymbol{\mathcal{G}}_m - \gamma_0 \mathbf{I}_{N_\mathrm{C} N_\mathrm{r}}\right)\mathbf{H}\mathbf{s}_m + \mathbf{n}_m\end{aligned}$$

$$\boldsymbol{\mathcal{G}}_m = \begin{pmatrix} \gamma_0 & \gamma_{-1} & \cdots & \gamma_{-(N_\mathrm{C}-1)} \\ \gamma_1 & \gamma_0 & \cdots & \gamma_{-(N_\mathrm{C}-2)} \\ \vdots & \vdots & \ddots & \vdots \\ \gamma_{N_\mathrm{C}-1} & \gamma_{N_\mathrm{C}-2} & \cdots & \gamma_0 \end{pmatrix}$$

$$\gamma_q = \frac{\sin(\pi(\delta-q))}{N_\mathrm{C}\sin(\frac{\pi}{N_\mathrm{C}}(\delta-q))} e^{j\frac{\pi(N_\mathrm{C}-1)}{N_\mathrm{C}}(\delta-q)} e^{j\frac{2\pi\delta}{N_\mathrm{C}}(mN_\mathrm{S}+N_\mathrm{g})} \qquad \delta = \Delta f N_\mathrm{c} T_\mathrm{s}$$

Frequency synchronization (FS)

- *Estimate* the CFO and *correct* for its influence
- Time or frequency domain approach?

frequency: $\mathbf{x}_m = (\gamma_0 \mathbf{I}_{N_\mathrm{c} N_\mathrm{r}}) \mathbf{H}\mathbf{s}_m + (\boldsymbol{\mathcal{G}}_m - \gamma_0 \mathbf{I}_{N_\mathrm{c} N_\mathrm{r}}) \mathbf{H}\mathbf{s}_m + \mathbf{n}_m$

time: $\hat{y}_m(n) = y_m(n) \exp(j 2\pi \Delta f T_\mathrm{s}(m N_\mathrm{s} + n)) + v_m(n)$

- The phase difference between two (repeated) samples is linear dependent on CFO

$$t_m(n + N_\mathrm{p}) = t_m(n) \exp(j 2\pi \Delta f T_\mathrm{s} N_\mathrm{p})$$

- Frequency offset estimation by correlation:

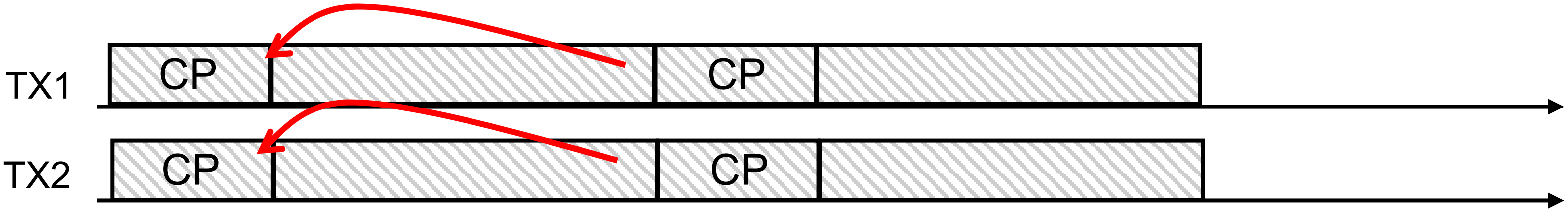
$$\Delta f = \frac{\angle \sum_{n=0}^{N_\mathrm{p}-1} t_m(n+N_\mathrm{p}) t_m^*(n)}{2\pi T_\mathrm{s} N_\mathrm{p}}$$

Tim Schenk (tim.schenk@philips.com), June 8, 2007

FS – time domain

- Use preamble (proposed for SISO [Moose95, Schmidl97])
  - Also used for channel estimation
  - Correlation repeated symbols is measure for CFO

- Use of Cyclic prefix (proposed for SISO [VanDeBeek97])
  - Poor performance in high delay spread channel (due to ISI)
  - No additional overhead



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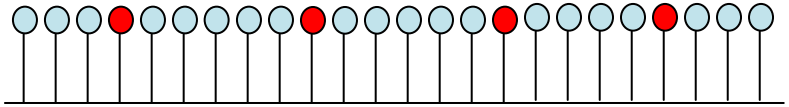
FS – frequency domain (I)

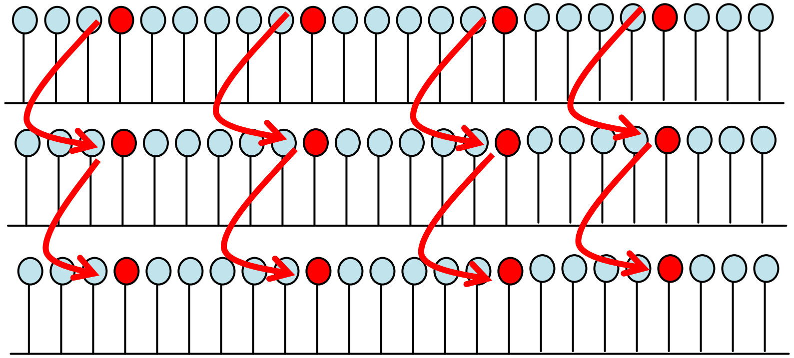
- Using pilot carriers
  - All carriers in a symbol experience the same rotation: estimate and rotate back.



$$\gamma_{0,n} \approx \exp(j\vartheta_1)$$

☹ Rotating back in freq. domain does not correct for ICI

  - using the relation between the phase rotations of the different symbols to estimate CFO: correct in time domain



$$\gamma_{0,n-1} \sim \exp(j(\alpha - \beta)\Delta f)$$

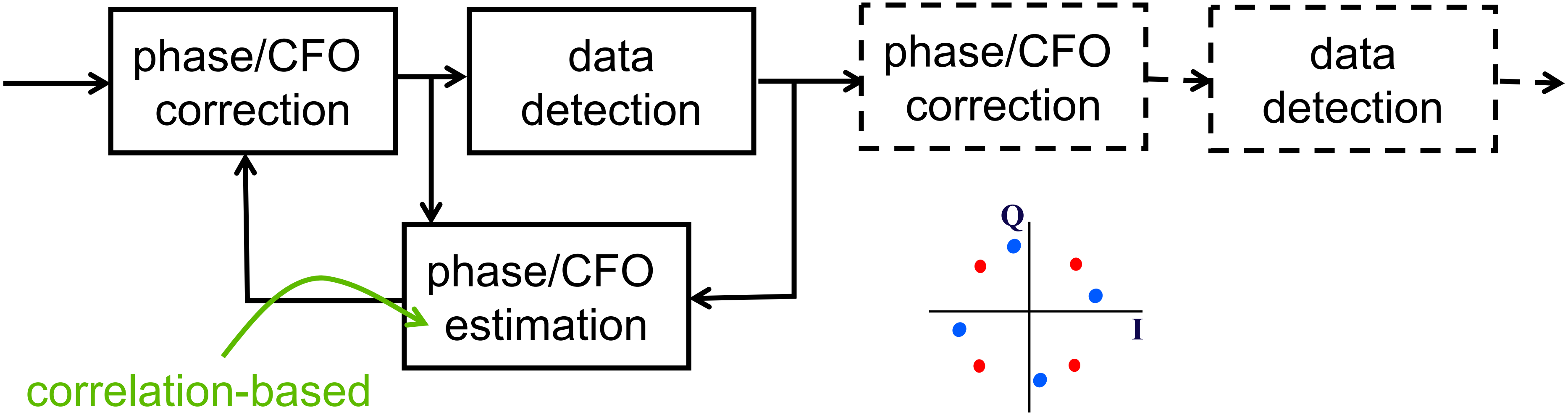
$$\gamma_{0,n} \sim \exp(j\alpha\Delta f)$$

$$\gamma_{0,n+1} \sim \exp(j(\alpha + \beta)\Delta f)$$

BPSK symbols as pilot

FS – blind techniques (I)

- Use detected data as “pilots” for phase of CFO estimation: decision directed

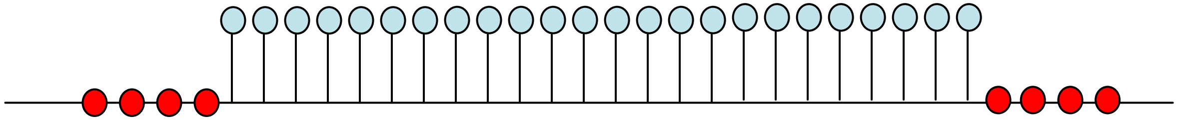


correlation-based

- Maximum detectable offset depends on used constellation: phase ambiguity

FS – blind techniques (II)

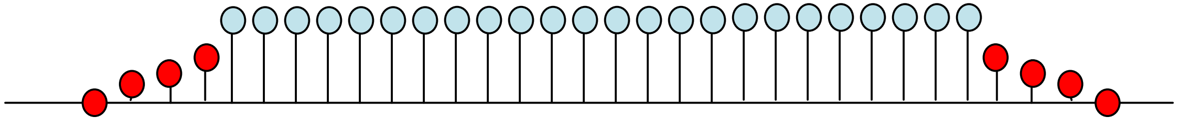
TX signal /
no CFO
RX signal



○ $\mathbf{x}_m(k) = \mathbf{H}(k)\mathbf{s}_m(k) + \mathbf{n}_m(k)$

● $\mathbf{x}_m(k) = \mathbf{n}_m(k)$

CFO-impaired
RX signal



○ $\mathbf{x}_m(k) = \gamma_0 \mathbf{H}(k)\mathbf{s}_m(k) + \sum_{\kappa,\kappa\neq k} \gamma_{k-\kappa}\mathbf{H}(\kappa)\mathbf{s}_m(\kappa) + \mathbf{n}_m(k)$

● $\mathbf{x}_m(k) = \sum_{\kappa,\kappa\neq k} \gamma_{k-\kappa}\mathbf{H}(\kappa)\mathbf{s}_m(\kappa) + \mathbf{n}_m(k)$

- ICI creates correlation between subcarriers
- Find CFO estimate that minimizes this correlation, using a cost function like:

$$\Delta\hat{f} = \arg\min_{\Delta f} \sum_{k\in K} \mathbf{x}_m(k)\mathbf{W}(\Delta f)\mathbf{x}_m$$

weighting matrix

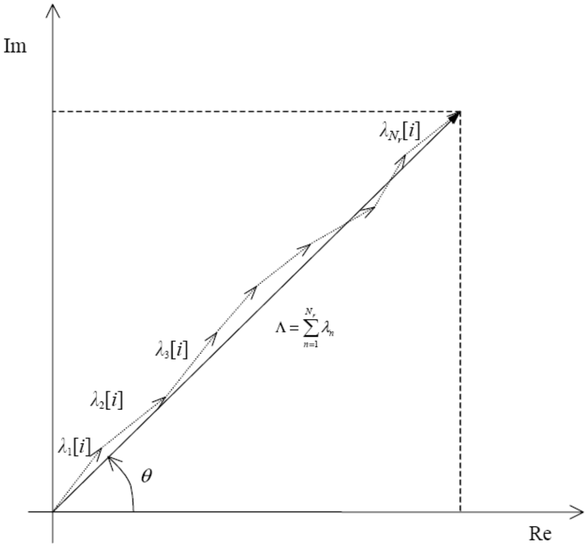
subset of carriers

FS - performance (I)

- As example we look at preamble-based technique
- Maximum CFO that can be estimated is limited by distance between correlated samples

$$|\delta_{\mathrm{max}}| = |\theta_{\mathrm{max}}| N_{\mathrm{c}}/2\pi N_{\mathrm{p}} = N_{\mathrm{c}}/2N_{\mathrm{p}} \qquad \delta = \Delta f N_{\mathrm{c}} T_{\mathrm{s}}$$

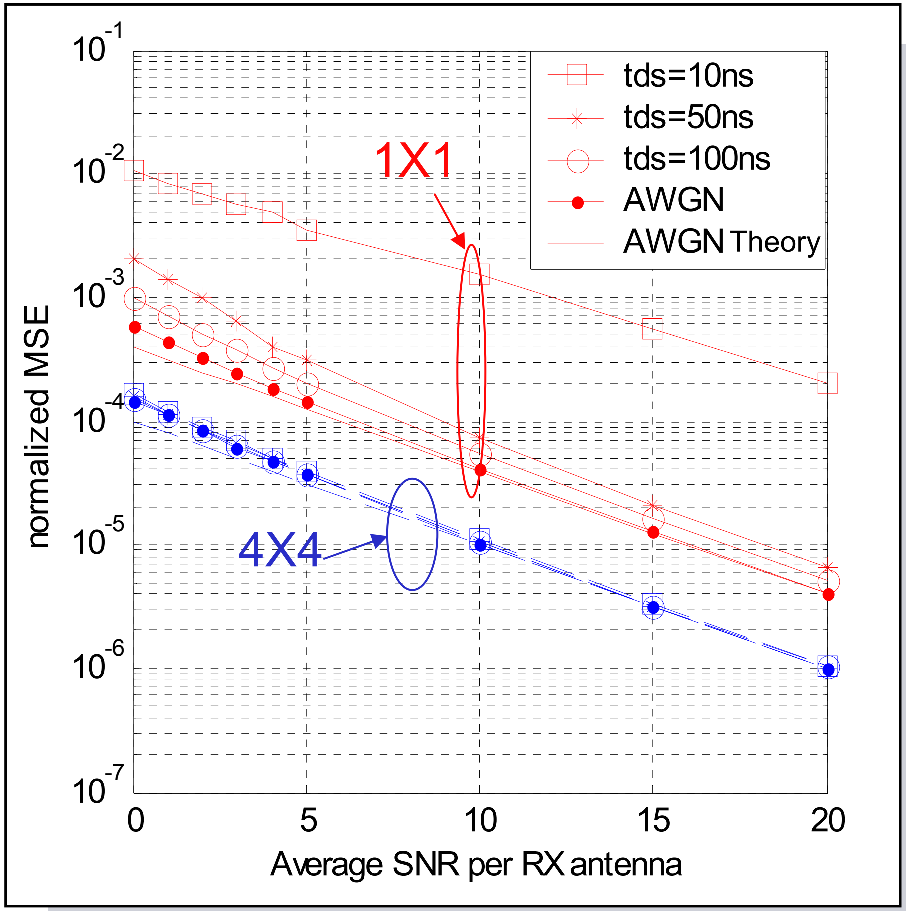
- MSE decreases with increasing N_{p}
- For MIMO sum the correlation outputs to achieve MRC like performance: not averaging of estimates!



FS - performance (II)

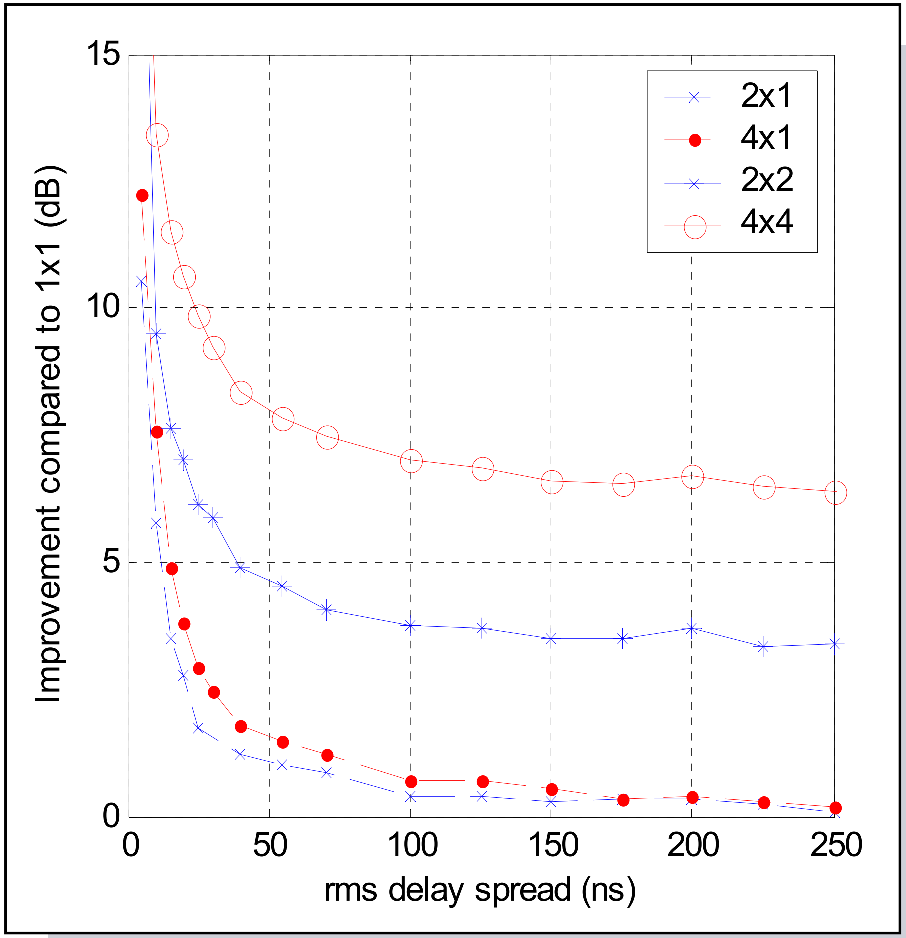
$$\mathrm{MSE}_{\delta,\mathrm{AWGN}} \approx \frac{N_\mathrm{c}^2}{(2\pi)^2 N_\mathrm{r} N_\mathrm{p}^3 \mathrm{SNR}}$$

- Results from theory and simulations for 1x1 and 4x4 AWGN / multipath
- Uncorrelated multipath channels, exp. PDP, Rayleigh fading every tap.
- Preamble-based estimation, length constant
- IEEE 802.11a parameters
- Degradation due to frequency fading is smaller in MIMO case.
- CFO $\delta = 0.2$



FS – MSE performance improvement

- Improvement in MSE compared to the SISO version.
- When frequency diversity arises, the gain of space diversity reduces.
- Allows for reduction of preamble length for higher order MIMO (for CFO estimation)



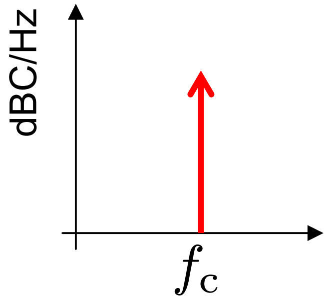
CFO/FS in (MIMO) OFDM – Additional reading

- Moose, “A technique for orthogonal frequency division multiplexing frequency offset correction”, IEEE Trans. on Commun., Oct 1994.
- Schmidl, et al., “Robust frequency and timing synchronization for OFDM,” IEEE Trans. on Commun., 1997.
- Yingwei Yao et. al., “Blind carrier frequency offset estimation in SISO, MIMO, and multiuser OFDM systems,” IEEE Trans. on Commun., Jan. 2005
- Schenk et al., “Frequency Synchronization for MIMO OFDM Wireless LAN Systems,” Proc. VTC-Fall 2003.
- Tureli et al., “Multicarrier synchronization with diversity,” Vehicular Technology Conference, 2001. Proc. IEEE VTC-Fall 2001.

