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Blind detection, spectral estimation, and demodulation of OFDM signals for ultra-wideband cognitive radio applications*

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in the frame of European research project FP7-2007-ICT-1-216785 (“Ultra-wide band real-time interference monitoring and cellular management strategies (UCELLS)”)

Outline

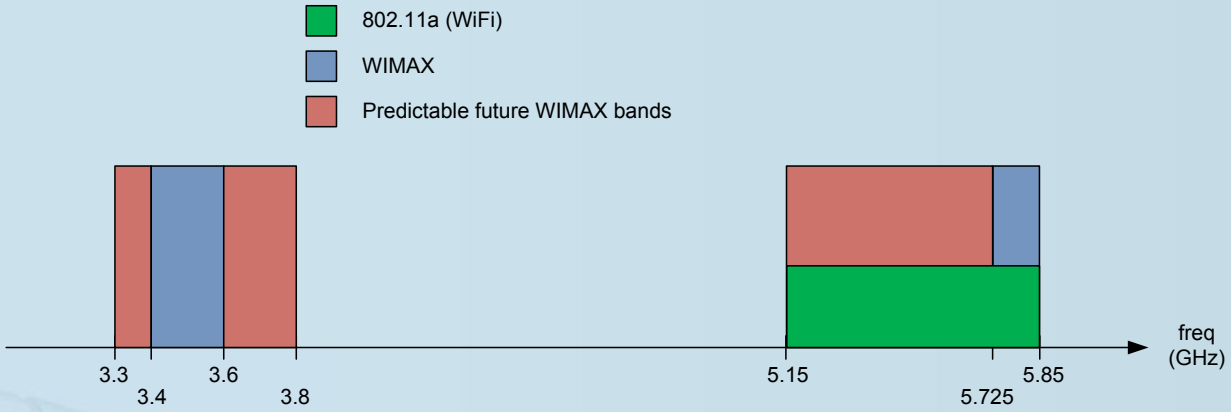
- Introduction
  - ultra-wideband cognitive radio applications
  - observed signal model
  - power spectral density of OFDM signals
- Blind detection and spectral estimation of OFDM signals
  - blind detection and spectral estimation block scheme
  - joint carrier frequency estimation and detection algorithm
  - joint bandwidth-power-SNR estimation algorithm
- Blind demodulation of OFDM signals
  - blind synchronization by time offset estimation
  - blind OFDM demodulation by modulation parameter estimation
  - blind equalization by phase offset and channel estimation
  - blind subcarrier demodulation by constellation estimation
- Conclusion

Introduction: UWB CR applications (1)

- Ultra-wideband cognitive radio applications:
  - UWB management: adapt UWB carrier frequency (CF) and/or power to activity of other radio signals (UWB and non-UWB)
    - ✓ blind detection of other radio signals
    - ✓ blind estimation of frequency band occupied by other radio signals
    - ✓ blind estimation of signal powers
  - UWB reception: demodulate and recover UWB data without knowledge of UWB transmission parameters
    - ✓ blind detection of UWB signals
    - ✓ blind estimation of UWB RF modulation parameters (CF, bandwidth)
    - ✓ blind estimation of UWB OFDM modulation parameters (symbol duration, time guard interval duration, number of subcarriers)
    - ✓ blind estimation of channel parameters (channel model, noise variance)
    - ✓ blind estimation of subcarrier modulation parameters (constellation)

Introduction: UWB CR applications (2)

- Radio environment:
  - UWB covers 3.1–10.6 GHz range (WiMedia ECMA-368)
  - other radio signals in this range are WiFi (802.11a) and WiMAX



- Consequences for estimation/detection algorithm design:
  - all signals of interest are OFDM signals
  - spectral overlap possible in 3.3–3.8 and 5.15–5.85 GHz bands
  - algorithms operate at 20 GHz (Nyquist sampling)
  - OFDM time guard interval can be cyclic prefix or zero padding
  - signal bandwidth ranges from 1.25 (WiMAX) to 528 MHz (UWB)