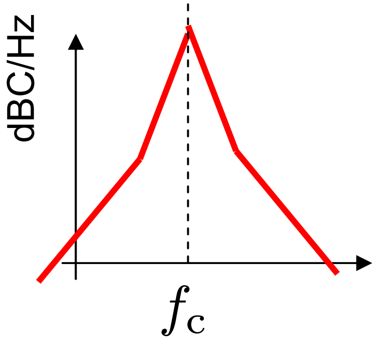
Phase Noise

- Imperfections in RF oscillators have big impact on (MIMO) OFDM performance
- Amplitude disturbances in RF oscillators are marginal
- Random frequency deviation of RF carrier are often modelled as excess phase deviation: *phase noise*.
- Oscillator stabillity becomes larger issue for low-cost implementations (crystal less) and high carrier frequencies

ideal



realistic



Influence of carrier frequency on Phase Noise

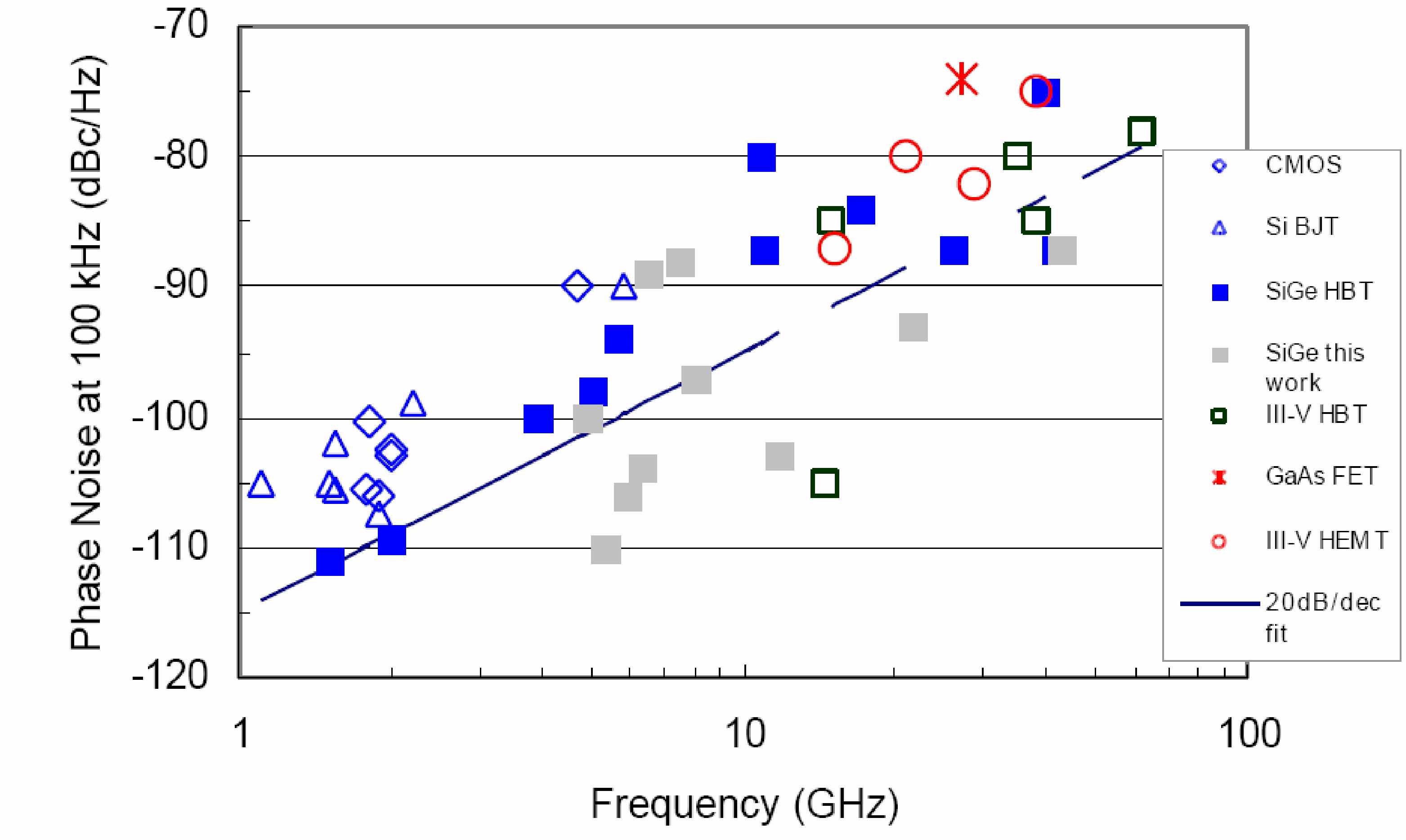
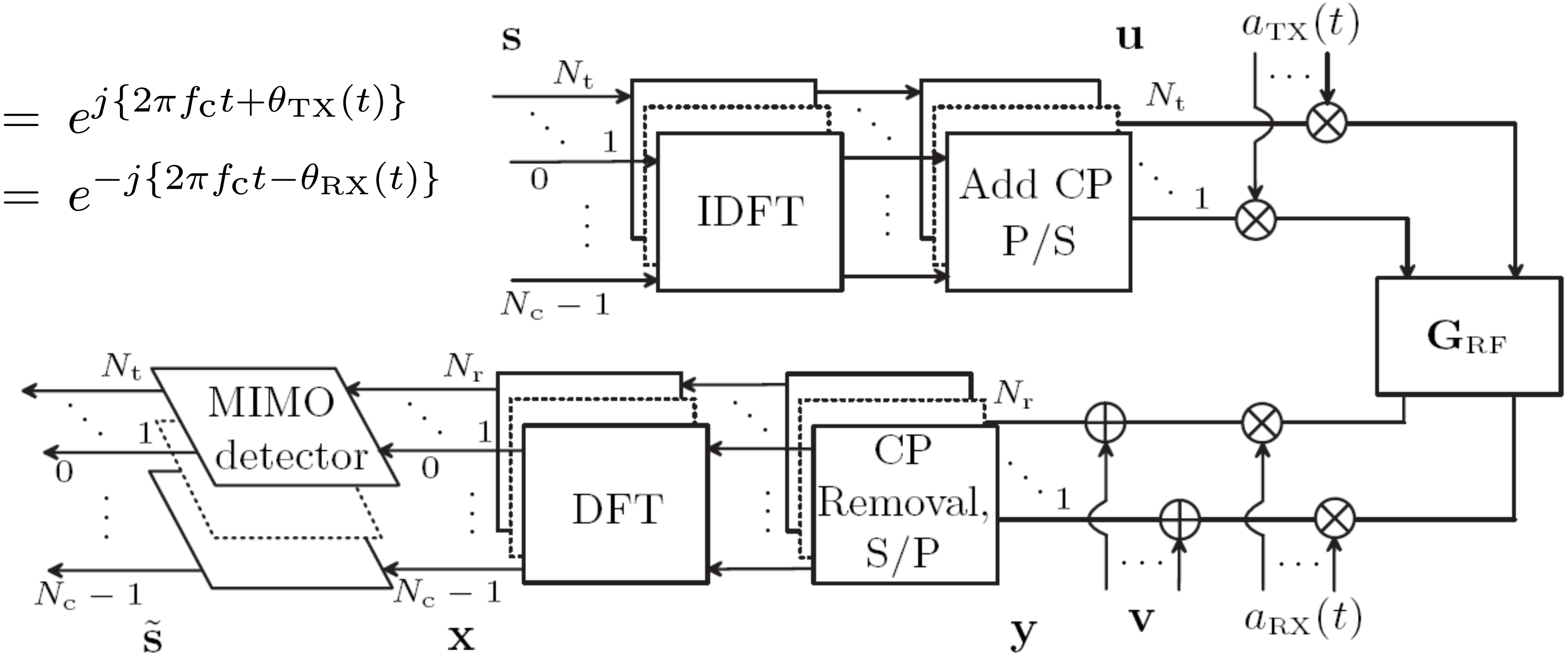


Figure: Collected phase-noise data of fully integrated VCOs

Li, et.al, (Ericsson AB), “High-frequency SiGe MMICs − an Industrial Perspective”

Phase Noise – system model

$$a_{\mathrm{TX}}(t) = e^{j\{2\pi f_c t + \theta_{\mathrm{TX}}(t)\}}$$
$$a_{\mathrm{RX}}(t) = e^{-j\{2\pi f_c t - \theta_{\mathrm{RX}}(t)\}}$$



DFT

CP removal

Add CP

$$\mathbf{x}_m = (\mathbf{F}\boldsymbol{\Upsilon} \otimes \mathbf{I}_{N_r})((\mathbf{E}_{\mathrm{RX},m} \otimes \mathbf{I}_{N_r})\breve{\mathbf{G}}(\mathbf{E}_{\mathrm{TX},m} \otimes \mathbf{I}_{N_t})(\boldsymbol{\Theta}\mathbf{F}^{-1} \otimes \mathbf{I}_{N_t})\mathbf{s}_m + \mathbf{v}_m)$$

phase noise vector

$$\mathbf{E}_{\mathrm{x},m} = \mathrm{diag}\{a_{\mathrm{x},m}(0), a_{\mathrm{x},m}(1), \ldots, a_{\mathrm{x},m}(N_s - 1)\}$$

with $a_{\mathrm{x},m}(n) = a_{\mathrm{x}}(mN_s + n)$ and $\mathrm{X} \in \{\mathrm{TX}, \mathrm{RX}\}$

PN – influence (I)

$$\begin{aligned}\mathbf{x}_m &= (\mathbf{F}\boldsymbol{\Upsilon}\otimes\mathbf{I}_{N_\mathrm{r}})((\mathbf{E}_{\mathrm{RX},m}\otimes\mathbf{I}_{N_\mathrm{r}})\breve{\mathbf{G}}(\mathbf{E}_{\mathrm{TX},m}\otimes\mathbf{I}_{N_\mathrm{t}})(\boldsymbol{\Theta}\mathbf{F}^{-1}\otimes\mathbf{I}_{N_\mathrm{t}})\mathbf{s}_m+\mathbf{v}_m)\\ &= (\boldsymbol{\mathcal{G}}_{\mathrm{RX},m}\otimes\mathbf{I}_{N_\mathrm{r}})\mathbf{H}(\boldsymbol{\mathcal{G}}_{\mathrm{TX},m}\otimes\mathbf{I}_{N_\mathrm{t}})\mathbf{s}_m+\mathbf{n}_m\end{aligned}$$

RX PN influence

TX PN influence

where

$$\begin{aligned}\boldsymbol{\mathcal{G}}_{\mathrm{TX},m} &= \mathbf{F}\boldsymbol{\Upsilon}\mathbf{E}_{\mathrm{TX},m}\boldsymbol{\Theta}\mathbf{F}^{-1}\\ \boldsymbol{\mathcal{G}}_{\mathrm{RX},m} &= \mathbf{F}\boldsymbol{\Upsilon}\mathbf{E}_{\mathrm{RX},m}\boldsymbol{\Theta}\mathbf{F}^{-1}\end{aligned}$$

$$\boldsymbol{\mathcal{G}}_{\mathrm{X},m}=\begin{bmatrix}\gamma^{\mathrm{X}}_{0,m} & \gamma^{\mathrm{X}}_{1,m} & \cdots & \gamma^{\mathrm{X}}_{N_\mathrm{C}-1,m}\\ \gamma^{\mathrm{X}}_{-1,m} & \gamma^{\mathrm{X}}_{0,m} & \cdots & \gamma^{\mathrm{X}}_{N_\mathrm{C}-2,m}\\ \vdots & \vdots & \ddots & \vdots\\ \gamma^{\mathrm{X}}_{-N_\mathrm{C}+1,m} & \gamma^{\mathrm{X}}_{-N_\mathrm{C}+2,m} & \cdots & \gamma^{\mathrm{X}}_{0,m}\end{bmatrix}$$

with $\gamma^{\mathrm{X}}_{k-l,m}=\frac{1}{N_\mathrm{C}}\sum_{i=0}^{N_\mathrm{C}-1}e^{j\theta_{\mathrm{X},m}(N_\mathrm{g}+i)}e^{-j\frac{2\pi\{k-l\}i}{N_\mathrm{C}}}$

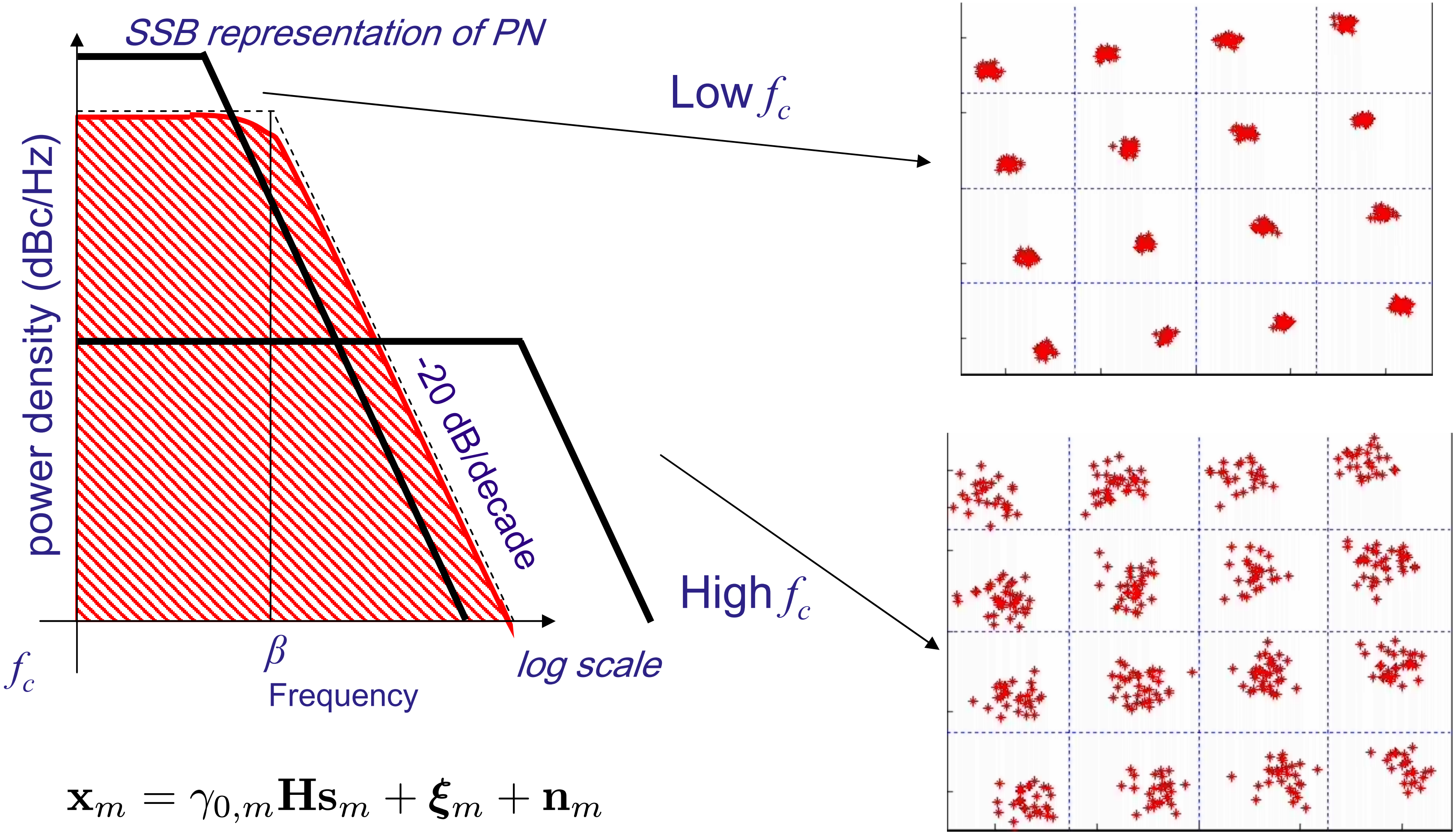
phase noise

$$\boxed{\mathbf{x}_m=\overbrace{\gamma^{\mathrm{RX}}_{0,m}\gamma^{\mathrm{TX}}_{0,m}}^{\gamma_{0,m}}\mathbf{H}\mathbf{s}_m+\boldsymbol{\xi}_m+\mathbf{n}_m}$$

common phase error (CPE)
= average PN over OFDM symbol

inter-carrier interference (ICI)

PN – influence (II)



$$\mathbf{x}_m = \gamma_{0,m}\mathbf{H}\mathbf{s}_m + \boldsymbol{\xi}_m + \mathbf{n}_m$$

CPE – ML estimation and correction

$$\mathbf{x}_m = \gamma_{0,m}\mathbf{H}\mathbf{s}_m + \boldsymbol{\xi}_m + \mathbf{n}_m \qquad \text{with} \qquad \gamma_{0,m} = \frac{1}{N_\mathrm{c}}\sum_{i=0}^{N_\mathrm{c}-1} e^{j\theta_m(N_\mathrm{g}+i)}$$

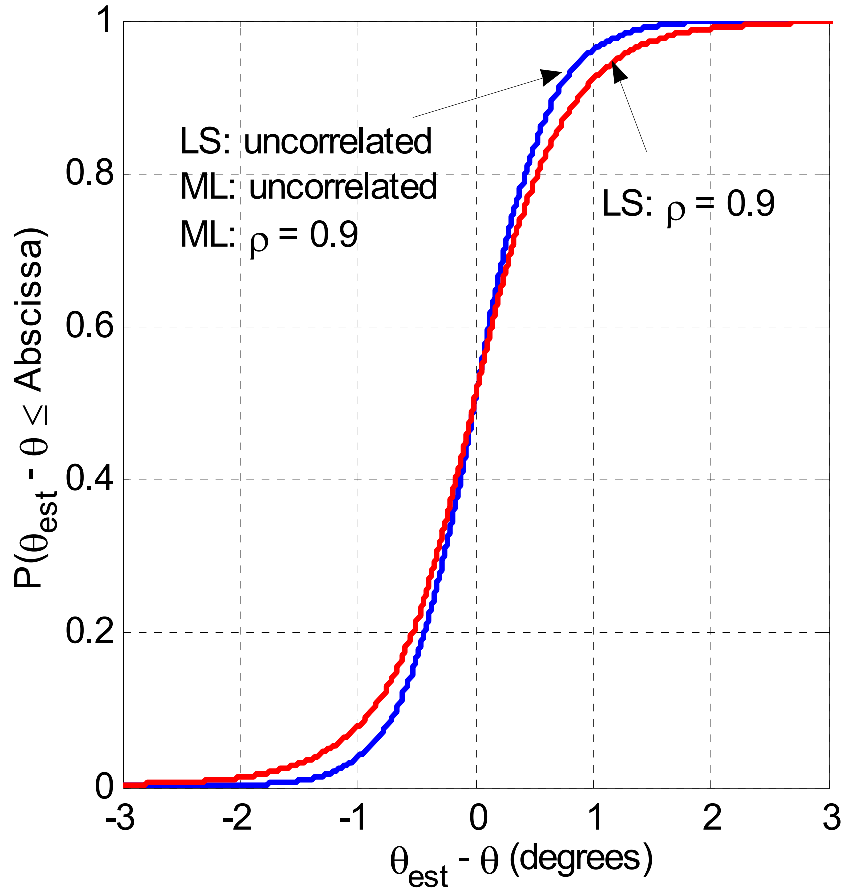
- Phase rotation due to CPE hase largest impact on performance → estimation and correction
- Pilots are “required”, since 1 estimation per OFDM symbol
- Let us derive the Maximum Likelihood Estimator (MLE). The error term is given by

$$\mathbf{z}_m = \mathbf{x}_m - \gamma_{0,m}\mathbf{H}\mathbf{s}_m = \boldsymbol{\xi}_m + \mathbf{n}_m$$

- Expression for ICI includes $\mathbf{H}$ so can exhibit (spatial) correlation.
- If no correlation: MLE reduces to least-squares estimator (LSE).

CPE – estimation performance (I)

- P_{PN} = −30 dBc and
  β = 200 kHz.
- Independent Rayleigh fading, spatial correlation.
- IEEE 802.11a based:
  64 subcarriers, 20 MHz,
  64-QAM, 4 pilots, no AWGN,
  no coding, no AWGN.
- **Performance equal for no correlation, MLE better for correlated channels**

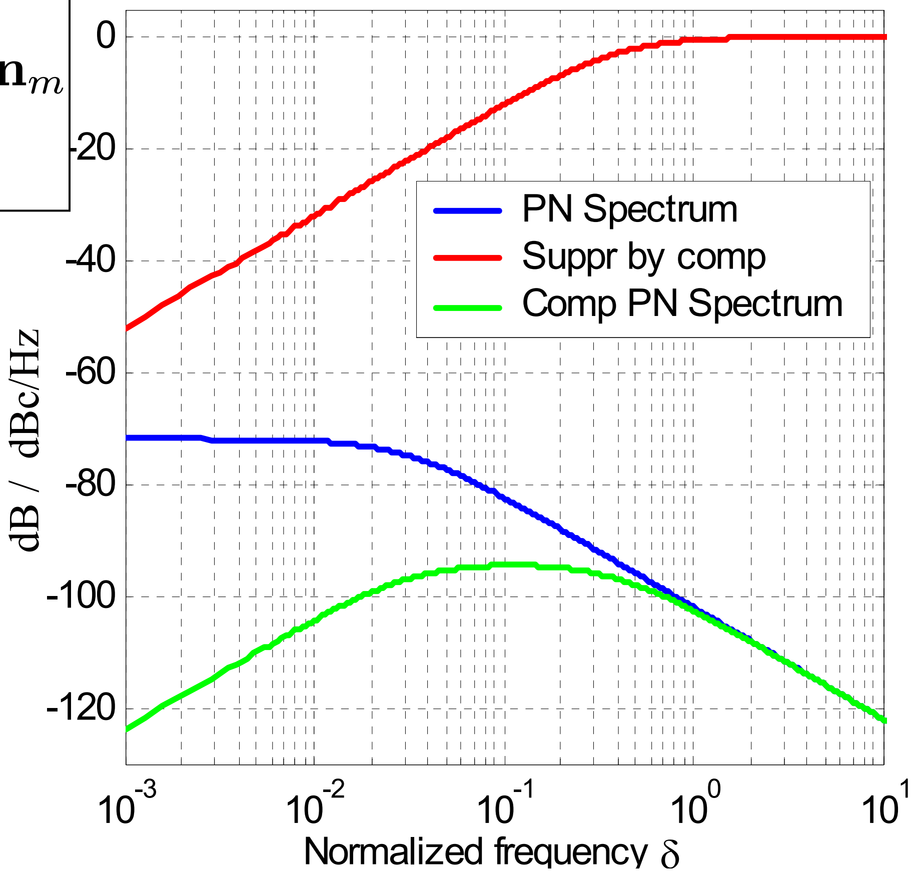


ECDF of the error in estimation of CPE for LS and ML estimator for 2x2.

PN spectrum w/wo compensation

$$\begin{aligned}\tilde{\mathbf{x}}_m &= \hat{\gamma}^*_{0,m}\mathbf{x}_m \\ &= \hat{\gamma}^*_{0,m}\gamma_{0,m}\mathbf{H}\mathbf{s}_m + \hat{\gamma}^*_{0,m}\boldsymbol{\xi}_m + \hat{\gamma}^*_{0,m}\mathbf{n}_m \\ &\approx \mathbf{H}\mathbf{s}_m + \boldsymbol{\xi}'_m + \mathbf{n}'_m\end{aligned}$$

- Using the estimated values of the CPE, we compensate the received signal.
- From that we find suppression and the resulting PN spectrum after compensation.
- Clearly lower frequencies in the PN spectrum are suppressed.



PN spectrum before/after compensation, β =3.2·10^{-2}, P_{PN} =*-30 dBc.*

PN – inter-carrier interference (ICI)

well studied problem

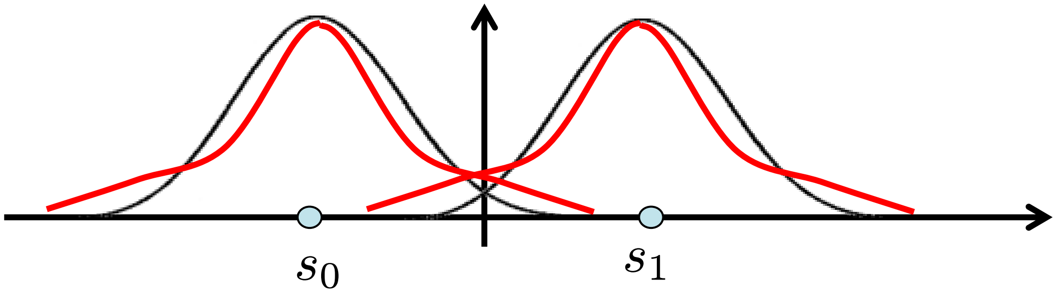
$$\mathbf{x} = \mathbf{H}\mathbf{s} + \mathbf{n} + \boldsymbol{\xi}$$

ICI

- ICI term generally assumed to have a zero-mean complex Gaussian distribution → Central Limit Theorem

$$\xi_m(k) = \sum_{l=0, l\neq k}^{N_\mathrm{C}-1} \gamma_{k-l,m} s_m(l)$$

- This is, however, not true and results in an underestimation of the bit-error probability. [Schenk, vd Hofstad, et al. *Trans. Wirel. Comm.*, 2007]



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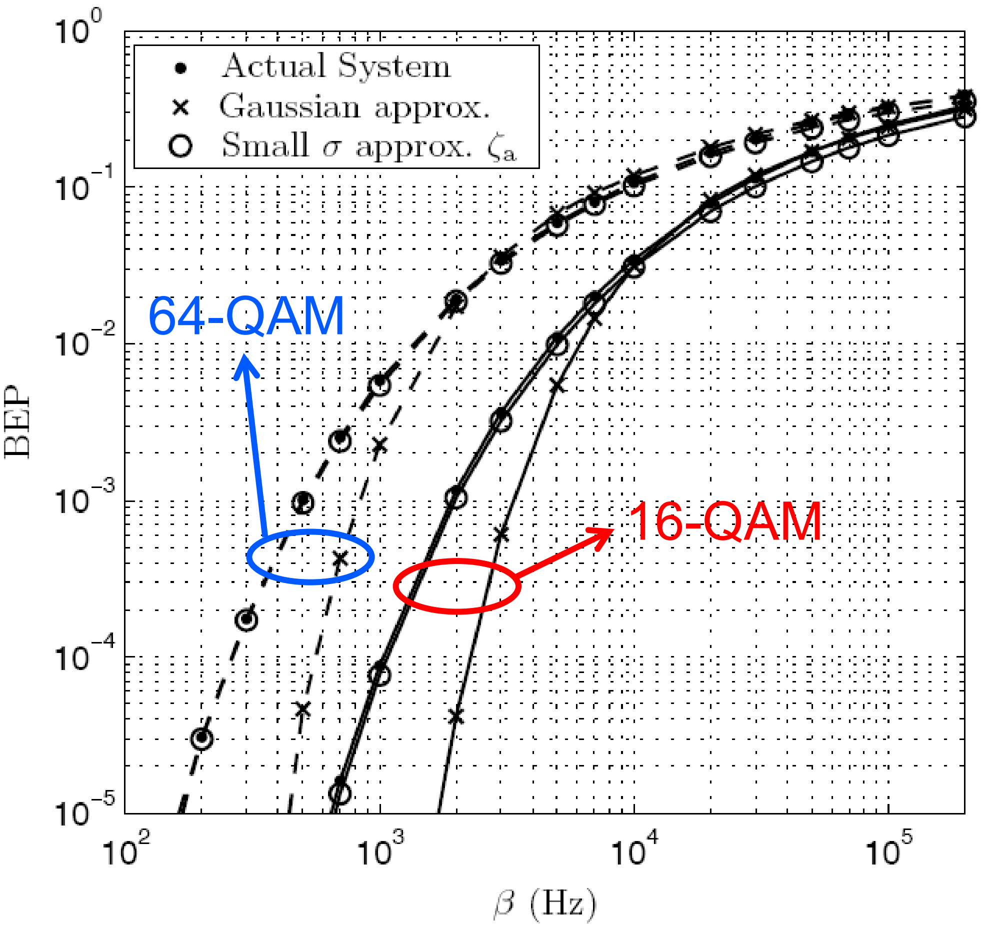
Distribution of the ICI

- ICI distribution has thicker tails than Gaussian distr. with same mean & var.
- Generally applied Gaussian approximation yields underestimation of BEP → under specification of oscillator.
- CLT does not hold due to fast decrease of γ_k
- Correct limit distribution available in Trans. WComm. Paper.



Results for 1x1 802.11a-like system

BER-impact of ICI in MIMO OFDM

- Study of BER performance for linear MIMO system applying ZF-detection. $\hat{\mathbf{s}}(k) = \mathbf{H}^{\dagger}(k)\mathbf{x}(k) = \mathbf{s}(k) + \mathbf{H}^{\dagger}(k)\boldsymbol{\xi}(k) + \mathbf{H}^{\dagger}(k)\mathbf{n}(k)$

$$= \mathbf{s}(k) + \boldsymbol{\Xi}(k) + \mathbf{H}^{\dagger}(k)\mathbf{n}(k)$$

- For (dominant) TX phase noise, the ICI-caused error term can be written as $\boldsymbol{\Xi}(k) = \sum_{l=1,l\neq k}^{N_\mathrm{C}} \gamma_{k-l}^{\mathrm{TX}}\mathbf{s}(l)$ → no influence of channel

$$\sigma^2_{\boldsymbol{\Xi}(k)} = \sigma_\mathrm{S}^2 \sum_{l=1,l\neq k}^{N_\mathrm{C}} \mathbb{E}\left[\left|\gamma_{k-l}^{\mathrm{TX}}\right|^2\right] = \frac{2\sigma_\mathrm{S}^2\pi\beta T_\mathrm{S}(N_\mathrm{C}^2-1)}{3N_\mathrm{C}}$$

- For (dominant) RX phase noise, the ICI-caused error term can be written as $\boldsymbol{\Xi}(k) = \mathbf{H}^{\dagger}(k)\sum_{l=1,l\neq k}^{N_\mathrm{C}} \gamma_{k-l}^{\mathrm{RX}}\mathbf{H}(l)\mathbf{s}(l)$ → interaction between carrier k and l

Flat fading: $\sigma^2_{\boldsymbol{\Xi}(k)} = \sigma_\mathrm{S}^2 \sum_{l=1,l\neq k}^{N_\mathrm{C}} \mathbb{E}\left[\left|\gamma_{k-l}^{\mathrm{RX}}\right|^2\right] = \frac{2\sigma_\mathrm{S}^2\pi\beta T_\mathrm{S}(N_\mathrm{C}^2-1)}{3N_\mathrm{C}}$

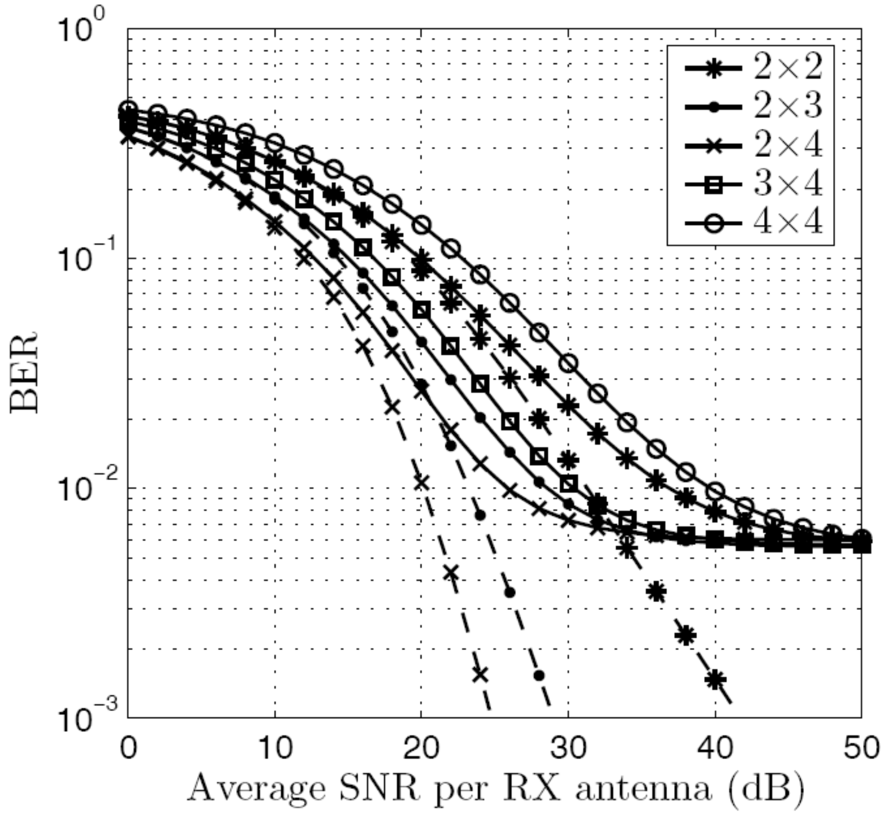
Independent fading: $\sigma^2_{\boldsymbol{\Xi}(k)} = \sigma_\mathrm{S}^2 \mathrm{tr}\left\{\mathbb{E}\left[\boxed{\left(\mathbf{H}(k)\mathbf{H}^{H}(k)\right)^{\dagger}}\right]\right\} \sum_{l=1,l\neq k}^{N_\mathrm{C}} \mathbb{E}\left[\left|\gamma_{k-l}^{\mathrm{RX}}\right|^2\right]$

$$\sigma^2_{\boldsymbol{\Xi}(k)} = \frac{N_\mathrm{t}}{N_\mathrm{r}-N_\mathrm{t}}\sigma_\mathrm{S}^2 \sum_{l=1,l\neq k}^{N_\mathrm{C}} \mathbb{E}\left[\left|\gamma_{k-l}^{\mathrm{RX}}\right|^2\right] = \frac{2\sigma_\mathrm{S}^2\pi\beta T_\mathrm{S}N_\mathrm{t}(N_\mathrm{r}-N_\mathrm{t})(N_\mathrm{C}^2-1)}{3N_\mathrm{C}} \quad \text{for} \quad N_\mathrm{r} > N_\mathrm{t}$$

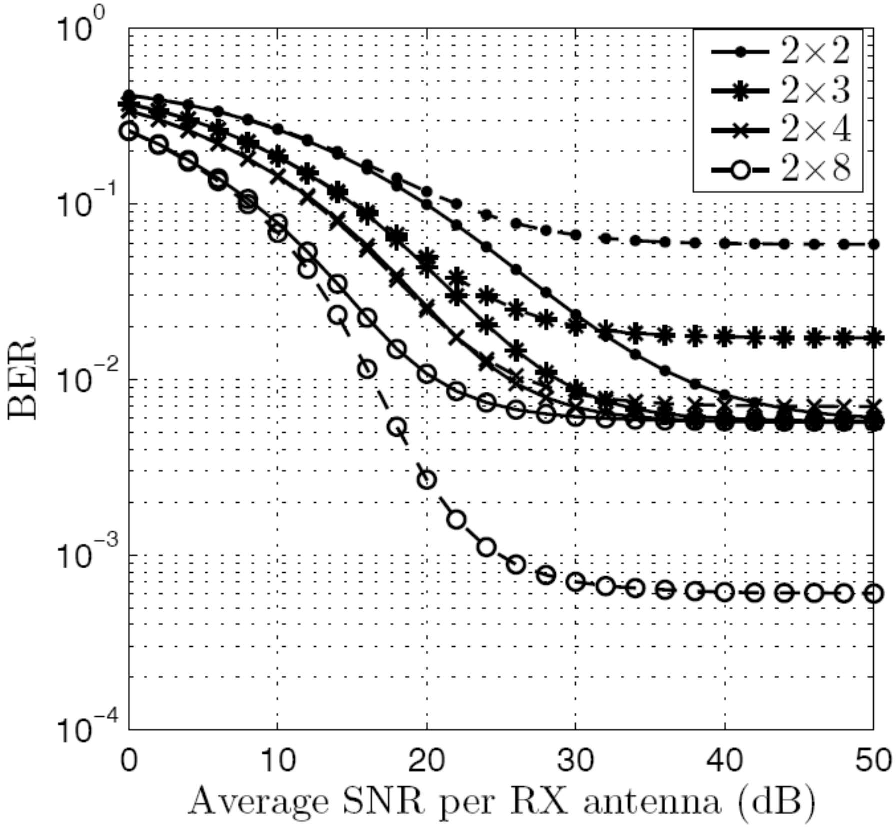
→ complex inverse Wishart distr.

Tim Schenk (tim.schenk@philips.com), June 8, 2007

BER-impact ICI – numerical



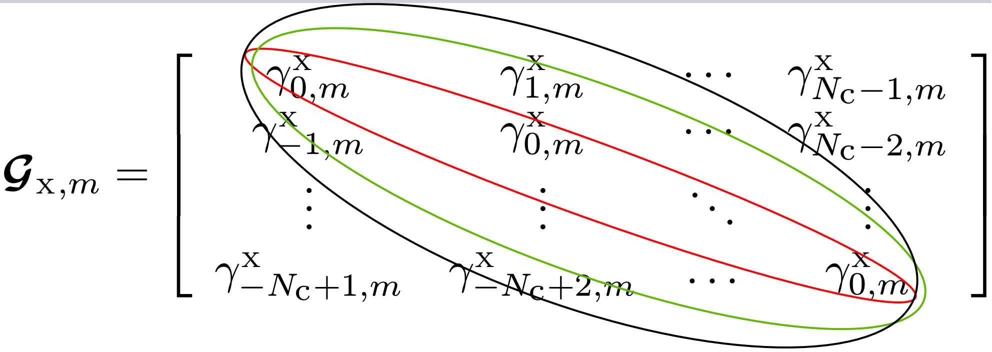
Flat fading, no PN (dashed),
β = 1000 Hz (solid), 64 QAM



Independent fading, *β* = 1000 Hz,
TX PN (solid), RX PN (dashed),
64 QAM

ICI – estimation and suppression (I)

$$\mathbf{x}_m = (\boldsymbol{\mathcal{G}}_{\mathrm{RX},m} \otimes \mathbf{I}_{N_\mathrm{r}})\mathbf{H}\mathbf{s}_m + \mathbf{n}_m$$



$$\boldsymbol{\mathcal{G}}_{\mathrm{x},m} = \begin{bmatrix} \gamma^{\mathrm{x}}_{0,m} & \gamma^{\mathrm{x}}_{1,m} & \cdots & \gamma^{\mathrm{x}}_{N_\mathrm{c}-1,m} \\ \gamma^{\mathrm{x}}_{-1,m} & \gamma^{\mathrm{x}}_{0,m} & \cdots & \gamma^{\mathrm{x}}_{N_\mathrm{c}-2,m} \\ \vdots & \vdots & \ddots & \vdots \\ \gamma^{\mathrm{x}}_{-N_\mathrm{c}+1,m} & \gamma^{\mathrm{x}}_{-N_\mathrm{c}+2,m} & \cdots & \gamma^{\mathrm{x}}_{0,m} \end{bmatrix}$$

- More terms of PN terms can be estimated than CPE only, using DFT bases.
- Only lower orders of γ_k are of importance. First harmonics are most important
- Estimation requires too much pilot data → decision direct approach.

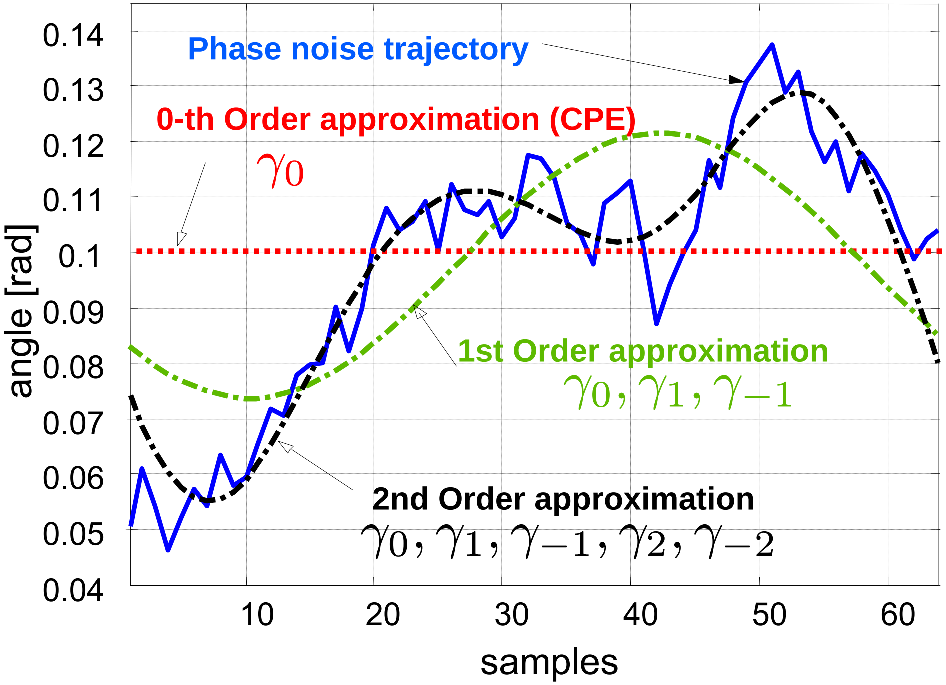


Figure: Denis Petrovic, Rhode & Schwarz

ICI – estimation and suppression (II)

- Decision directed approach.
- Aims to estimate phase noise waveform.
- Performance improvement compared to standard CPE correction
  → sensitive to fading channels

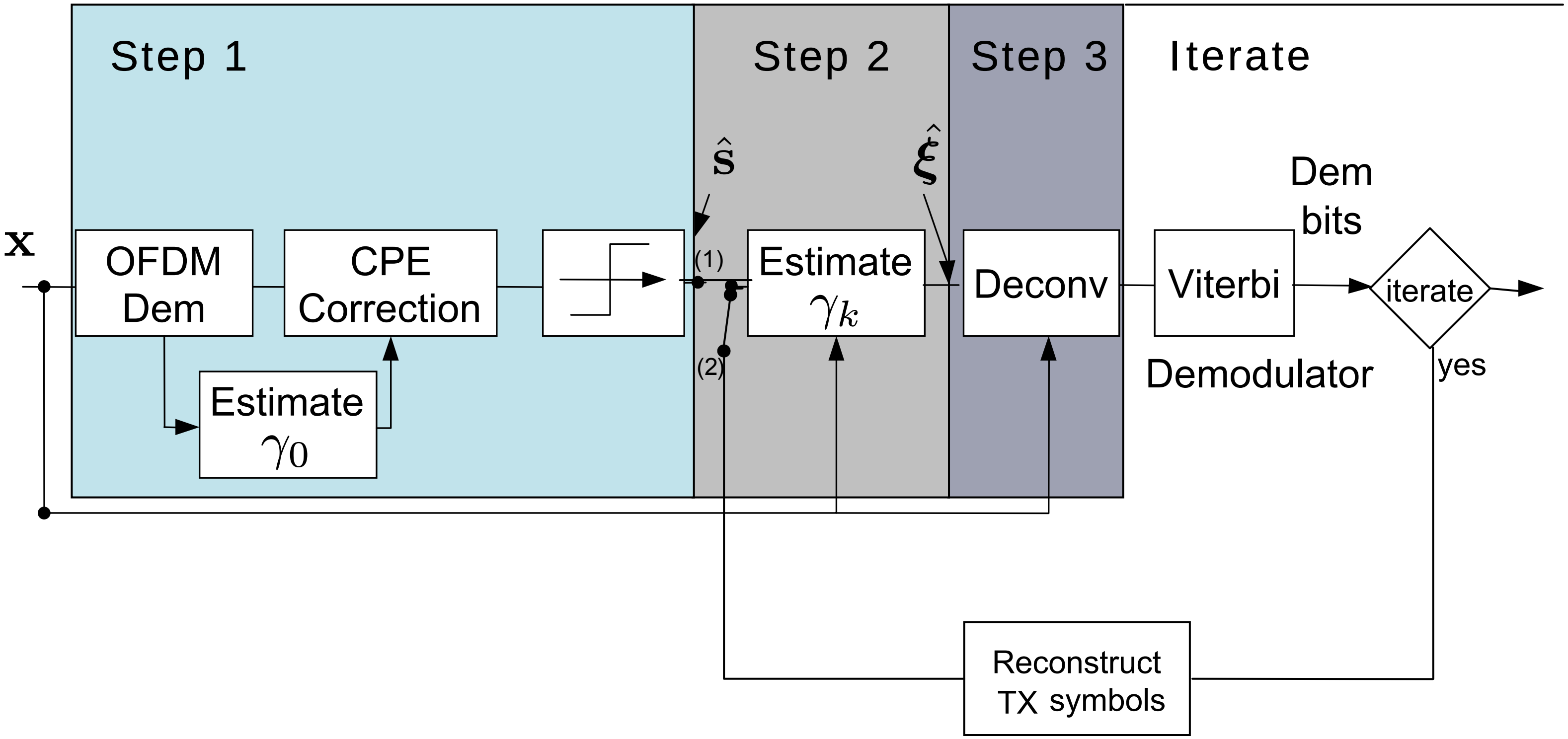


Figure: Denis Petrovic, Rhode & Schwarz

PN in (MIMO) OFDM – Additional reading

- Pollet, et al., “BER sensitivity of OFDM systems to carrier frequency offset and Wiener phase noise,” *IEEE Trans. on Commun.*, Feb./Mar./Apr. 1995.
- Steendam et al., “The effect of carrier phase jitter on the performance of OFDMA systems,” *IEEE Trans. Comm.*, April 1998.
- Piazzo, et al., “Analysis of phase noise effects in OFDM modems,” *IEEE Trans. on Commun.*, Oct. 2002.
- Casas, et al., “Time domain phase noise correction for OFDM signals,” *IEEE Trans. on Broadcasting*, Sept. 2002.
- Petrovic, et. al, “Intercarrier interference due to Phase Noise in OFDM - estimation and suppression,” *Proc. IEEE VTC-Fall 2004.*
- Schenk, et al., “Influence and suppression of phase noise in multi-antenna OFDM”, *Proc. IEEE VTC-Fall 2004.*
- Schenk, et al., “Distribution of the ICI term in Phase Noise impaired OFDM systems,” *IEEE Trans. Wirel. Comm.*, April 2007.