Introduction: Observed signal model (1)

- Observed signal model (continuous time):

$$y(t) = \sum_{p=1}^{P} \left\{ h^{(p)}(t) * \left[A_c^{(p)} e^{j(2\pi f_c^{(p)} t + \phi_c^{(p)})} x^{(p)}(t) \right] \right\} + n(t)$$

  - $x^{(p)}(t)$ = *p*th transmitter's analog baseband signal (compl. env.)
  - $A_c^{(p)}, f_c^{(p)}, \phi_c^{(p)}$ = *p*th transmitter's RF carrier signal parameters
  - $h^{(p)}(t)$ = channel impulse response w.r.t. *p*th transmitter
  - $n(t)$ = additive white Gaussian noise (AWGN)

Introduction: Observed signal model (2)

- Observed signal model (continuous time):

$$y(t) = \sum_{p=1}^{P} \left\{ h^{(p)}(t) * \left[A_c^{(p)} e^{j(2\pi f_c^{(p)} t + \phi_c^{(p)})} x^{(p)}(t) \right] \right\} + n(t)$$

- Transmitted OFDM signal model (continuous time):

$$x^{(p)}(t) = \sum_{n=-\infty}^{\infty} x^{(p)}[n] g_{\mathrm{I}}^{(p)}(t - nT_s^{(p)})$$

  - $x^{(p)}[n]$ = *p*th transmitter's discrete-time baseband signal
  - $T_s^{(p)}$ = *p*th transmitter's sampling period
  - $g_{\mathrm{I}}^{(p)}(t)$ = *p*th transmitter's D/A interpolation filter

Introduction: Observed signal model (3)

- Observed signal model (continuous time):

$$y(t) = \sum_{p=1}^{P} \left\{ h^{(p)}(t) * \left[A_c^{(p)} e^{j(2\pi f_c^{(p)} t + \phi_c^{(p)})} x^{(p)}(t) \right] \right\} + n(t)$$

- Transmitted OFDM signal model (continuous time):

$$x^{(p)}(t) = \sum_{n=-\infty}^{\infty} x^{(p)}[n] g_{\mathrm{I}}^{(p)}(t - nT_s^{(p)})$$

- Transmitted OFDM signal model (discrete time):

$$x^{(p)}[n] = \sum_{l=-\infty}^{\infty} \left(\sum_{k=0}^{N^{(p)}-1} c_{k,l}^{(p)} g_{\mathrm{P}}^{(p)}[n - lM^{(p)}] e^{j(2\pi/N^{(p)})k(n-lM^{(p)})} \right)$$

  - $c_{k,l}^{(p)}$ = *p*th transmitter's complex data symbol (fin. alph. constell.)
  - $g_{\mathrm{P}}^{(p)}[n]$ = *p*th transmitter's pulse shaping window
  - $N^{(p)}$ = *p*th transmitter's number of OFDM subcarriers
  - $M^{(p)}$ = *p*th transmitter's OFDM symbol length

Introduction: Power spectral density (1)

- Power spectral density (PSD) definition:

$$P_x(f) = \lim_{\mathcal{T}\to\infty}\left(\frac{1}{\mathcal{T}}E\left\{\left|\mathcal{F}\left\{x_{\mathcal{T}}(t)\right\}\right|^2\right\}\right)$$

  - $x_{\mathcal{T}}(t)$ = truncated version of $x(t)$ in time interval $(-\mathcal{T}/2, \mathcal{T}/2)$
  - $E\{\cdot\}, \mathcal{F}\{\cdot\}$ = expectation and Fourier operators



Introduction: Power spectral density (2)

- Power spectral density (PSD) definition:

$$P_x(f) = \lim_{\mathcal{T}\to\infty} \left(\frac{1}{\mathcal{T}} E\left\{ \left| \mathcal{F}\left\{ x_{\mathcal{T}}(t) \right\} \right|^2 \right\} \right)$$