Summary – Part II

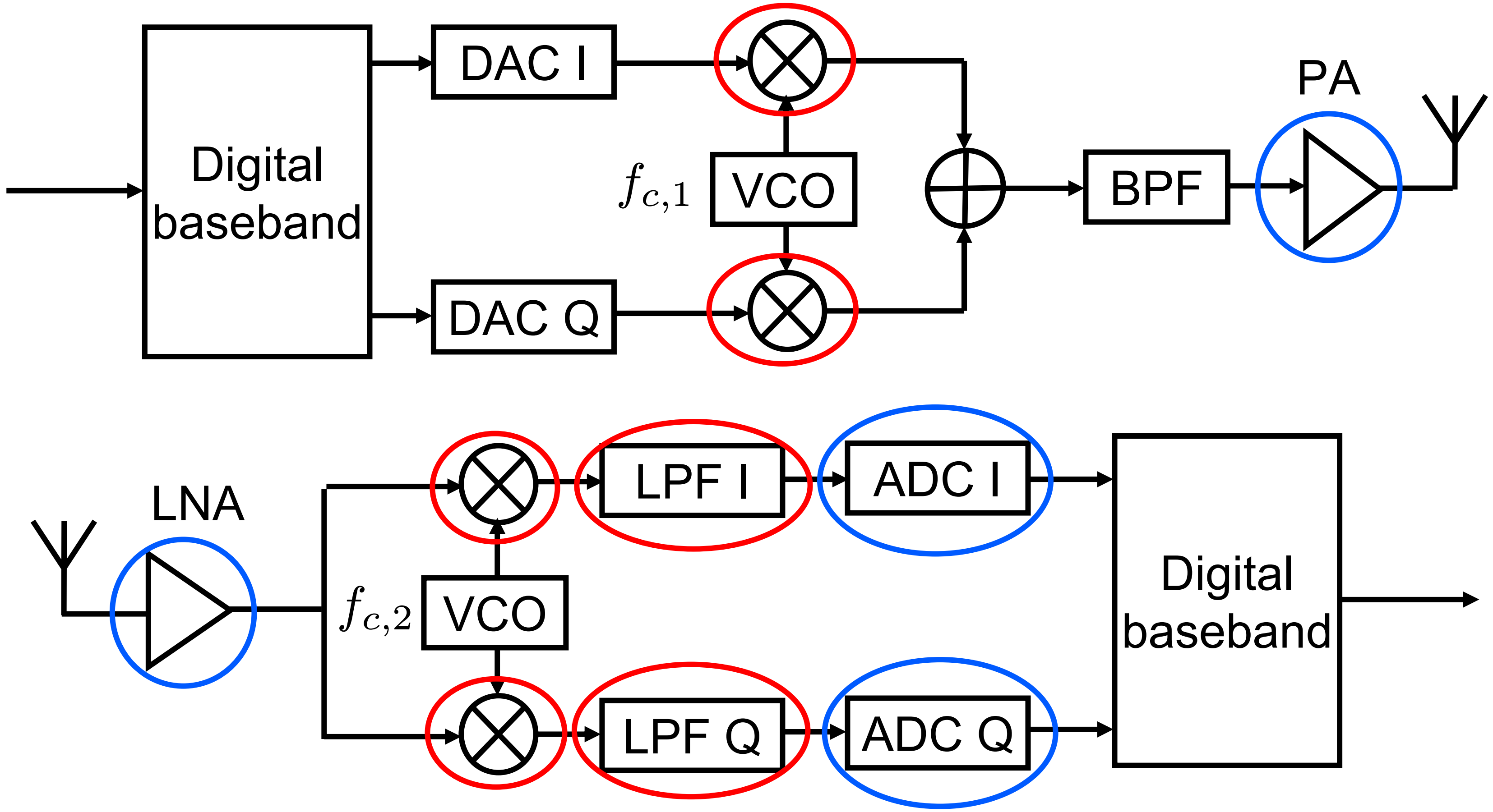
- Digital compensation requires good understanding of RF impairments
- Both CFO and Phase Noise introduce
  - constellation rotation, and
  - inter-carrier interference
- Different estimation and compensation approaches have been presented

Tim Schenk (tim.schenk@philips.com), June 8, 2007

Part III: IQ imbalance and Nonlinearities

- IQ imbalance
  - What is the influence?
  - How to treat it?
- Nonlinearities
  - What is the influence?
  - How to treat it?

Transceiver structure - impairments

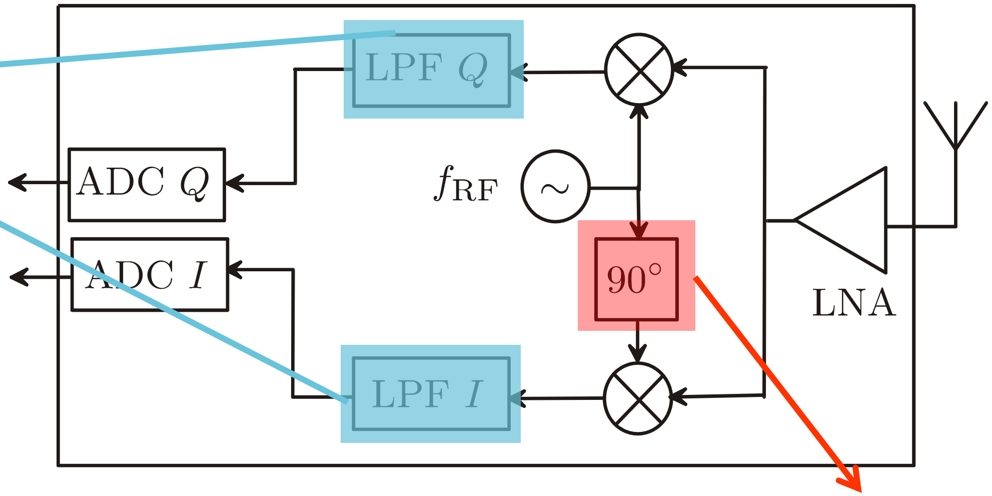


IQ imbalance

nonlinearities

IQ mismatch

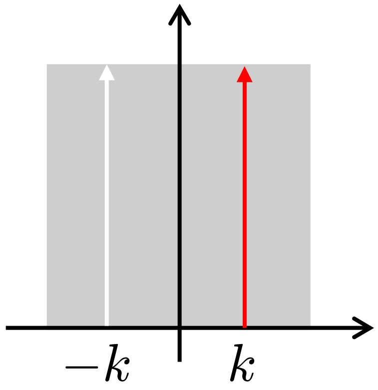
- Direct-conversion enables monolithic integration
- Quadrature mixing is performed in analog part. Can result in IQ mismatch: Phase and amplitude difference between I and Q arms
- Occurs in TX and RX, varies per front-end, time-invariant



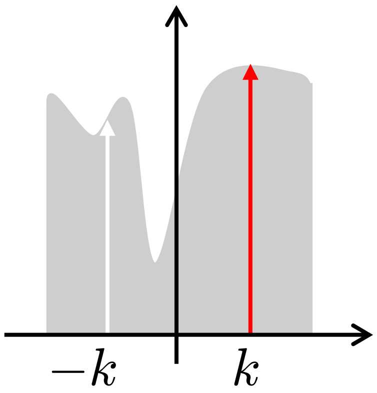
Frequency Selective

Frequency Independent

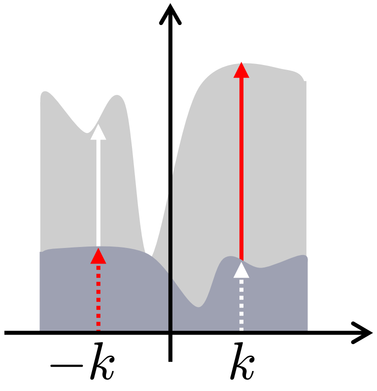
Influence of IQ imbalance (I)



TX baseband



RX RF

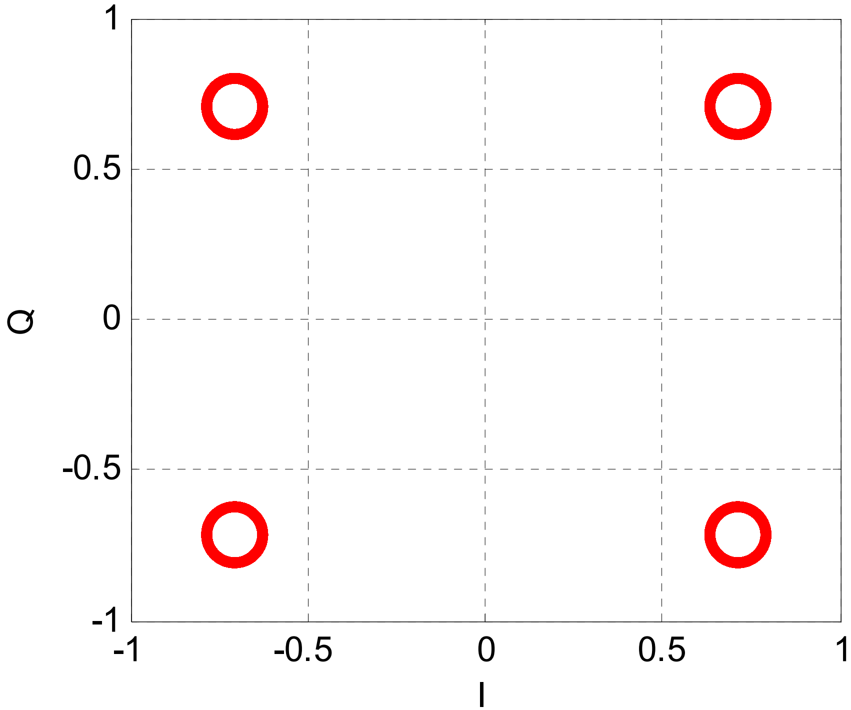


RX baseband

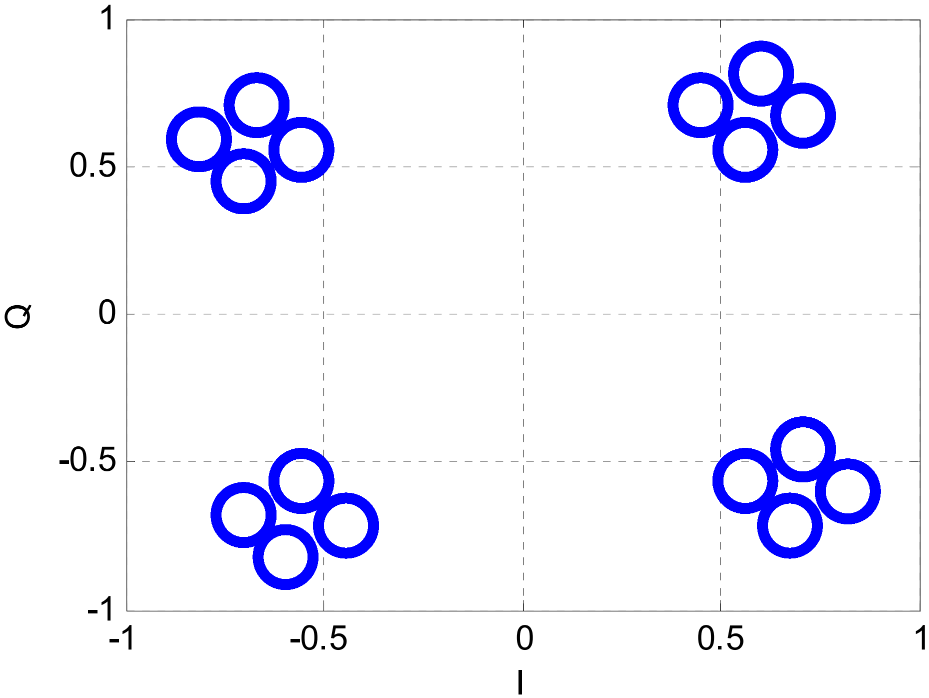
- Influence: limited suppression of mirror
- Mirror signal lies in-band for direct-conversion system
- For analysis: IQ imbalance frequency independent, but varies per front-end

Influence of IQ imbalance (II)

- Influence of moderate IQ imbalance on QPSK modulation

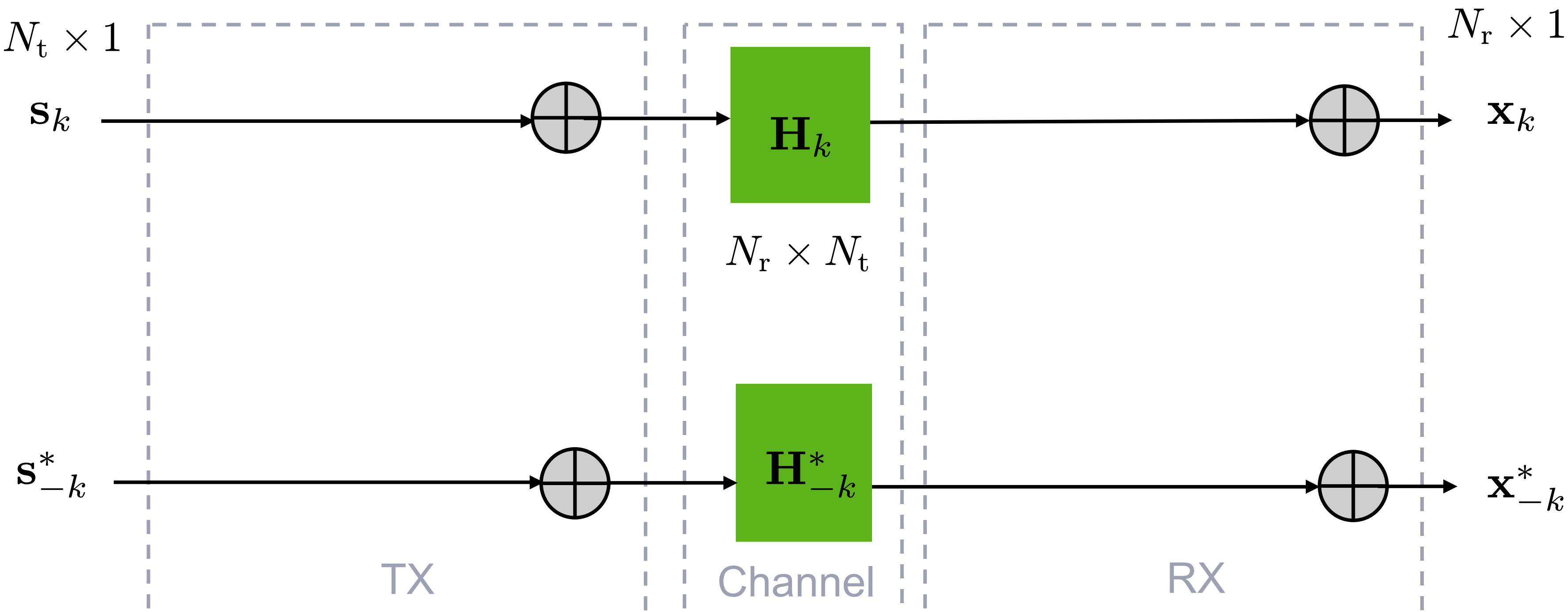


without IQ imbalance



with IQ imbalance

Influence of IQ imbalance (II)



$$\mathbf{x}_k = \mathbf{H}_k \mathbf{s}_k$$

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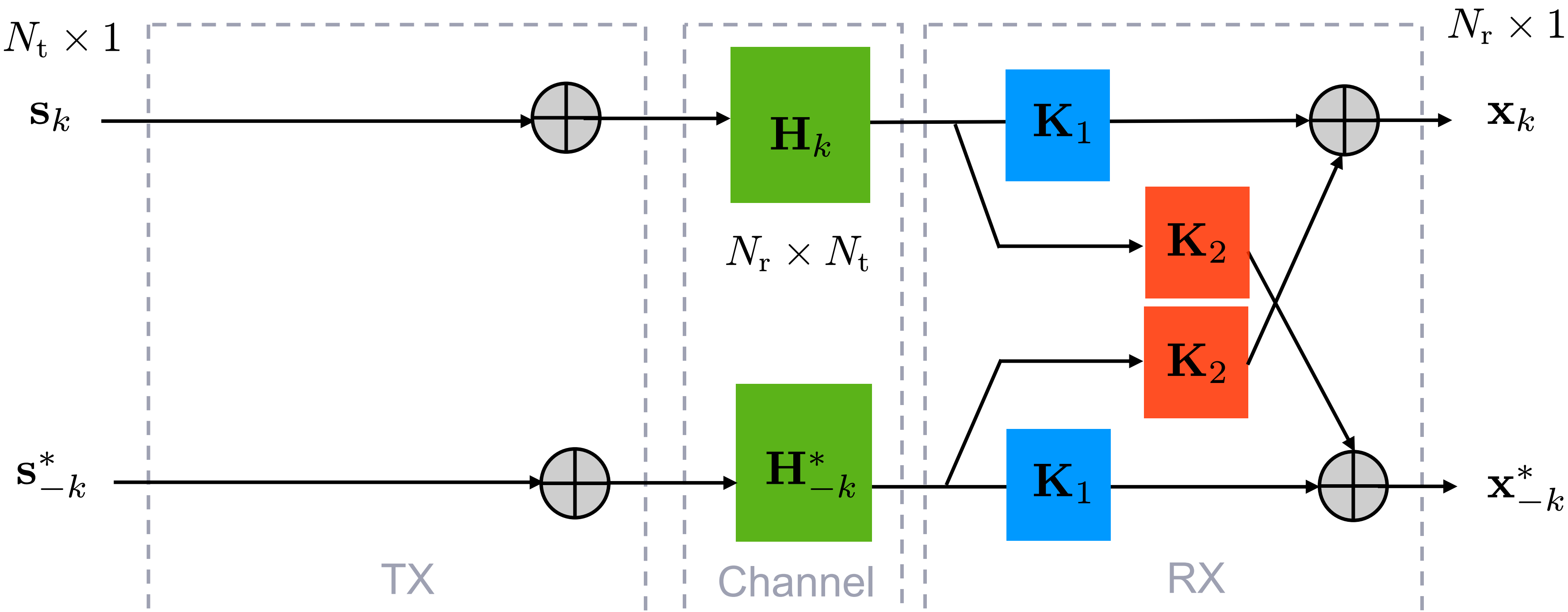
Influence of IQ imbalance (II)



$$\mathbf{x}_k = \mathbf{K}_1 \mathbf{H}_k \mathbf{s}_k + \mathbf{K}_2 \mathbf{H}^*_{-k} \mathbf{s}^*_{-k}$$

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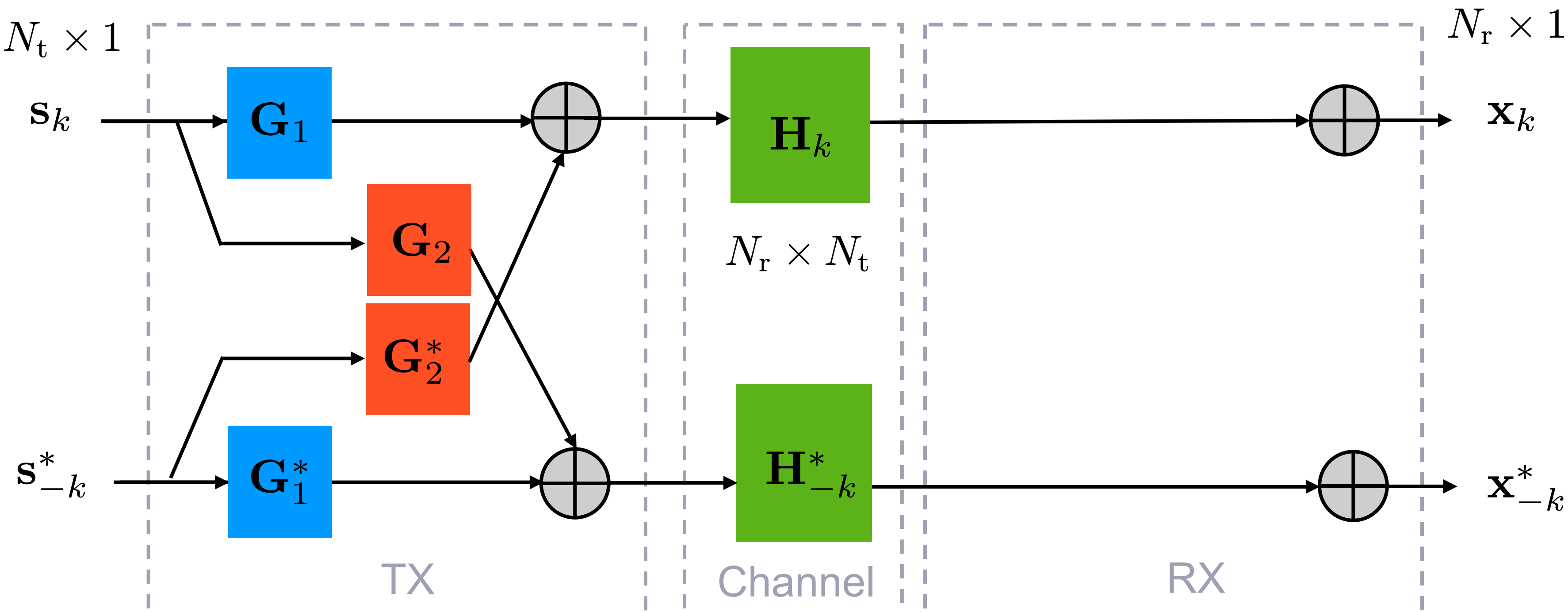
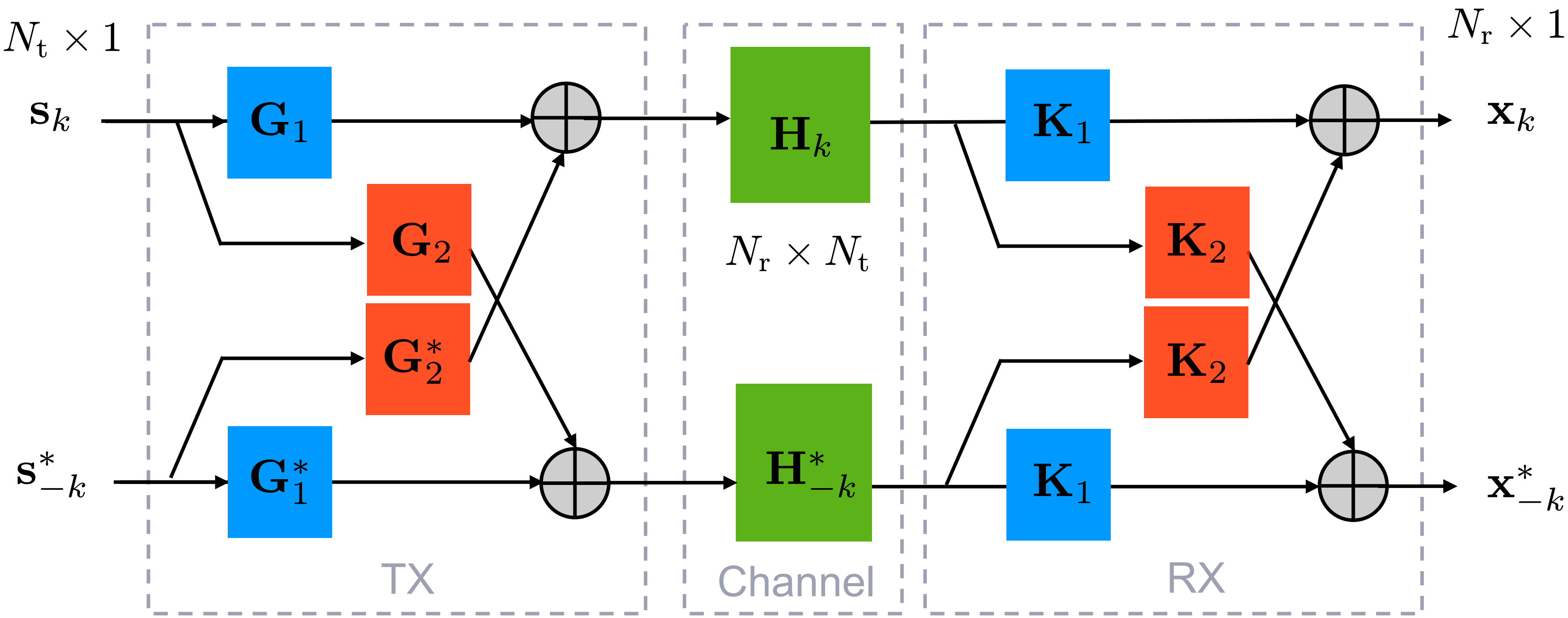
Influence of IQ imbalance (II)



$$\mathbf{x}_k = \mathbf{H}_k \mathbf{G}_1 \mathbf{s}_k + \mathbf{H}_k \mathbf{G}_2^* \mathbf{s}^*_{-k}$$

Influence of IQ imbalance (II)



$$\mathbf{x}_k = (\mathbf{K}_1\mathbf{H}_k\mathbf{G}_1 + \mathbf{K}_2\mathbf{H}^*_{-k}\mathbf{G}_2)\mathbf{s}_k + (\mathbf{K}_2\mathbf{H}^*_{-k}\mathbf{G}_1^* + \mathbf{K}_1\mathbf{H}_k\mathbf{G}_2^*)\mathbf{s}^*_{-k}$$

MIMO processing – Zero Forcing

- We apply zero-forcing MIMO detection, using the estimated MIMO channel matrix
- The estimated channel matrix is given by

$$\tilde{\mathbf{H}}_k = \mathbf{K}_1 \mathbf{H}_k \mathbf{G}_1 + \mathbf{K}_2 \mathbf{H}^*_{-k} \mathbf{G}_2$$

- The estimated TX vector is than given by

$$\tilde{\mathbf{s}}_k = \tilde{\mathbf{H}}^\dagger_k \mathbf{x}_k = \mathbf{s}_k + \varepsilon_k$$

- The error in the estimation is given by

$$\varepsilon_{\mathrm{t},k} = \mathbf{G}_\mathrm{e} \mathbf{s}^*_{-k} + \left(\mathbf{H}_k \mathbf{G}_1\right)^\dagger \mathbf{n}_k \quad \text{where} \quad \mathbf{G}_\mathrm{e} = \mathbf{G}_1^\dagger \mathbf{G}_2^*$$

$$\varepsilon_{\mathrm{r},k} = \mathbf{H}_k^\dagger \mathbf{K}_\mathrm{e} \mathbf{H}^*_{-k} \mathbf{s}^*_{-k} + \mathbf{H}_k^\dagger \left(\mathbf{K}_\mathrm{e} \mathbf{n}^*_{-k} + \mathbf{n}_k\right) \quad \text{where} \quad \mathbf{K}_\mathrm{e} = \mathbf{K}_1^\dagger \mathbf{K}_2$$

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Performance analysis – TX IQ imbalance

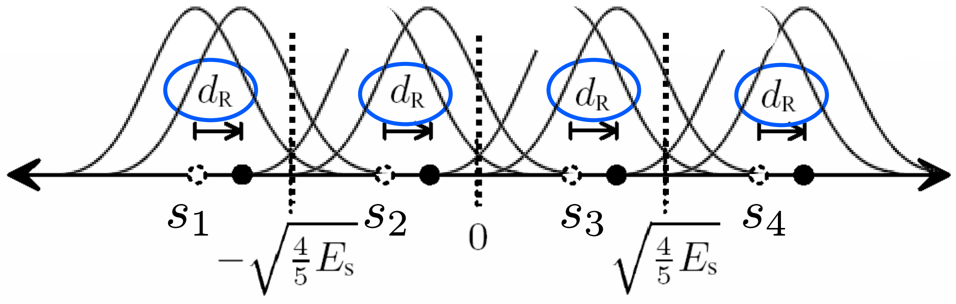
$$\varepsilon_{\mathrm{t},k} = \mathbf{G}_{\mathrm{e}}\mathbf{s}^{*}_{-k} + \left(\mathbf{H}_k\mathbf{G}_1\right)^{\dagger}\mathbf{n}_k$$

Scaled & rotated constellation points

“Regular” noise term

- For a given IQ imbalance and mirror signal, we have a “shift” d_{R} of the constellation.
- If me assume *M*-QAM modulation, we can seperately study the real and imaginairy part of the constellation

$$P^{d}_{\mathrm{e},M\text{-QAM},E_s} = 1 - \left(1 - P^{d_{\mathrm{R}}}_{\mathrm{e},\sqrt{M}\text{-PAM},E_{\mathrm{s}}/2}\right)\left(1 - P^{d_{\mathrm{I}}}_{\mathrm{e},\sqrt{M}\text{-PAM},E_{\mathrm{s}}/2}\right)$$



Performance analysis – RX IQ imbalance

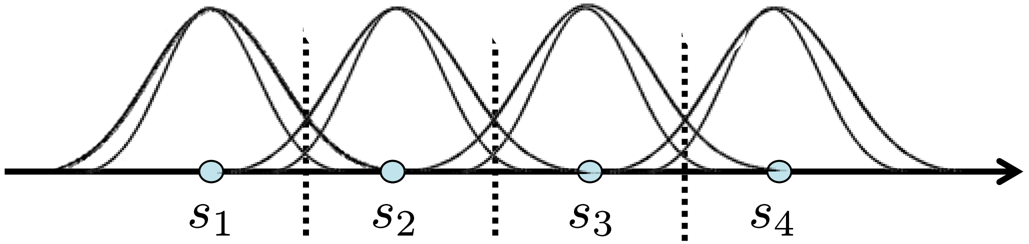
$$\varepsilon_{\mathrm{r},k} = \mathbf{H}_k^\dagger \mathbf{K}_\mathrm{e}\mathbf{H}_{-k}^* \mathbf{s}_{-k}^* + \mathbf{H}_k^\dagger \left(\mathbf{K}_\mathrm{e}\mathbf{n}_{-k}^* + \mathbf{n}_k \right)$$

Leakage from mirror

Extra noise term

“Regular” noise term

- For complex normally distributed channel:
  $\mathbf{H}_k$ and $\mathbf{H}_{-k}$ approximately i.i.d →We can treat the leakage as an extra noise term
- Shown to be valid assumption by [Windisch_ICC2006]
- We can calculate an effective SNR

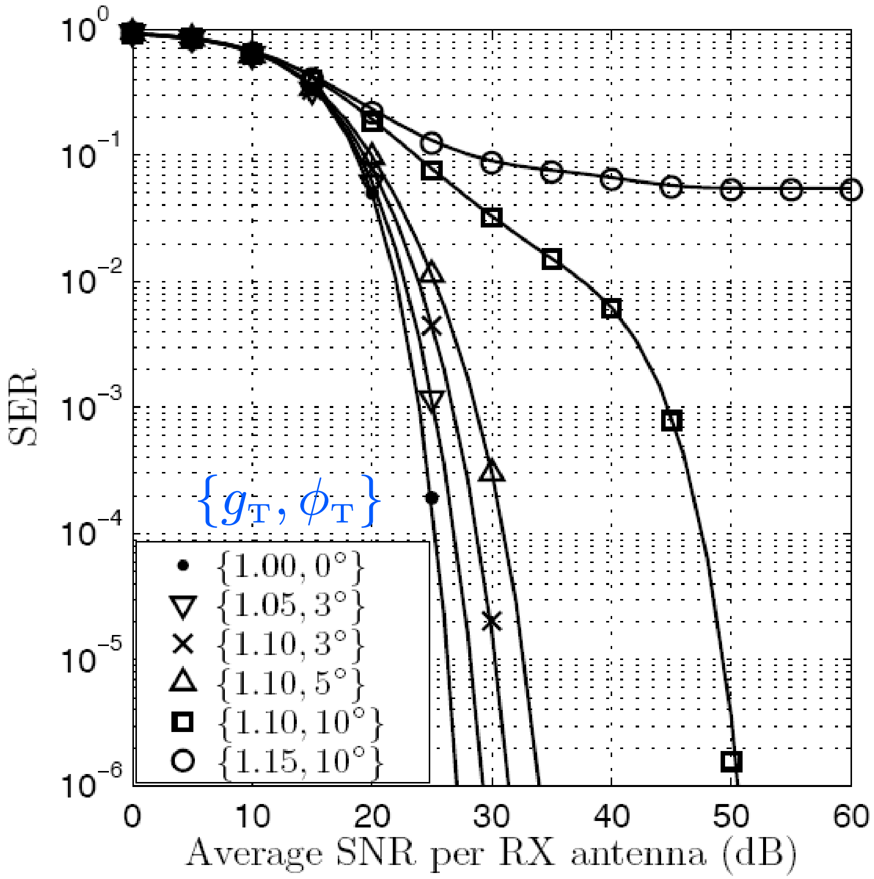


Performance analysis – cont'd

- Symbol-error rate (SER) calculation procedure:
  1) For given SNR calculate influence of shift
  2) Calculation of the effective SNR
  3) Integration over the distribution of the SNR (influence of channel)
- We derived closed form solutions for the SER
  - For TX IQ imbalance
  - For RX IQ imbalance

TX IQ imbalance – Numerical results (I)

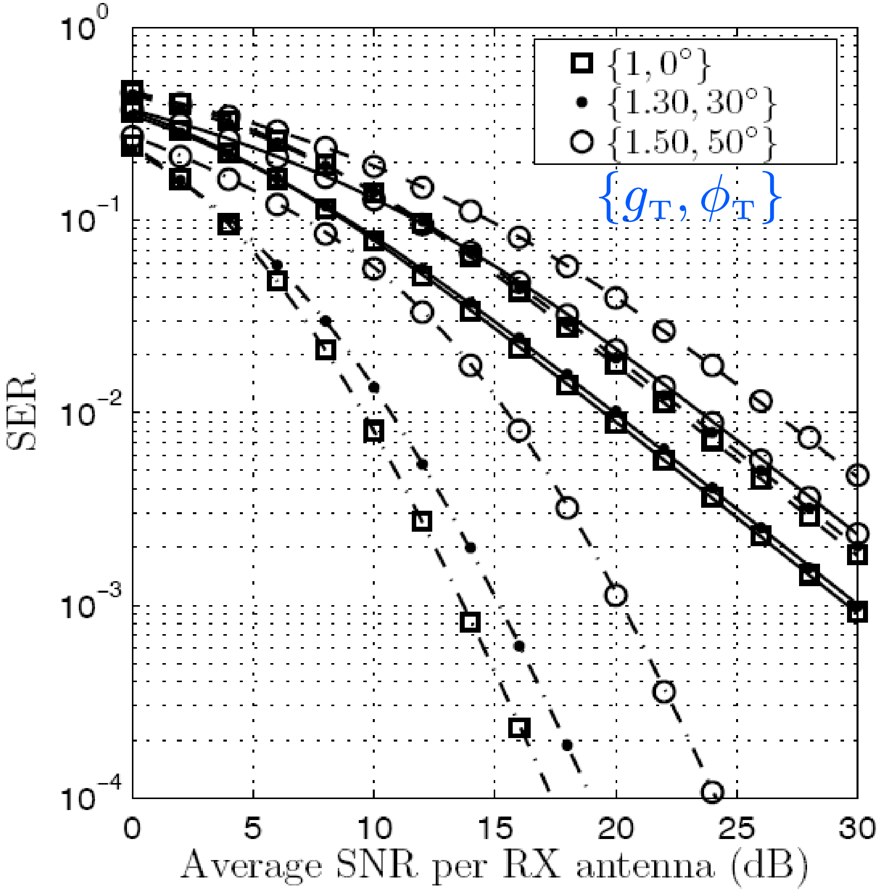
- AWGN channel performance
- Analytical results in lines and simulation results in markers.
- No flooring until "average realization" of constellation is shifted outside the decision region.
- For AWGN channel influence of TX/RX IQ imbalance is similar.



SER performance, TX IQ imbalance and an 1x1 AWGN channel, 64-QAM.

TX IQ imbalance – Numerical results (II)

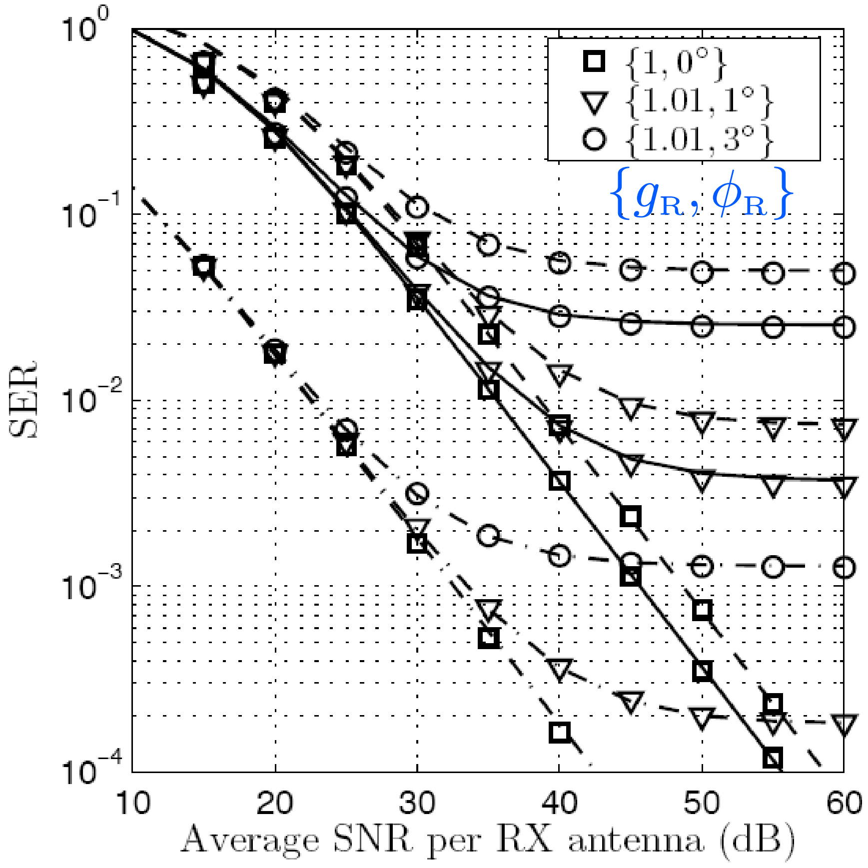
- Rayleigh-faded channel, independent per subcarrier
- 1x1(solid lines), 2x2 (dashed lines) and 2x4 (dash-dot lines)
- Analytical results in lines and simulation results in markers.
- TX IQ results in SNR shift of the SER curve.



SER performance, TX IQ imbalance, Rayleigh faded channel 1x1 (solid lines), 2x2 (dashed lines) and 2x4 (dash-dot lines) 4-QAM system.

RX IQ imbalance – Numerical results

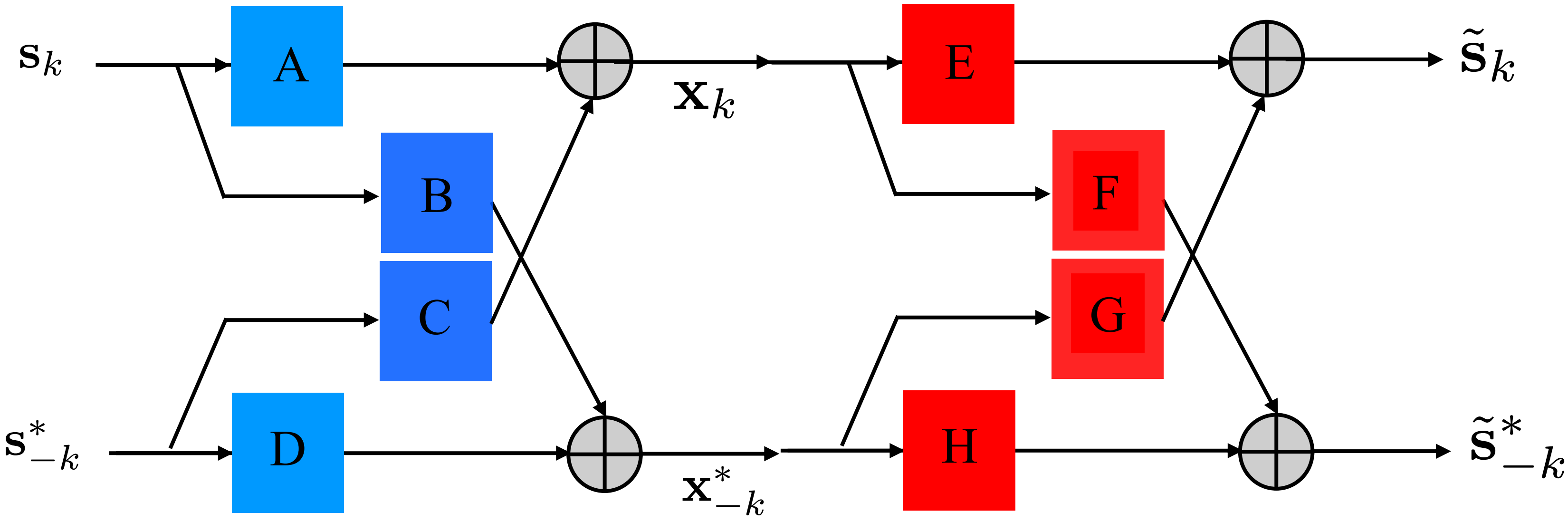
- Rayleigh-faded channel, independent per subcarrier
- 1x1(solid lines), 2x2 (dashed lines) and 2x4 (dash-dot lines)
- Analytical results in lines and simulation results in markers.
- No degradation at low SNR, but flooring at high SNR.
- RX IQ imbalance less destructive (than TX) for low SNRs, and more destructive for high SNRs.



SER performance, RX IQ imbalance, Rayleigh faded channel 1x1 (solid lines), 2x2 (dashed lines) 64-QAM system and 2x2 4-QAM (dash-dot lines).

Compensation IQ imbalance

- Correct for the channel between carrier k and $-k$. Problem goes from $N_r x N_t$ problem to $2N_r x 2N_t$ problem.
- Either estimate the $2N_r x 2N_t$ channel and compensate.
- Or estimate $N_r x N_t$ channel + IQ imbalance parameters, and compensate.