- Existing OFDM PSD expressions:
  1. sum of subcarrier spectra [Pauli & Kuchenbecker, 1998] [Zhao, 2000] [Jayalath & Tellambura, 2001] [Liu & Li, 2004]

$$P_x(f) = \frac{\sigma_c^2}{MT_s} \sum_{k=0}^{N-1} \left(\mathrm{sinc}\left[(f - k\Delta_f) MT_s \right] \right)^2$$

  - σ_c^2 = variance of data symbols $c_{k,l}$
  - $\Delta_f = 1/(NT_s)$ = subcarrier separation
  - based on assumption of rectangular pulse shaping window
  - transmitter side sampling & interpolation filtering disregarded ($T_s \to 0$)

Introduction: Power spectral density (3)

- Power spectral density (PSD) definition:

$$P_x(f) = \lim_{\mathcal{T}\to\infty} \left(\frac{1}{\mathcal{T}} E\left\{ \left| \mathcal{F}\left\{ x_{\mathcal{T}}(t) \right\} \right|^2 \right\} \right)$$

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$$P_x(f) = \frac{\sigma_c^2}{MT_s} \sum_{k=0}^{N-1} \left(\mathrm{sinc}\left[(f - k\Delta_f) MT_s \right] \right)^2$$

  2. sum of periodic subcarrier spectra [Cuypers *et al.*, 1998] [Talbot & Farhang-Boroujeny, 2008]

$$P_x(f) = \frac{\sigma_c^2}{MT_s} \sum_{k=0}^{N-1} \left(M\mathrm{sinc}_M\left[(f - k\Delta_f) T_s \right] \right)^2$$

  - $\mathrm{sinc}_M(\cdot)$ = periodic sinc function (Dirichlet kernel)
  - based on assumption of sampled rectangular pulse shaping window
  - based on assumption of brick-wall interpolation filter

Introduction: Power spectral density (4)

- Derivation of new OFDM PSD expression (1):
  - general PSD expression for digital signals [Couch, 1997]:

$$x(t) = \sum_{i=-\infty}^{\infty} a_i f(t - iT) \Rightarrow P_x(f) = \frac{|F(f)|^2}{T}\left[R(0) + 2\sum_{m=1}^{\infty} R(m)\cos(2\pi f m T)\right]$$

  - $F(f)$ = frequency response of pulse shape $f(t)$
  - $R(m)$ = (time-averaged) autocorrelation function of data symbols a_i
  - T = pulse width

Introduction: Power spectral density (5)

- Derivation of new OFDM PSD expression (1):
  - general PSD expression for digital signals [Couch, 1997]:

$$x(t) = \sum_{i=-\infty}^{\infty} a_i f(t - iT) \Rightarrow P_x(f) = \frac{|F(f)|^2}{T} \left[R(0) + 2 \sum_{m=1}^{\infty} R(m) \cos(2\pi f m T) \right]$$

  - recall the continuous-time transmitted OFDM signal model

$$x(t) = \sum_{n=-\infty}^{\infty} \left[\sum_{l=-\infty}^{\infty} \left(\sum_{k=0}^{N-1} c_{k,l} g_{\mathrm{P}}[n - lM] e^{j(2\pi/N)k(n-lM)} \right) \right] g_{\mathrm{I}}(t - nT)$$

  - one particular mapping between the above expressions, with

$$a_n = \sum_{l=-\infty}^{\infty} \left(\sum_{k=0}^{N-1} c_{k,l} g_{\mathrm{P}}[n - lM] e^{j(2\pi/N)k(n-lM)} \right)$$

  results in

$$P_x(f) = \frac{|G_{\mathrm{I}}(f)|^2}{T_s} \left[R(0) + 2 \sum_{m=1}^{\infty} R(m) \cos(2\pi f m T_s) \right]$$

Introduction: Power spectral density (6)

- Derivation of new OFDM PSD expression (2):
  - the (time-averaged) autocorrelation function can be reduced to

$$R(m) = \begin{cases} \dfrac{N\sigma_c^2}{M} \displaystyle\sum_{n=0}^{2N-1} (g_{\mathrm{P}}[n])^2, & \text{for } m = 0 \\ \dfrac{N\sigma_c^2}{M} \displaystyle\sum_{n=N}^{M-1} g_{\mathrm{P}}[n] g_{\mathrm{P}}[n-N], & \text{for } m = \pm N \\ 0, & \text{for } m \neq 0, \pm N \end{cases}$$

  under the assumption that:

1. the data symbols $c_{k,l}$ are i.i.d.
2. the pulse shape is sufficiently localized in time, i.e.,

$$g_{\mathrm{P}}[n] = 0, \; n \notin \{0, \ldots, 2N-1\}$$

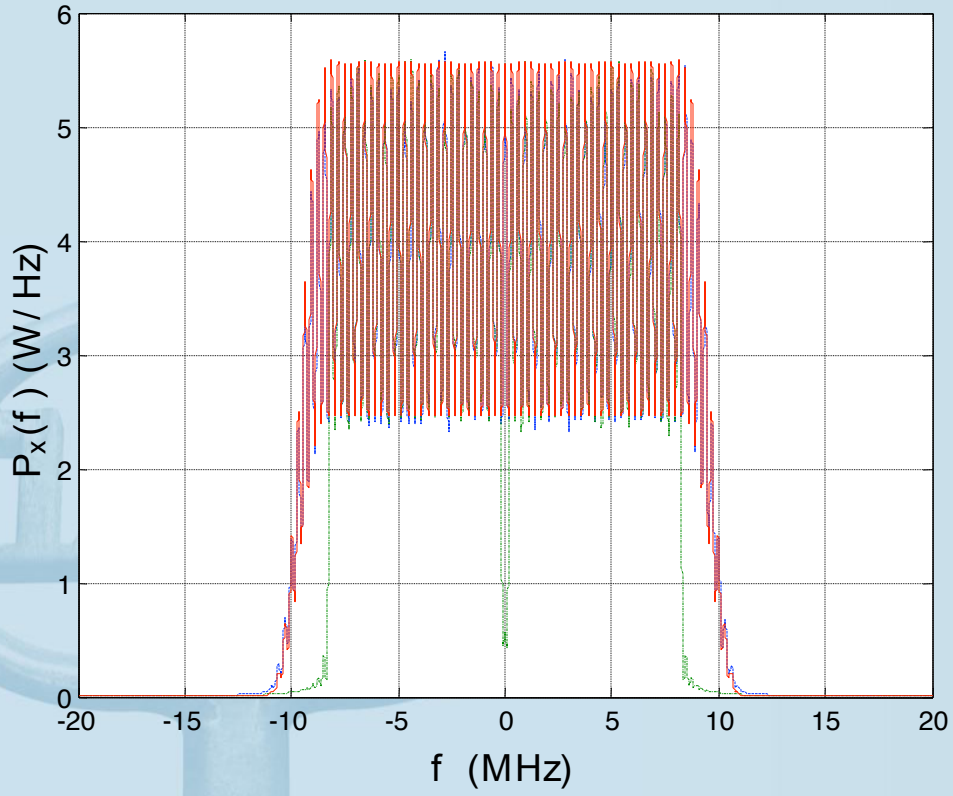
[T. van Waterschoot, V. Le Nir, J. Duplicy, and M. Moonen, “Analytical expressions for the power spectral density of CP-OFDM and ZP-OFDM signals", *IEEE Signal Process. Lett.*, vol. 17, no. 4, Apr. 2010, pp. 371-374.]