Tim Schenk (tim.schenk@philips.com), June 8, 2007

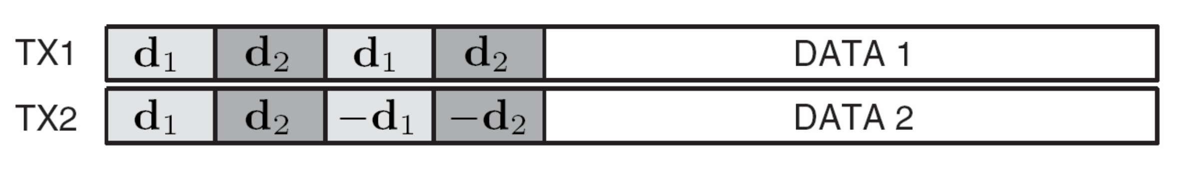
Preamble design

- Preamble used for MIMO channel and IQ imbalance matrices estimation $(\mathbf{H}(k),\mathbf{K}_1,\mathbf{K}_2,\mathbf{G}_1,\mathbf{G}_2)$
  →Approach exploits that imbalance matrices are time- and frequency-invariant
- Or preamble can be applied for estimation of the effective channel
- Orthogonality required:
  - between carrier and mirror
  - between TX branches

$s_k = s_{-k} \rightarrow \mathbf{C}_k^{+}$

$s_k = -s_{-k} \rightarrow \mathbf{C}_k^{-}$

- We propose the use of Hadamard matrices



TX1	d_1	d_2	d_1	d_2	DATA 1
TX2	d_1	d_2	$-d_1$	$-d_2$	DATA 2

	k	-k
$\mathbf{d}_1$	c	c
$\mathbf{d}_2$	c	-c

Estimation TX IQ imbalance

- Effective channel estimates

  $$\mathbf{C}_k^{+} = \mathbf{H}_k$$

  $$\mathbf{C}_k^{-} = \mathbf{H}_k\,(\mathbf{G}_1-\mathbf{G}_2^{*}) \;=\; \mathbf{H}_k\mathbf{g}_\mathbf{T}\exp(j\boldsymbol{\varphi}_\mathbf{T})$$

- IQ imbalance estimates

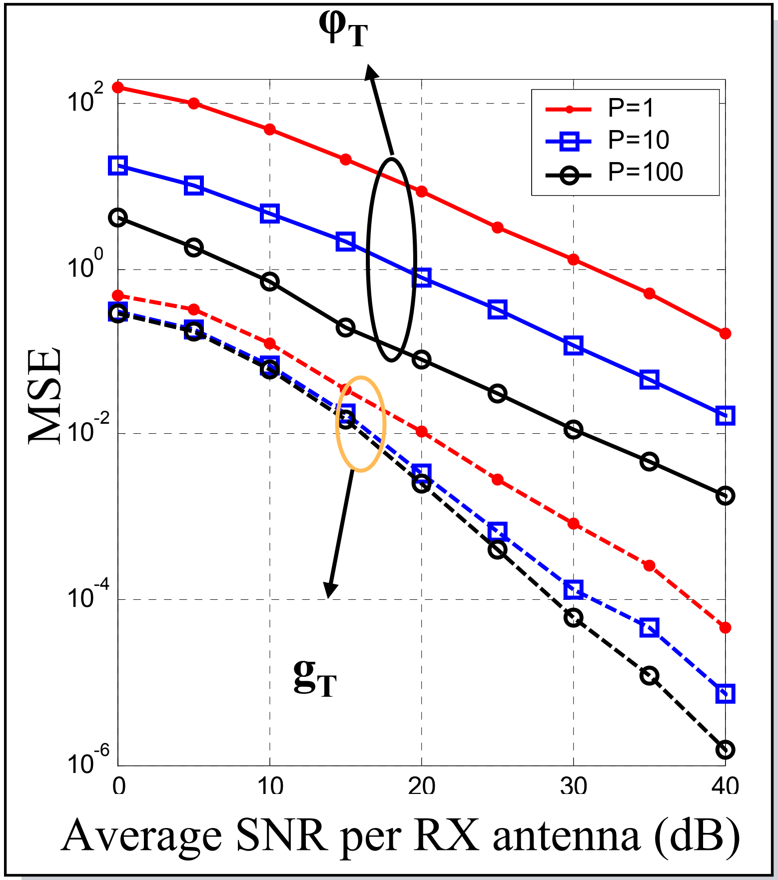
  $$\boldsymbol{\varphi}_\mathbf{T} = \angle\{(\mathbf{C}_k^{+})^{-1}\mathbf{C}_k^{-}\}$$

  $$\mathbf{g}_\mathbf{T} = |(\mathbf{C}_k^{+})^{-1}\mathbf{C}_k^{-}|$$

- Improvement of estimates:
  - averaging over the carriers
  - averaging over *P* packets
- Mean-squared error (MSE) results for 2x2 extension IEEE 802.11a.
- rms delay spread = 50 ns.

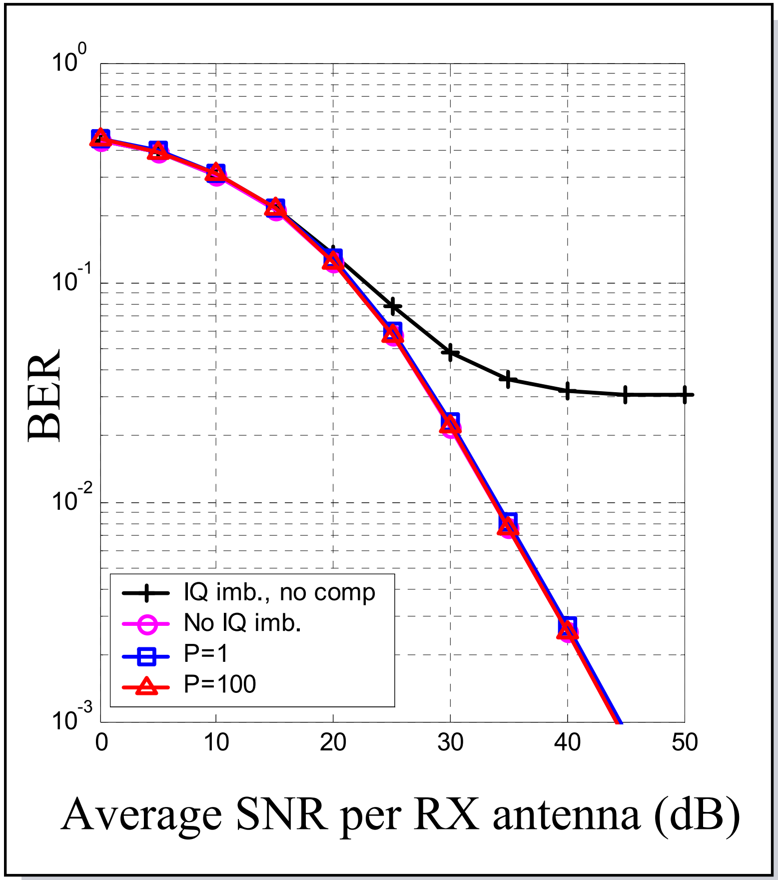
  $$\boldsymbol{\varphi}_\mathrm{T} = \mathrm{diag}\{3^\circ, -3^\circ\}$$

  $$\mathbf{g}_\mathrm{T} = \mathrm{diag}\{1.1,\ 0.9\}.$$



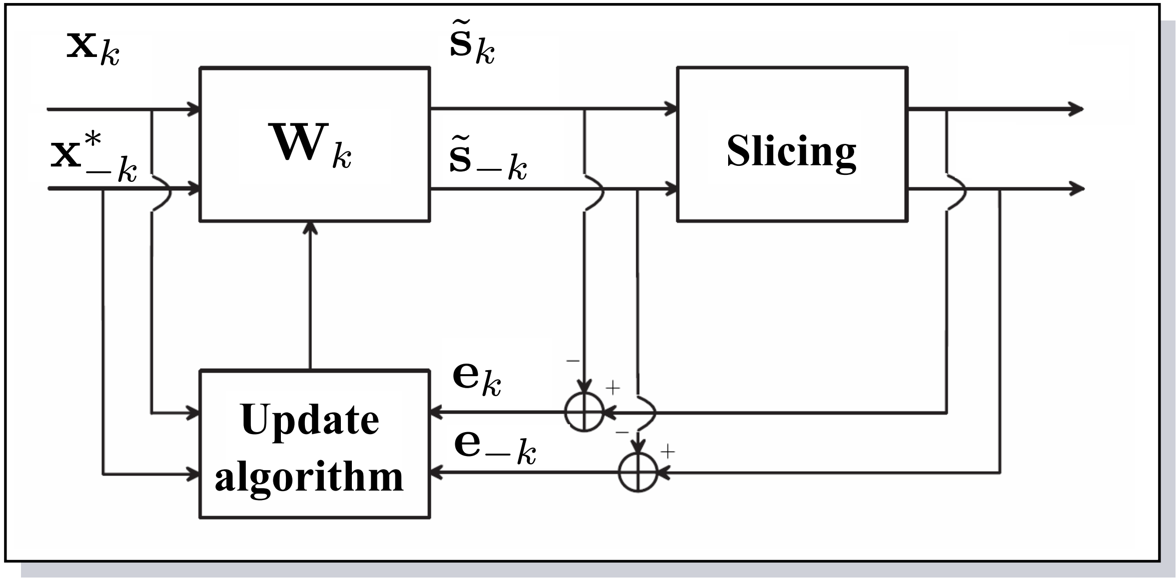
BER RX IQ imbalance

- 2x2 MIMO system applying Zero-Forcing estimation, 64-QAM modulation and no coding.
- $\boldsymbol{\varphi}_R = \mathrm{diag}\{3^\circ, -3^\circ\}$ and $\mathbf{g}_R = \mathrm{diag}\{1.1,\ 0.9\}$. rms delay spread = 50 ns.
- For regarded SNR range perfect compensation of the effect
- Same possible for TX and TX/RX



Adaptive Filter based compensation

- No estimation but adaptive filtering (AF) based compensation
- Especially applicable for frequency dependent IQ imbalance
  → too many pilots required
- LMS/RLS-algorithm used to update of weighting matrix **W**
- Implementation can be limited to outer carriers
- Redetection can be applied to improve convergence behavior



Summary: IQ imbalance

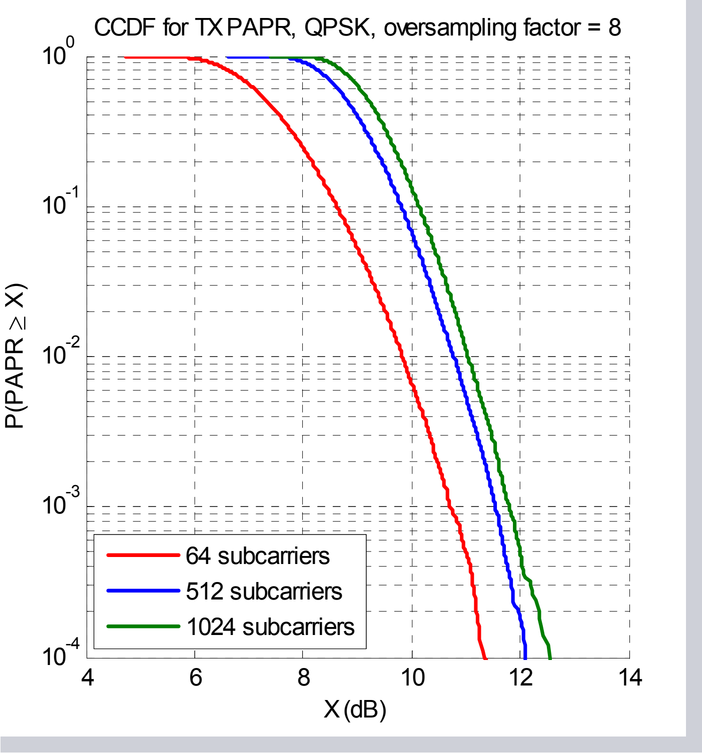
- IQ imbalance in ZIF systems results in leakage between subcarrier and its mirror carrier.
- Two data-aided estimation/compensation approaches are proposed:
  - Using effective channel
  - Estimation of the IQ imbalance and wireless channel parameters
- Latter exploits that imbalance matrices are time- and frequency-invariant
- Frequency dependent IQ imbalance compensated using adaptive MIMO filtering
- Algorithms can **significantly reduce** the influence of IQ mismatch

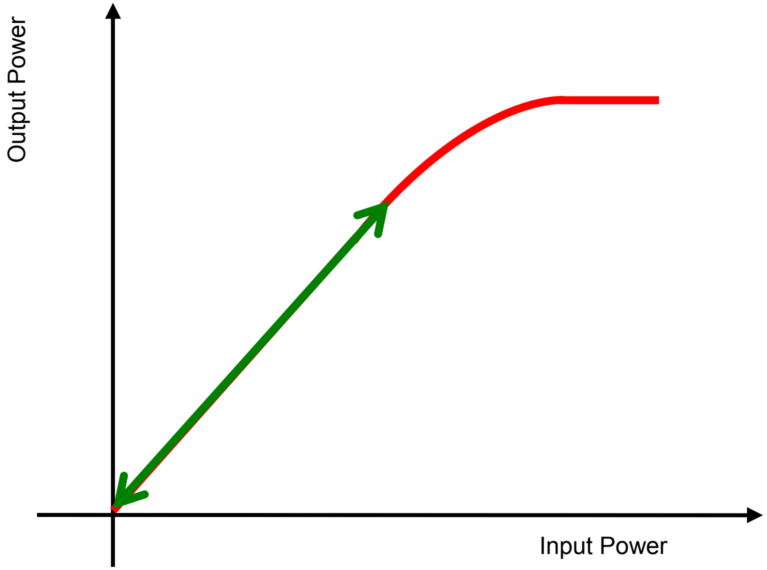
IQ imbalance in (MIMO) OFDM – Additional reading

- M. Valkama, et. Al., “Advanced methods for I/Q imbalance compensation in communication receivers,” *IEEE Trans. on Signal Proc.*, Oct. 2001.
- A. Tarighat, et al., “MIMO OFDM receivers for systems with IQ imbalances,” *IEEE Trans. on Signal Proc.*, Sept. 2005.
- M. Windisch et al., “Standard-independent I/Q imbalance compensation in OFDM direct-conversion receivers,” in *Proc. 9th InOWo Workshop*, Sept. 2004.
- E. Tsui and J. Lin, “Adaptive IQ imbalance correction for OFDM systems with frequency and timing offsets,” in *Proc. IEEE Globecom* 2004
- T.C.W. Schenk, et al., “Estimation and compensation of TX and RX IQ imbalance in OFDM based MIMO systems,” in *Proc. IEEE RWS 2006*, Jan. 2006.
- T.C.W. Schenk, et al., “Performance impact of IQ mismatch in direct-conversion MIMO OFDM transceivers”, *Proc. IEEE RWS 2007*, Jan 2007.

Nonlinearities

- Mainly in PAs and LNAs
- OFDM signals exhibits a high peak-to-average power ratio
- Depends on the number of carriers and modulation depth





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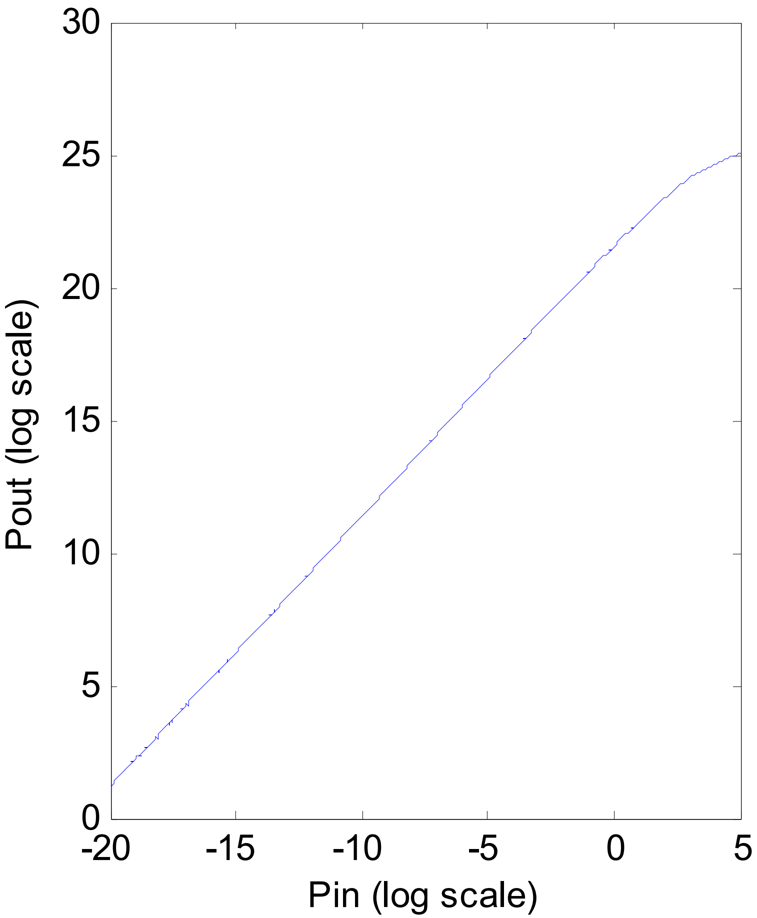
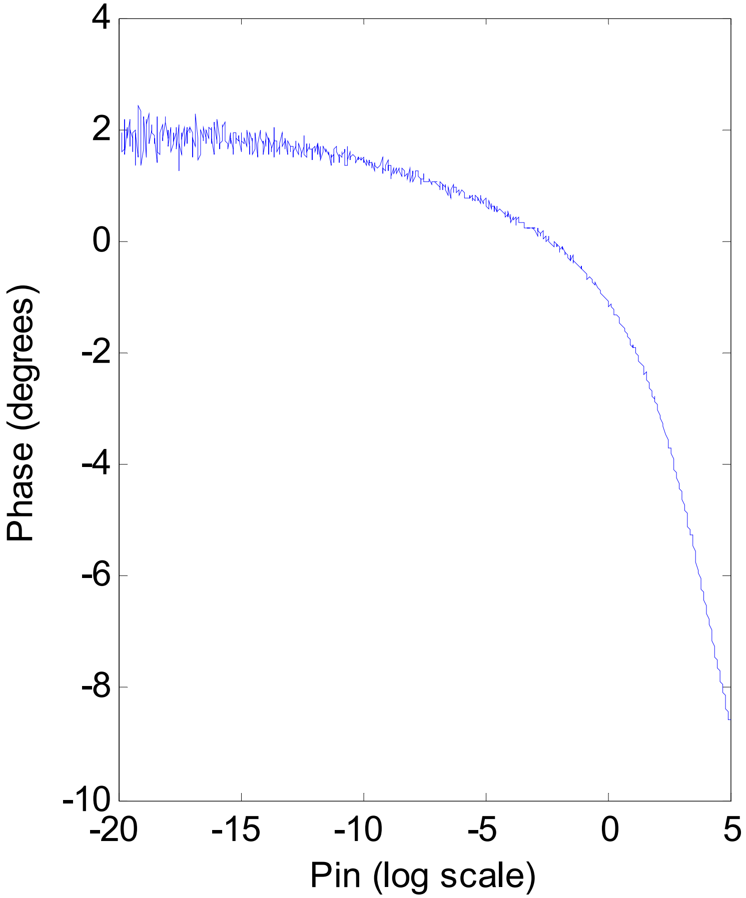
Amplitude and phase distortion

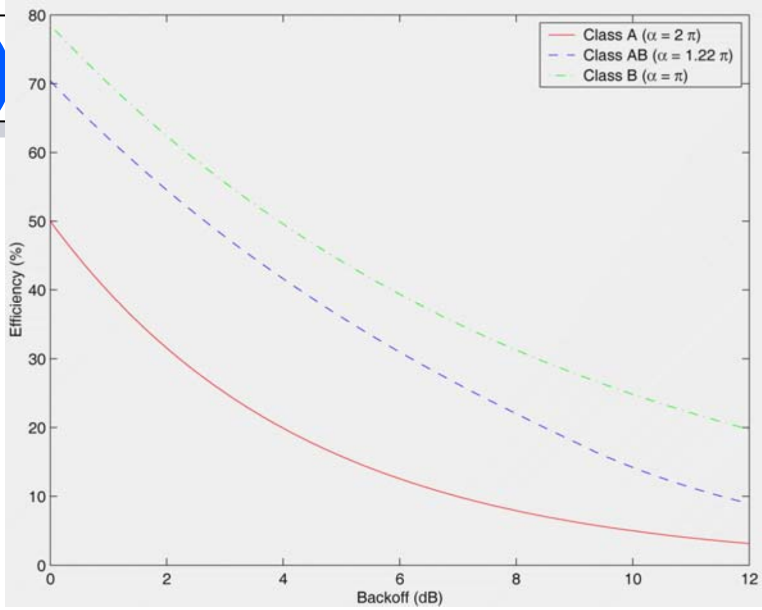


AM-AM

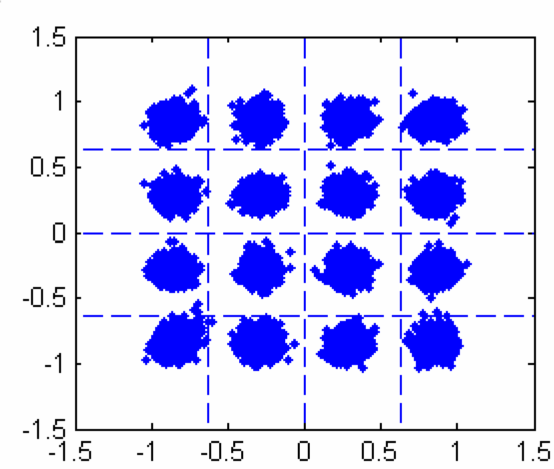


AM-PM

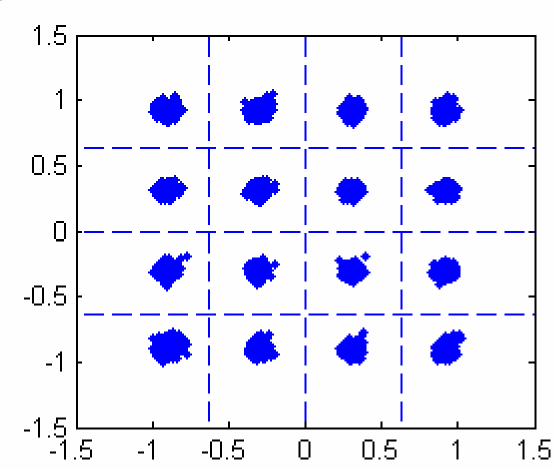
Influence of nonlinearities (I)



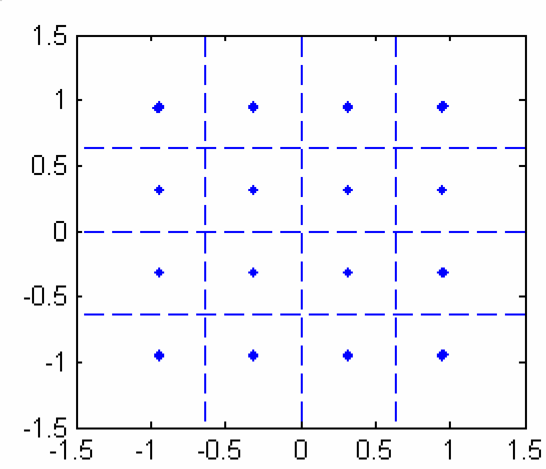
- Increasing the back-off (BO) of the signals on the amplifier
  → very power inefficient



BO = 3 dB



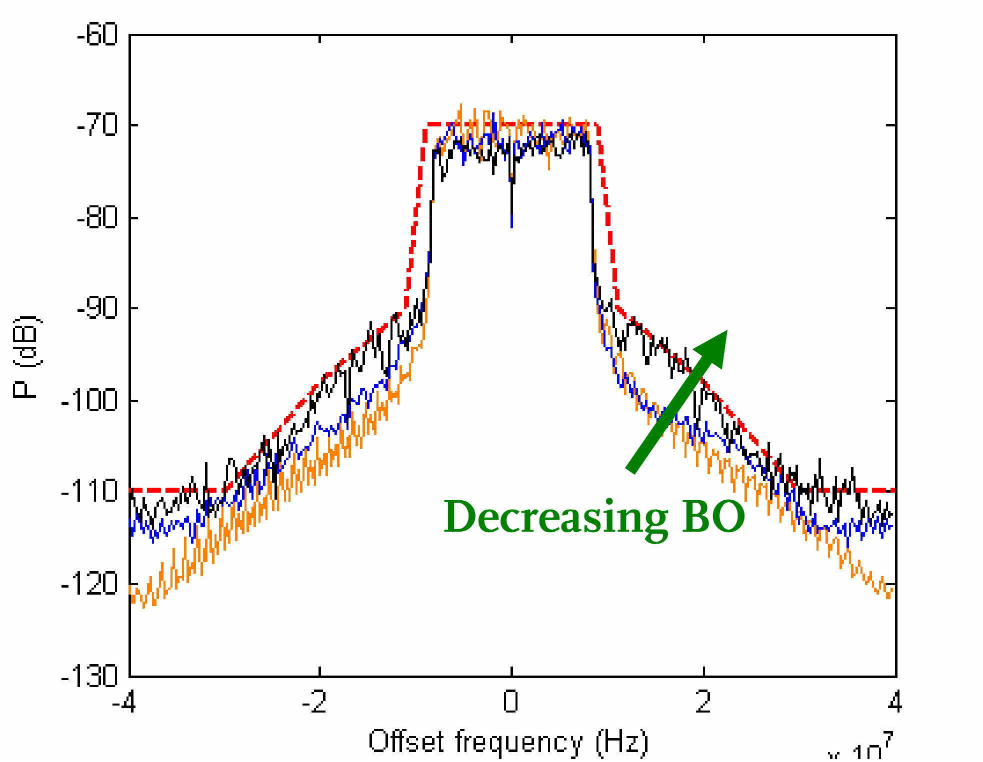
BO = 5 dB



BO = 11 dB

Influence of nonlinearities (II)

- Clipping/Nonlinear distortion results in spectral regrowth.