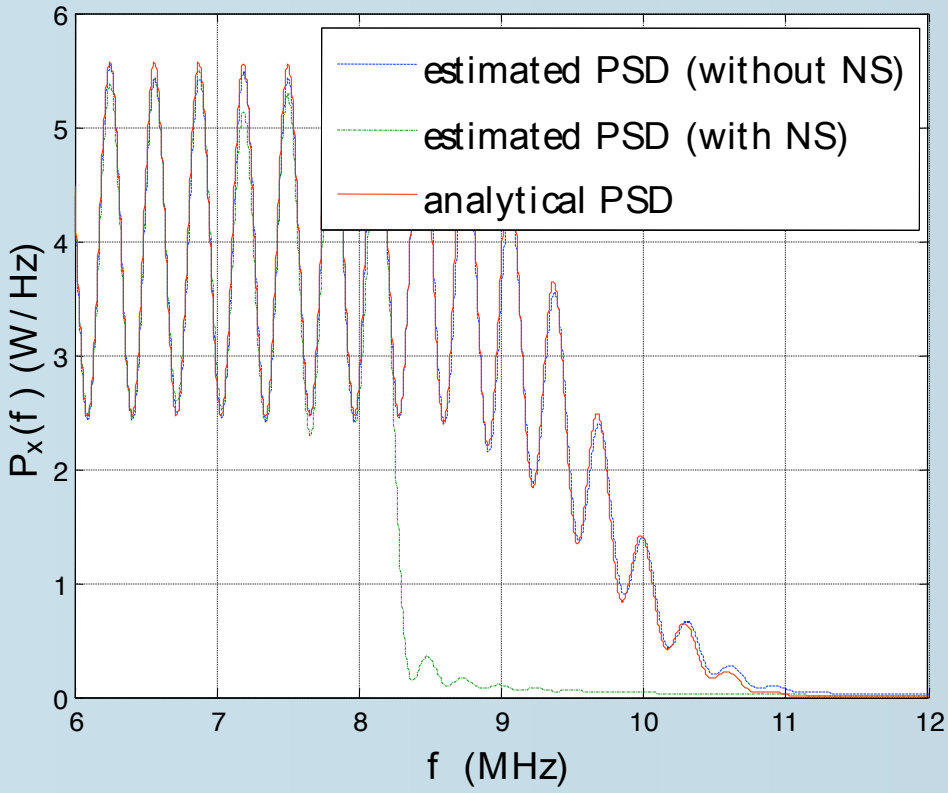
Introduction: Power spectral density (7)

- PSD expression for CP-OFDM signals:

$$P_x(f) = \frac{N|\sigma_c G_{\mathrm{I}}(f)|^2}{MT_s}\left[\sum_{n=0}^{2N-1}(g_{\mathrm{P}}[n])^2 + 2\cos\left(\frac{2\pi f}{\Delta_f}\right)\sum_{n=N}^{M-1} g_{\mathrm{P}}[n]g_{\mathrm{P}}[n-N]\right]$$