Signal modelling + MIMO processing

- Gaussian signal input to a nonlinearity *u* [Bussgang52]:

$$u_{\mathrm{d}} = g(u) = \alpha u + d$$

- A MIMO OFDM signal is approximately Gaussian, thus

$$\mathbf{u}_{\mathrm{d}} = g(\mathbf{u}) = (\mathbf{I} \otimes \alpha)\mathbf{u} + \mathbf{d}$$

- For TX nonlinearities we can write the RX signal as

$$\mathbf{x}_{\mathrm{t}} = \mathbf{H}(\mathbf{I}_{N_{\mathrm{c}}} \otimes \alpha_{\mathrm{t}})\mathbf{s} + \mathbf{H}\mathbf{e}_{\mathrm{t}} + \mathbf{n}$$

- For RX nonlinearities we can write the RX signal as

$$\mathbf{x}_{\mathrm{r}} = (\mathbf{I}_{N_{\mathrm{c}}} \otimes \alpha_{\mathrm{r}})\mathbf{H}\mathbf{s} + \mathbf{e}_{\mathrm{R}} + \mathbf{n}$$

- After MIMO processing with perfect CSI we get

$$\begin{cases} \tilde{\mathbf{s}} = \mathbf{H}^{\dagger}\mathbf{x}_{\mathrm{t}} = (\mathbf{I}_{N_{\mathrm{c}}} \otimes \alpha_{\mathrm{t}})\mathbf{s} + \mathbf{e}_{\mathrm{t}} + \mathbf{H}^{\dagger}\mathbf{n} = \hat{\mathbf{s}}_{\mathrm{t}} + \varepsilon_{\mathrm{t}} \\ \tilde{\mathbf{s}} = \mathbf{H}^{\dagger}\mathbf{x}_{\mathrm{r}} = (\mathbf{I}_{N_{\mathrm{c}}} \otimes \alpha_{\mathrm{r}})\mathbf{s} + \mathbf{H}^{\dagger}(\mathbf{e}_{\mathrm{r}} + \mathbf{n}) = \hat{\mathbf{s}}_{\mathrm{r}} + \varepsilon_{\mathrm{r}} \end{cases}$$

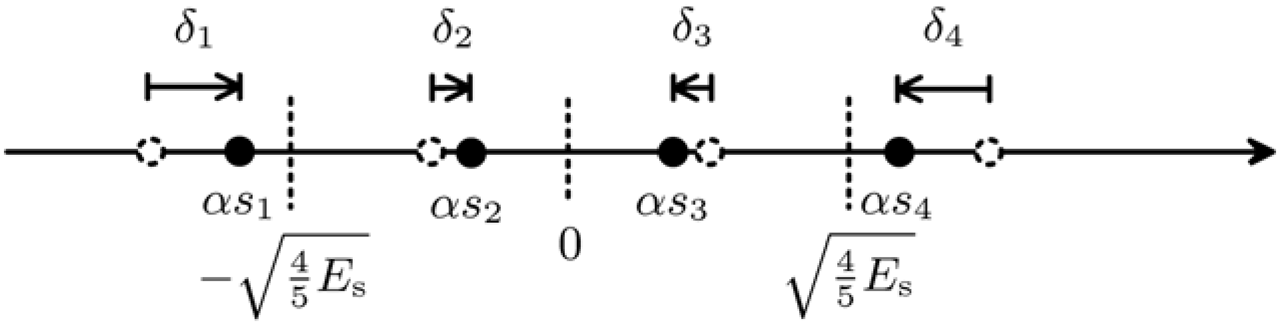
Performance analysis – nonlinearities

$$\begin{cases} \tilde{\mathbf{s}} = \mathbf{H}^{\dagger}\mathbf{x}_{\mathrm{t}} = (\mathbf{I}_{N_{\mathrm{c}}} \otimes \alpha_{\mathrm{t}})\mathbf{s} + \mathbf{e}_{\mathrm{t}} + \mathbf{H}^{\dagger}\mathbf{n} = \hat{\mathbf{s}}_{\mathrm{t}} + \varepsilon_{\mathrm{t}} \\ \tilde{\mathbf{s}} = \mathbf{H}^{\dagger}\mathbf{x}_{\mathrm{r}} = (\mathbf{I}_{N_{\mathrm{c}}} \otimes \alpha_{\mathrm{r}})\mathbf{s} + \mathbf{H}^{\dagger}(\mathbf{e}_{\mathrm{r}} + \mathbf{n}) = \hat{\mathbf{s}}_{\mathrm{r}} + \varepsilon_{\mathrm{r}} \end{cases}$$

scaling

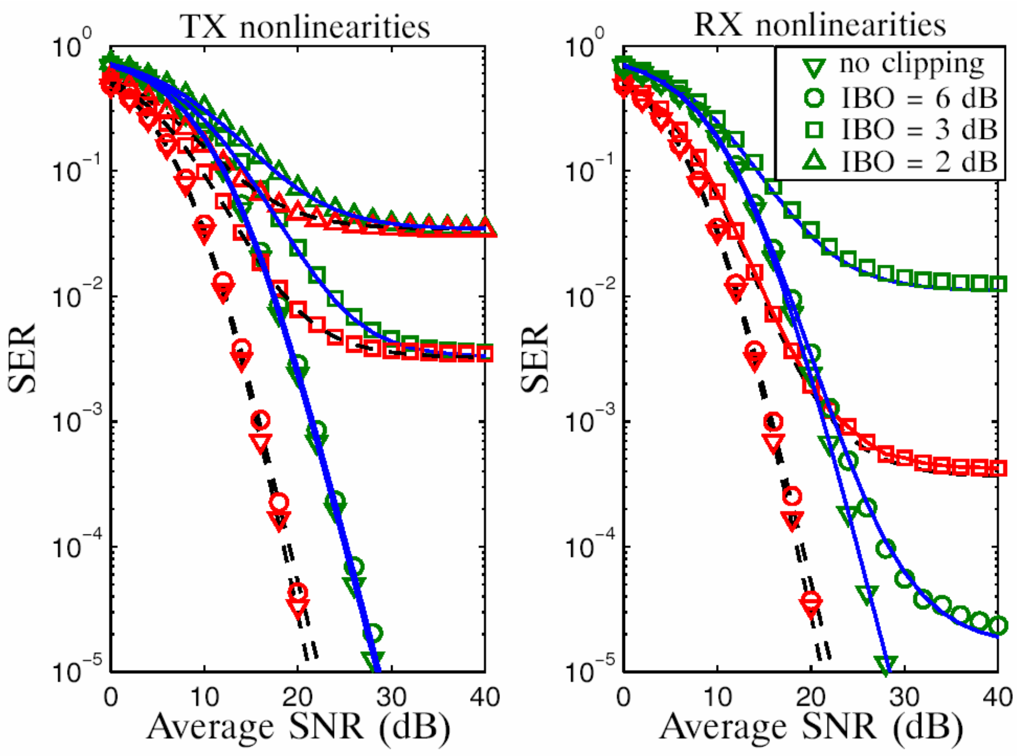
extra noise term

“regular” noise term



- Bit-error rate calculation procedure:
  1) For given SNR calculate influence of scaling (influence *M*-QAM)
  2) Calculation of the effective SNR
  3) Integration over the distribution of the SNR (influence of channel).

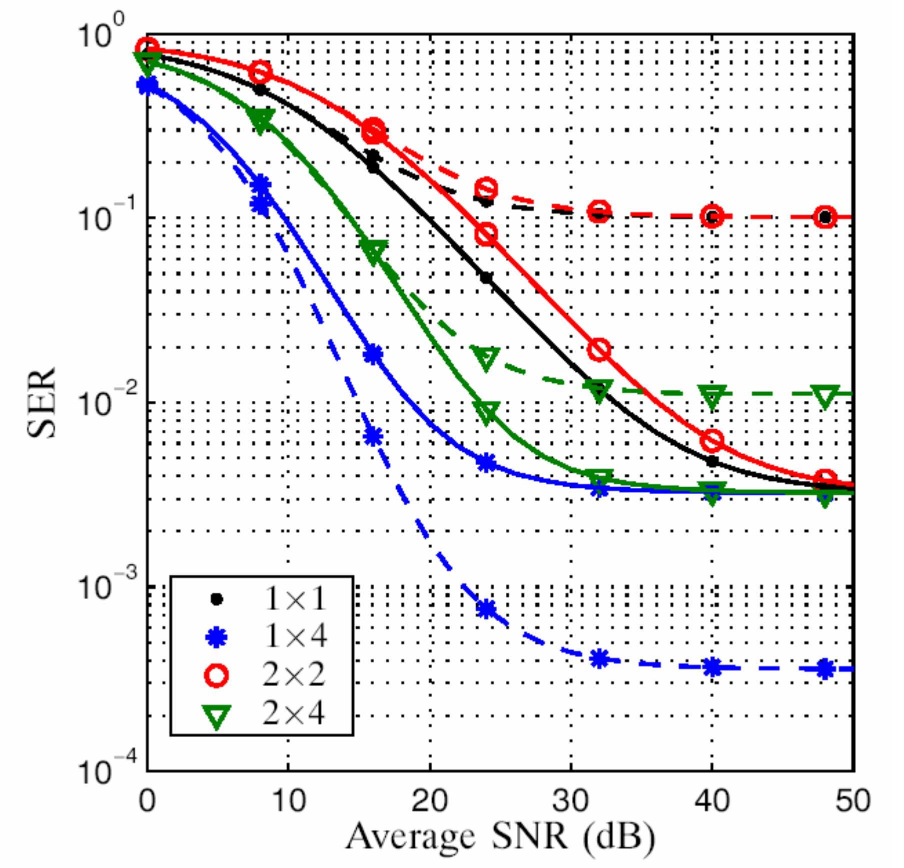
Numerical results (I)



SER performance for Rayleigh channel.
1x4 (dashed) and 2x4 (solid) 16-QAM systems

- Rayleigh faded channel
- Clipping amplifier
- Solid lines are analytical results and markers are from simulations.
- 1x4 results in **black** and 2x4 results in **blue**
- Flooring TX nonlinearities independent of MIMO configuration.
- RX nonlinearities benefit from spatial diversity.

Numerical results (II)



SER performance for Rayleigh channel. RX nonlinearities (dashed) and TX nonlinearities (solid) 16-QAM systems

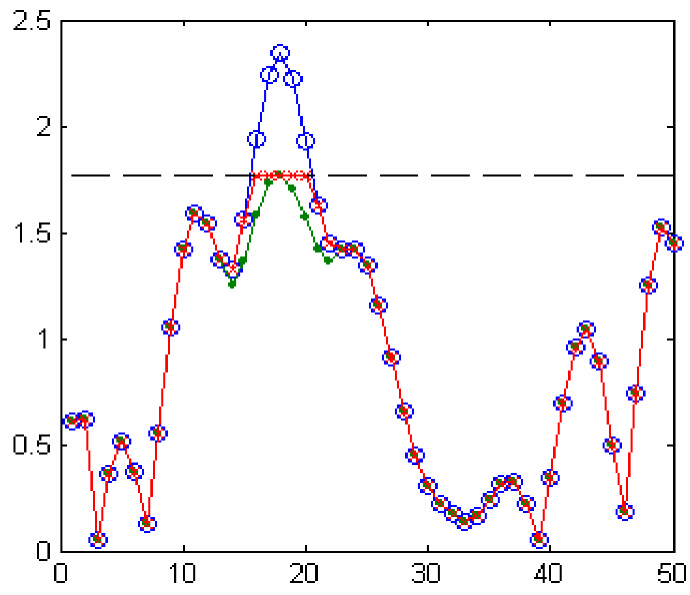
- Analytical results for both TX nonlin. (solid lines) and
  RX nonlin. (dashed lines).
- Clipping amplifier with 3dB input backoff
- Different MIMO configurations.
- Impact depends on MIMO configuration.

Nonlinearities – Mitigation (I)

Example on next slides

1) Decreasing the PAPR of the OFDM signals

- Block coding with Fourier like (Golay codes)
  - Error correcting capabilities less optimal
- Selective mapping → Spatial shifting
  - Significant reduction
  - Computational complex
  - Introduces small overhead
- "Smart" clipping
  - Reduces spectral regrowth
  - Introduces extra distortion in the band
- Constellation shaping
- ……



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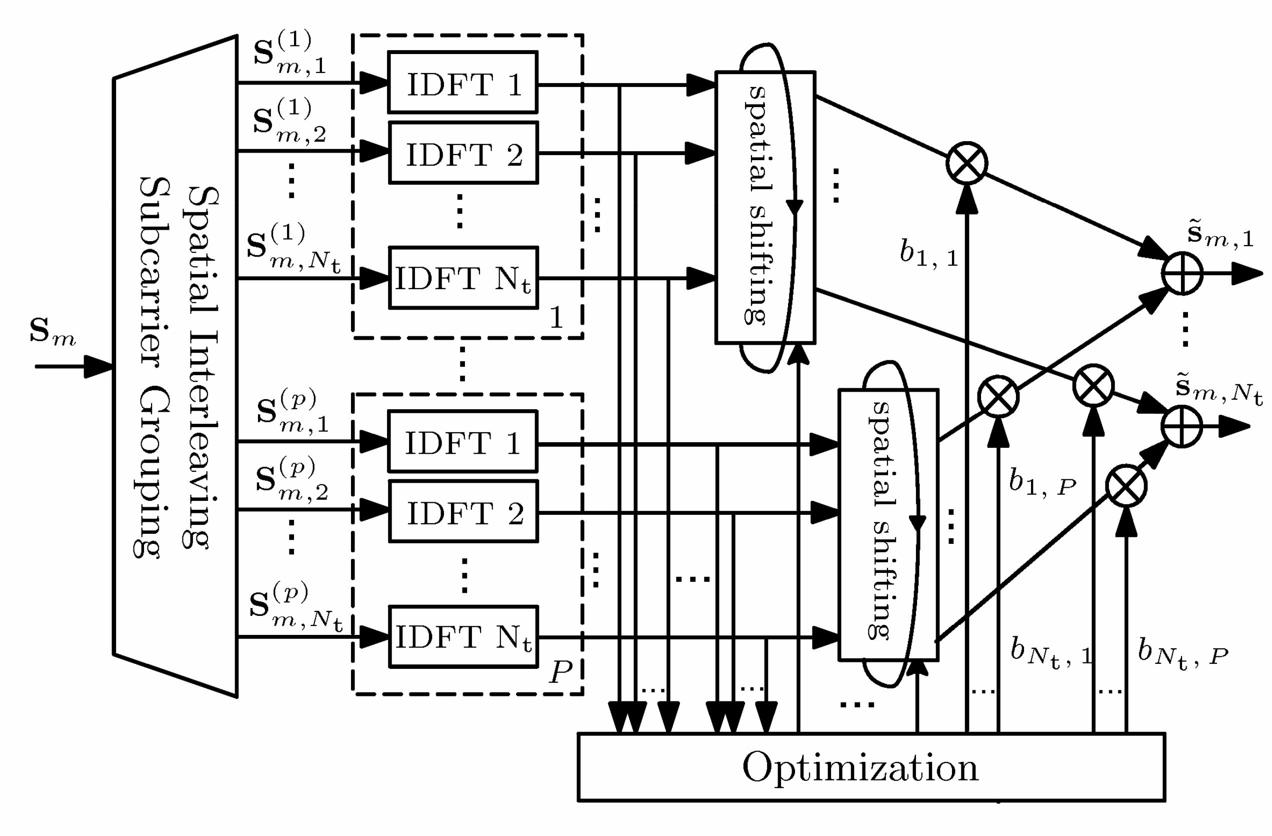
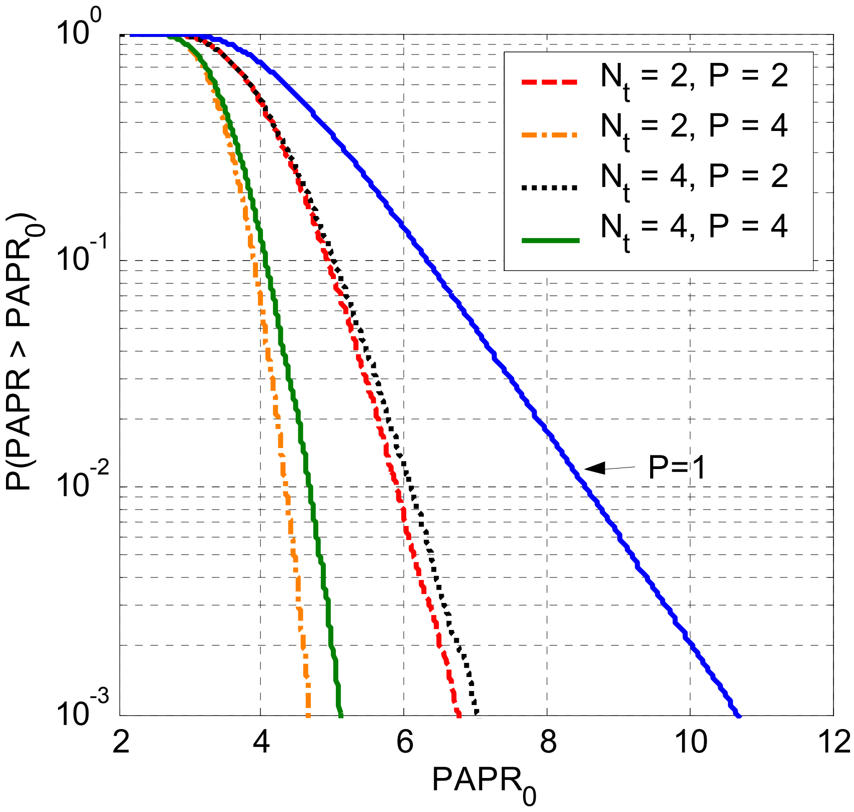
Spatial shifting + phase shifting



- SS: Reshuffeling of groups of subcarriers between TX branches.
- PS: Sum the groups with different phases
- Notify RX of chosen phases and shift → extra overhead

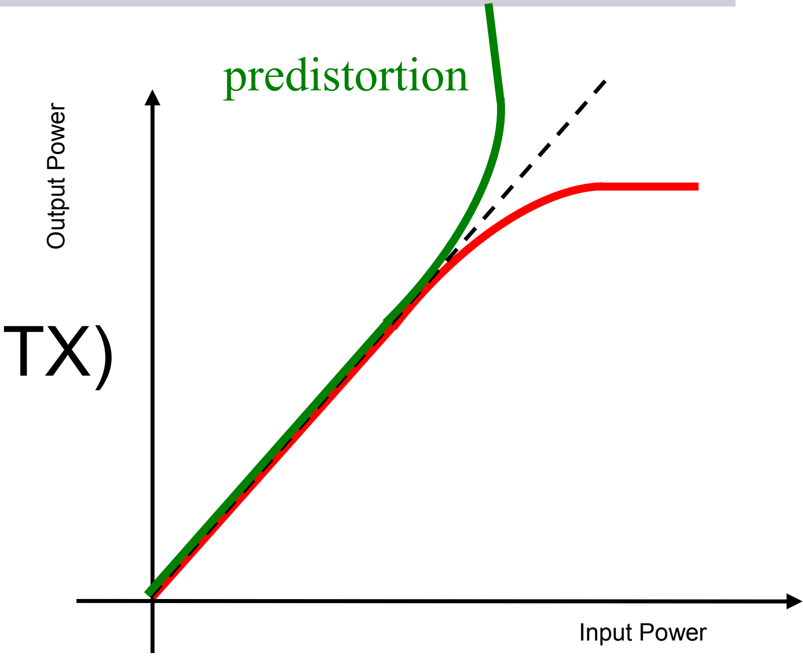
Performance of SS/PS

- Number of subcarrier groups = *P.*
- With increasing *P* performance increases, however overhead is also increased.
- No degradation in performance when SS/PS is transmitted correctly.