- Example: PSD of IEEE 802.11a WiFi signal



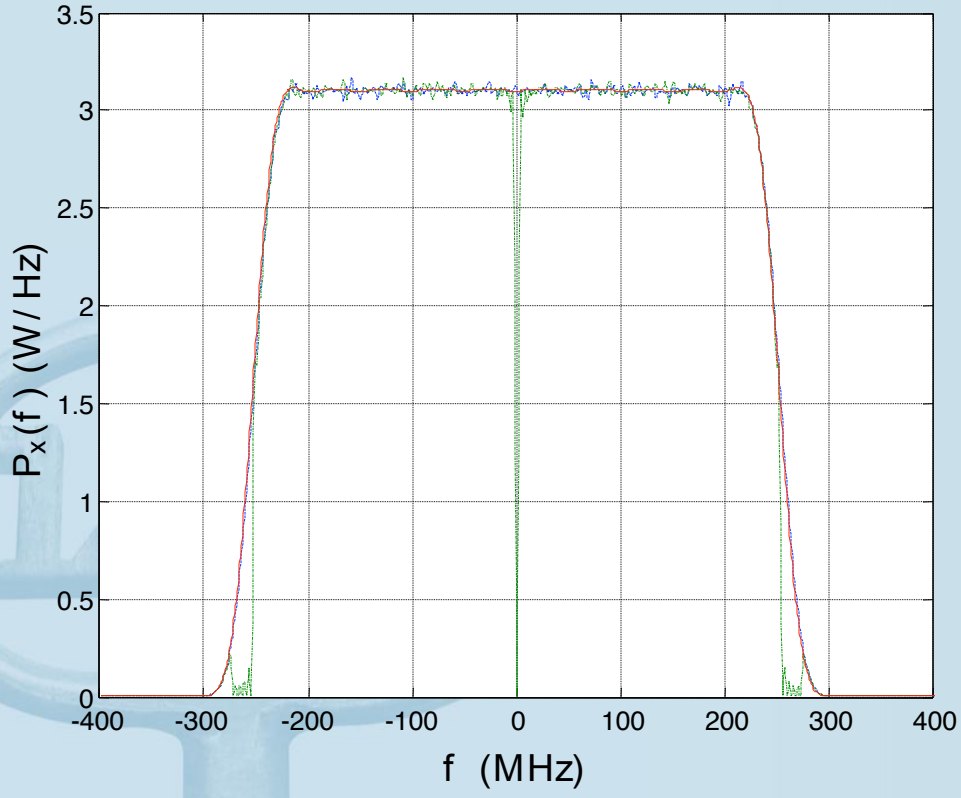


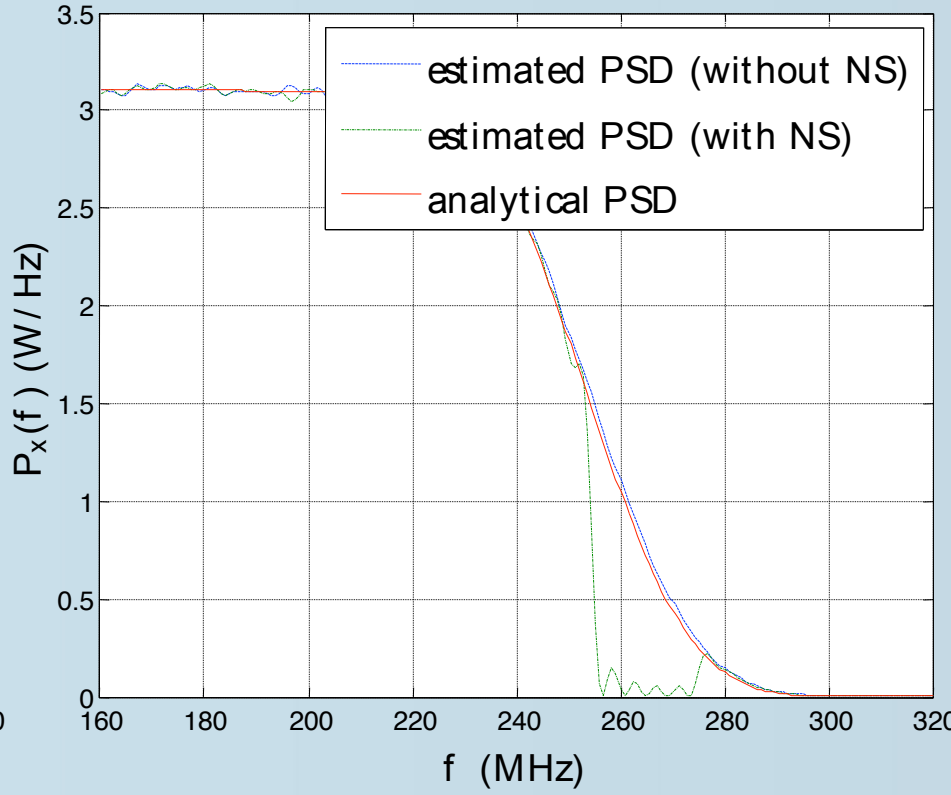
Introduction: Power spectral density (8)

- PSD expression for ZP-OFDM signals:

$$P_x(f) = \frac{N|\sigma_c G_{\mathrm{I}}(f)|^2}{MT_s} \sum_{n=0}^{N-1} (g_{\mathrm{P}}[n])^2$$

- Example: PSD of ECMA-368 UWB signal





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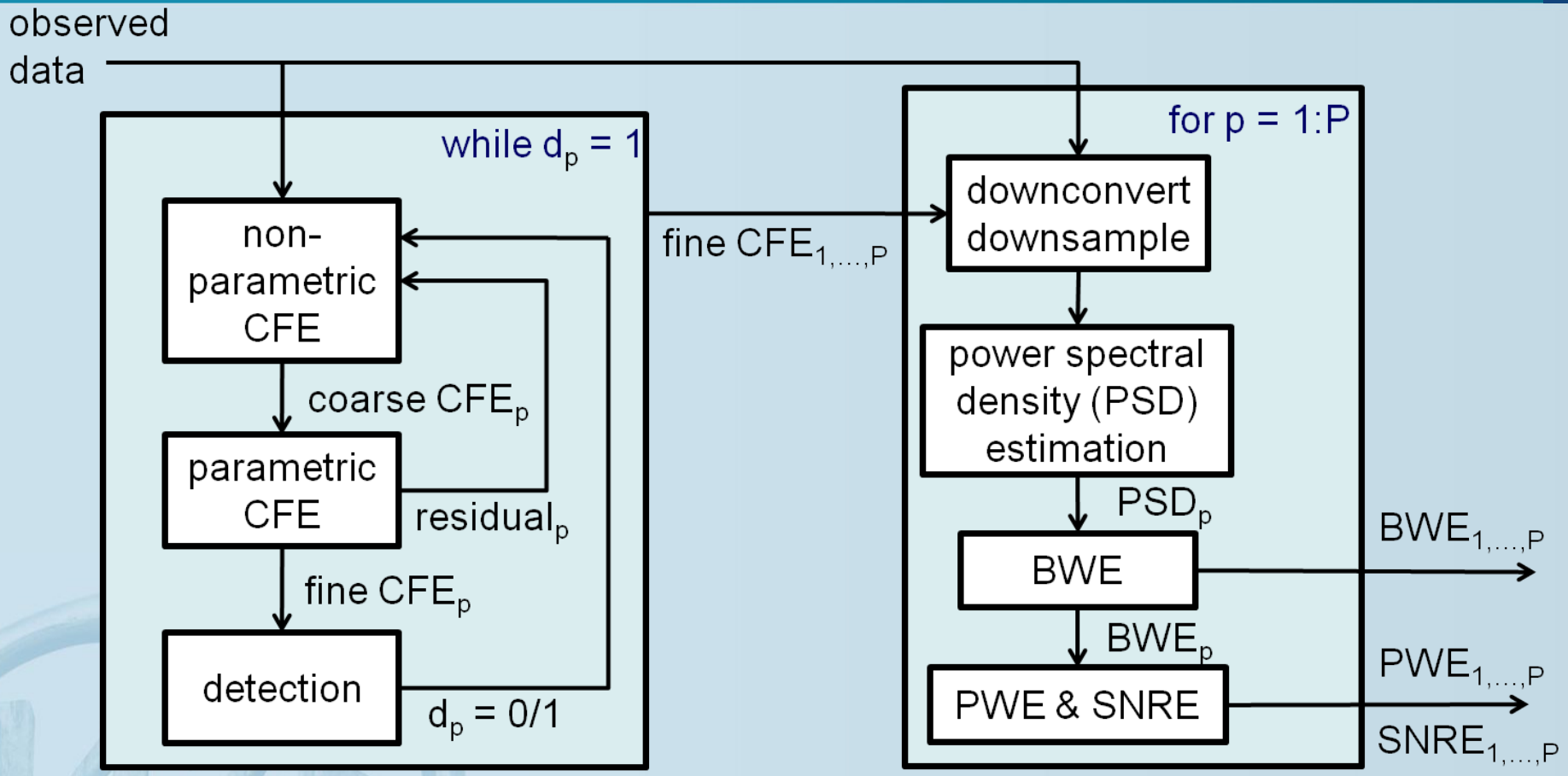
Blind detection and spectral estimation: Block scheme