Nonlinearities – Mitigation (II)

II) Compensation for nonlinearities



- Digital baseband pre-distortion (TX)
  - Nonlinear transfer found in calibration procedure
  - Gaining popularity
  - Does not solve the power limiting

- Recovering techniques (in the baseband RX)
  - Interference term is function of the data
  - Can be computational complex
  - How do we get the nonlinear transfer of the total system?

Example on next slides

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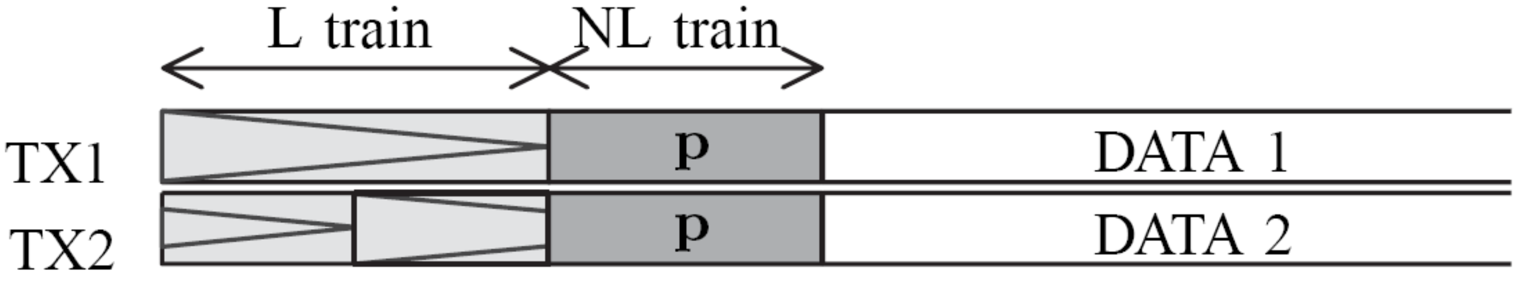
Estimation using preamble

- Constant modulus part “L Train” for estimation MIMO channel.
- High PAPR part “NL Train” for estimation multiple nonlinear transfers.



Linear Channel estimation

$$\mathbf{x}_{\mathrm{p}} = \overbrace{(\mathbf{F}\boldsymbol{\Upsilon} \otimes \mathbf{I}_{N_{\mathrm{R}}})}^{\text{RX OFDM proc.}}\overbrace{\mathbf{K}}^{\text{CIR}}\overbrace{g}^{\text{NL}}\big(\overbrace{(\boldsymbol{\Theta} \otimes \mathbf{I}_{N_{\mathrm{T}}})}^{\text{CP addition}}\overbrace{\mathbf{u}_{\mathrm{p}}}^{\text{training symbol}}\big)$$

$$= (\mathbf{F}\boldsymbol{\Upsilon} \otimes \mathbf{I}_{N_{\mathrm{R}}})\mathbf{K}(\boldsymbol{\Theta} \otimes \mathbf{I}_{N_{\mathrm{T}}})(\mathbf{I}_{N_{\mathrm{C}}} \otimes \boldsymbol{\eta})\mathbf{u}_{\mathrm{p}},$$

- Due to constant modulus property channel can be estimated upto constant diagonal matrix

$$\hat{\mathbf{H}} = \mathbf{H}(\mathbf{I}_{N_{\mathrm{C}}} \otimes \boldsymbol{\eta}),$$

- Constant can be estimated from nonlinearity and corrected for or corrected as part of nonlinearity.

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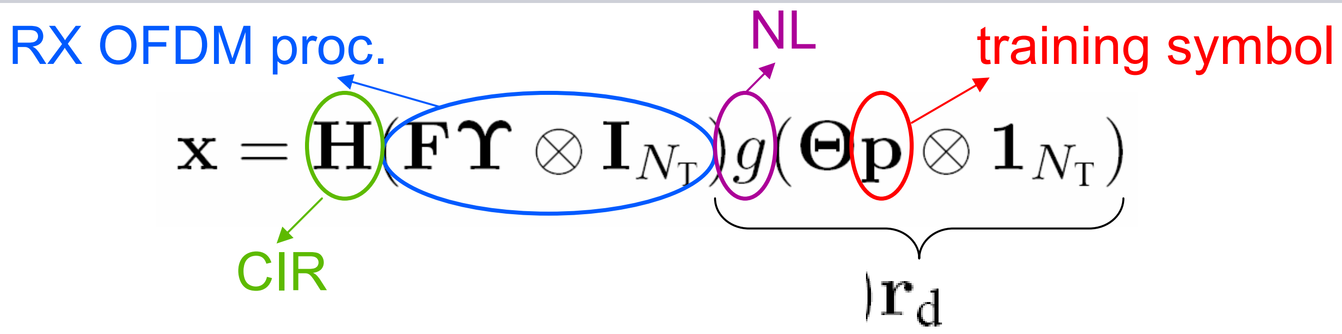
Nonlinearity estimation


RX OFDM proc.

NL

training symbol

$$\mathbf{x} = \mathbf{H}(\mathbf{F}\boldsymbol{\Upsilon} \otimes \mathbf{I}_{N_\mathrm{T}})\underbrace{g(\boldsymbol{\Theta}\mathbf{p} \otimes \mathbf{1}_{N_\mathrm{T}})}_{\mathbf{r}_\mathrm{d}}$$

CIR

- Estimated nonlinear distorted sequence is given by:

$$\hat{\mathbf{r}}_\mathrm{d} = (\mathbf{I}_{N_\mathrm{C}} \otimes \boldsymbol{\eta}^{-1})\mathbf{r}_\mathrm{d}.$$

- Which can be written in polynomial form as

$$\hat{\mathbf{r}}_{\mathrm{d},n_\mathrm{T}} = \eta_{n_\mathrm{T}}^{-1}\mathbf{r}_{\mathrm{d},n_\mathrm{T}} = \eta_{n_\mathrm{T}}^{-1} g_{n_\mathrm{T}}(\mathbf{p}).$$

$$\hat{r}_{\mathrm{d},n_\mathrm{T}}(n) = \eta_{n_\mathrm{T}}^{-1} p(n) \sum_{m=0}^{N-1} \beta_{m+1,n_\mathrm{T}} |p(n)|^m,$$

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Nonlinearity estimation (cont'd)

$$\hat{r}_{\mathrm{d},n_{\mathrm{T}}}(n) = \eta_{n_{\mathrm{T}}}^{-1} p(n) \sum_{m=0}^{N-1} \beta_{m+1,n_{\mathrm{T}}} |p(n)|^{m},$$

- This can be written in matrix notation as:

(i,k)th element $\mathbf{\Phi}$

$$\hat{\mathbf{r}}_{\mathrm{d},n_{\mathrm{T}}} = \eta_{n_{\mathrm{T}}}^{-1} \mathbf{\Phi} \boldsymbol{\beta}_{n_{\mathrm{T}}} = \mathbf{\Phi} \boldsymbol{\beta}'_{n_{\mathrm{T}}},$$

$$p(i)|p(i)|^{k-1}$$

- The estimated parameters of the nonlinearity are then given by

$$\hat{\boldsymbol{\beta}}'_{n_{\mathrm{T}}} = \boxed{\mathbf{\Phi}}^{\dagger} \hat{\mathbf{r}}_{\mathrm{d},n_{\mathrm{T}}},$$

- Using that linear part are in the MIMO channel estimate, we find:

$$\begin{aligned}
\hat{\eta}_{n_{\mathrm{T}}} &= \left(\hat{\beta}'_{n_{\mathrm{T}}}(1)\right)^{-1}, \\
\hat{\beta}_{3,n_{\mathrm{T}}} &= \hat{\eta}_{n_{\mathrm{T}}} \hat{\beta}'_{n_{\mathrm{T}}}(2), \\
\hat{\beta}_{5,n_{\mathrm{T}}} &= \hat{\eta}_{n_{\mathrm{T}}} \hat{\beta}'_{n_{\mathrm{T}}}(3),
\end{aligned}$$

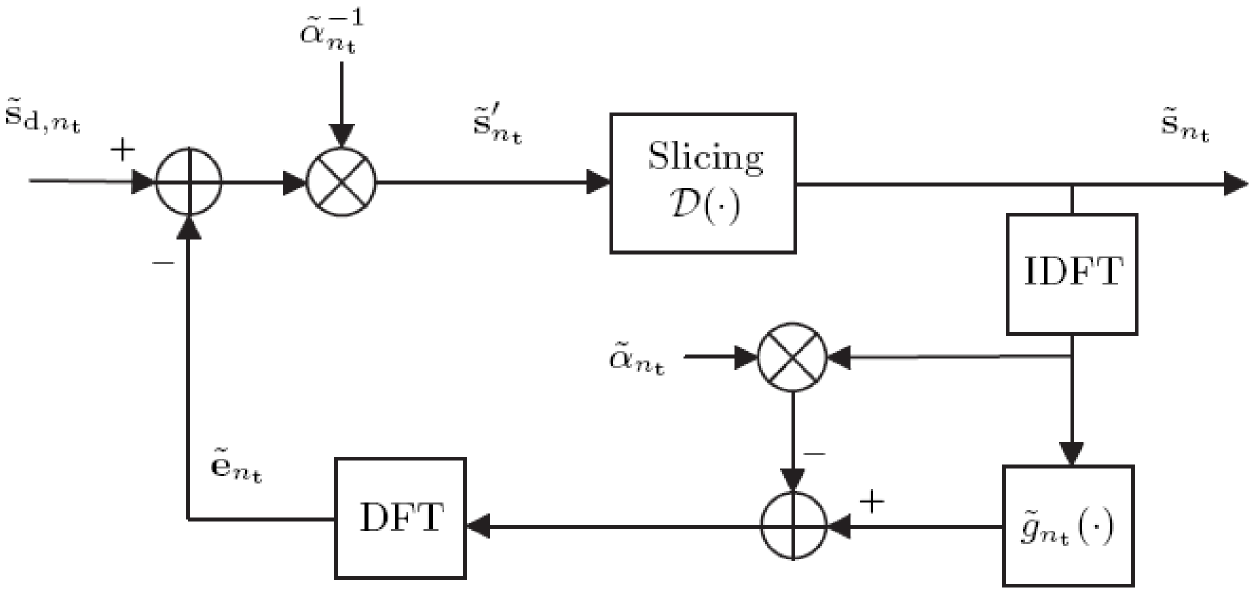
Compensation approaches

1) “Multiple” received signal with estimated inverse nonlinearity

2) Project nonlinearity on new basis. Then interpolation of received signal

3) Iterative distortion removal (IDR)

Iterative distortion removal



- Initial detection
- Estimate distrotion term
- Substract distortion term
- Apply a redetection
- Iterate

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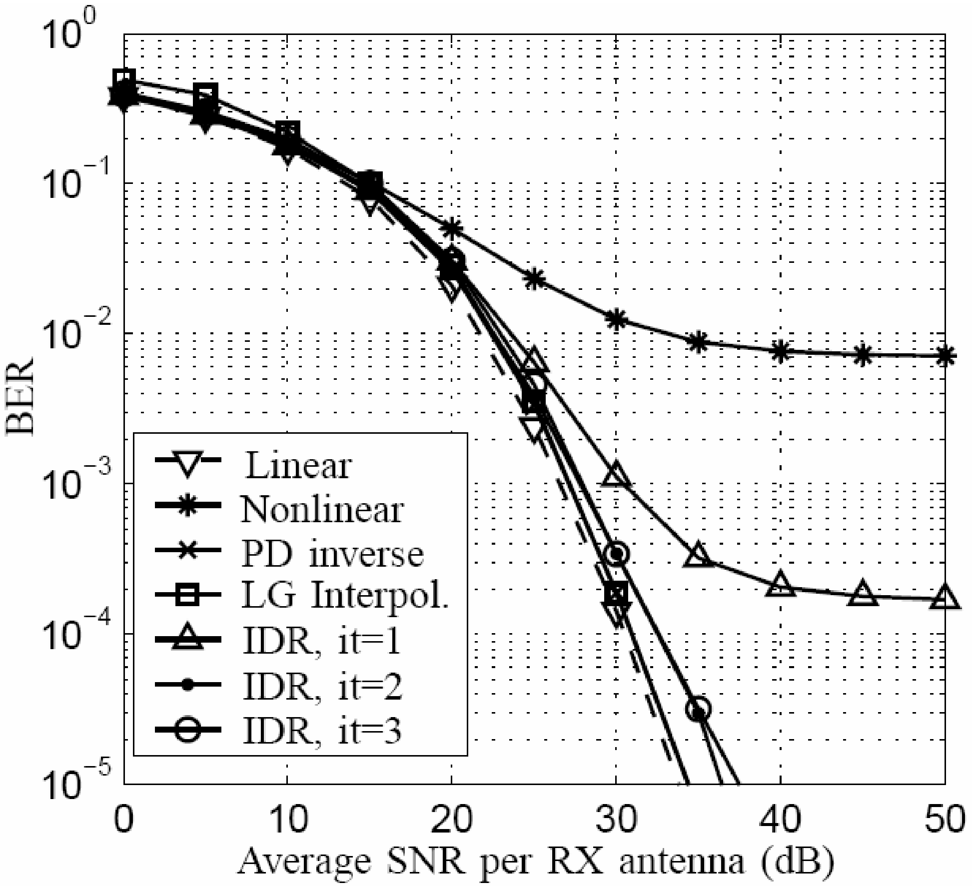
Numerical results (cont’d)

- 2x4, uncoded, 64-QAM
- Linear and nonlinear as reference
- Estimated MIMO channel
- postdistortion (PD) with estimated inverse
- Lagrange interpolation using estimated nonlinearity (5 point basis)
- 3 iterateterations for IDR



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Summary – nonlinearities

- Nonlinearities create decrease SNR and constelation scaling in OFDM systems
- Two major approaches:
  - Reduce PAPR of signals
  - Compensate for (influence) nonlinearities
- Some compensation approaches have been presented, which can significantly reduce the influence of the nonlinearities.

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Nonlinearities in (MIMO) OFDM – Additional reading

- E. Costa, et al., "Impact of amplifier nonlinearities on OFDM transmission system performance," *IEEE Commun. Letters*, Feb. 1999.
- D. Dardari, et al., "A theoretical characterization of nonlinear distortion effects in OFDM systems," *IEEE Trans. on Commun.*, Oct. 2000.
- H. W. Kang, et al., "On compensating nonlinear distortions of an OFDM system using an efficient adaptive predistorter," *IEEE Trans. on Commun.*, April 1999.
- H. Chen, et al., "Iterative estimation and cancellation of clipping noise for OFDM signals," *IEEE Commun. Letters*, July 2003.
- J. Tellado, et al., "Maximum-likelihood detection of nonlinearly distorted multicarrier symbols by iterative decoding," *IEEE Trans. on Commun.*, Feb. 2003.
- T.C.W. Schenk, et al., "Receiver-based compensation of transmitter-incurred nonlinear distortion in multiple-antenna OFDM systems," IEEE VTC 2007.
- T.C.W. Schenk, et al., "Impact of nonlinearities in multiple-antenna OFDM tranceivers," IEEE SCVT2006.

Some things to take home…

- MIMO+OFDM → very suitable for high datarate communications. Basis of many wireless standards.
- OFDM is very sensitive to imperfections of the RF front-end.
- Resulting errors are **not** yet extra noise sources
  → very distinct behavior.
- Mitigation techniques are promising to decrease to requirements on the RF front-end.
- For optimal usage, however, mutual understanding of baseband and RF-front-end limitations is **essential** → co-design

Thank you!

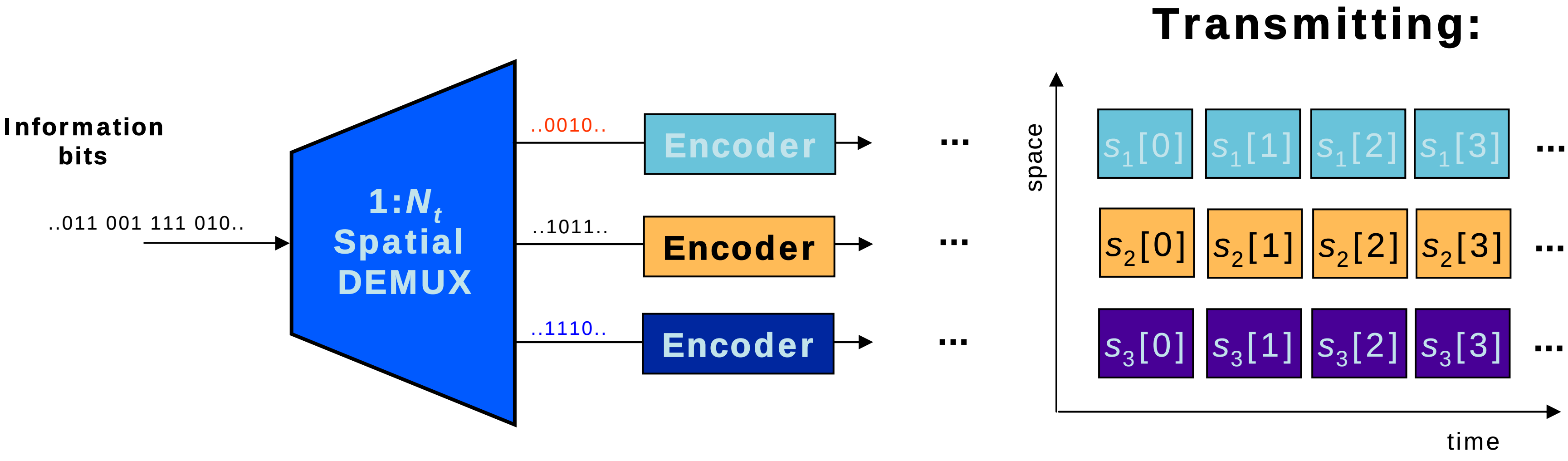
Tim Schenk (tim.schenk@philips.com), June 8, 2007

Back-up Slides

V-BLAST Encoder

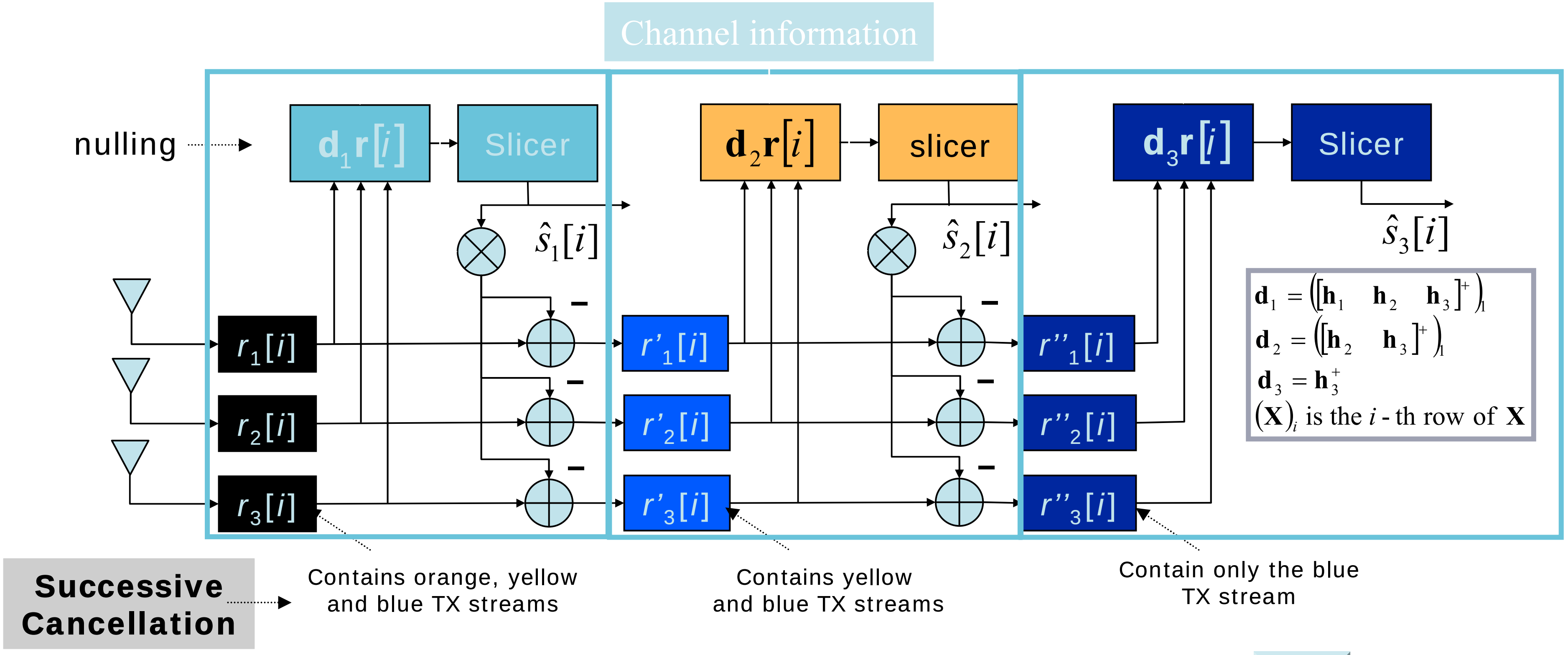
Vertical-BLAST, i.e., MMSE with Decision Feedback (DFB) and optimal ordering

Example: $(N_t,N_r) = (3,3)$



V-BLAST Decoder

Assume optimal ordering is performed; without loss of generality assume that $\rho_{s_1} > \rho_{s_2} > \rho_{s_3}$
where ρ_{s_i} is the post-detection SNR of TX stream s_i



$\mathbf{d}_1 = \left(\left[\mathbf{h}_1 \quad \mathbf{h}_2 \quad \mathbf{h}_3\right]^+\right)_1$

$\mathbf{d}_2 = \left(\left[\mathbf{h}_2 \quad \mathbf{h}_3\right]^+\right)_1$

$\mathbf{d}_3 = \mathbf{h}_3^+$

$(\mathbf{X})_i$ is the i-th row of $\mathbf{X}$

Contains orange, yellow and blue TX streams

Contains yellow and blue TX streams

Contain only the blue TX stream