OFDM carrier frequency estimation (CFE) and detection

OFDM bandwidth, power, SNR (B-P-SNR) estimation

[T. van Waterschoot, V. Le Nir, J. Duplicy, and M. Moonen, “Spectral estimation and detection of OFDM signals for ultra-wideband cognitive radio applications", ESAT-SISTA Technical Report TR 09-150, K.U.Leuven, Belgium, Jul. 2010]

Blind detection and spectral estimation: Carrier frequency estimation (1)

- Step 1: non-parametric carrier frequency estimation
  - DFT magnitude spectrum estimation followed by peak picking
  - provides a (coarse) initial estimate for parametric CFE algorithm
- Step 2: parametric carrier frequency estimation
  - CFE data model:

$$y[n] = \left(\prod_{p=1}^{P} \frac{1 - \rho_p e^{j\theta_p} z^{-1}}{1 - \nu_p e^{j\theta_p} z^{-1}} \right) e[n]$$

P bandpass signals in WGN ~ WGN shaped by *P* bandpass filters

  - CFE criterion:

$$\min_{\vartheta} V(\vartheta) = \min_{\vartheta} \frac{1}{L} \sum_{n=1}^{L} e^2[n, \vartheta] \text{ with } \begin{cases} e[n, \vartheta] = \left(\prod_{p=1}^{P} \frac{1 - \nu_p e^{j\theta_p} z^{-1}}{1 - \rho_p e^{j\theta_p} z^{-1}} \right) y[n] \\ \vartheta = [\theta_1 \ \ldots \ \theta_P]^T \end{cases}$$

least-squares criterion maximizes residual spectral flatness

$\rightarrow$ notch filters are expected to align with bandpass signals

Blind detection and spectral estimation: Carrier frequency estimation (2)

- Parametric carrier frequency estimation algorithm (1)
  - decoupling of CFE criterion into *P* subproblems:

$$\min_{\theta_p} V_p(\vartheta_p) = \min_{\theta_p} \frac{1}{L} \sum_{n=1}^{L} e_p^2[n, \vartheta_p] \text{ with } \begin{cases} e_p[n, \vartheta_p] = \left(\prod_{l=1}^{p} \frac{1 - \nu_l e^{j\theta_l} z^{-1}}{1 - \rho_l e^{j\theta_l} z^{-1}} \right) y[n] \\ \vartheta_p = [\theta_1 \ \ldots \ \theta_p]^T \end{cases}$$

    → each subproblem is scalar estimation problem

    → detection algorithm can be “interleaved ” with CFE algorithm

  - subproblem solution using line search optimization method:

$$\theta_p^{(k+1)} = \theta_p^{(k)} + \mu_k \psi^{(k)}$$

    - μ_k = step length
    - $\psi^{(k)}$ = search direction
  - step length calculation:

    backtracking with Armijo’s sufficient decrease condition

    [Nocedal & Wright, 2006]

Blind detection and spectral estimation: Carrier frequency estimation (3)

- Parametric carrier frequency estimation algorithm (2)
  - search direction calculation:

    scalar simplification of quasi-Newton method with damped Broyden-Fletcher-Goldfard-Shanno (BFGS) updating

    [Powell, 1977] [Nocedal & Wright, 2006] [van Waterschoot & Moonen, 2008]

$$\begin{cases} \psi^{(k)} = -B_k^{-1} \frac{\partial}{\partial \theta_p} V_p\left(\hat{\vartheta}_p^{(k)}\right) \\ s_k = \hat{\theta}_p^{(k+1)} - \hat{\theta}_p^{(k)} \\ v_k = \frac{\partial}{\partial \theta_p} V_p\left(\hat{\vartheta}_p^{(k+1)}\right) - \frac{\partial}{\partial \theta_p} V_p\left(\hat{\vartheta}_p^{(k)}\right) \\ B_{k+1} = \max\left(\frac{v_k}{s_k}, \gamma B_k\right) \end{cases}$$

  - gradient calculation:

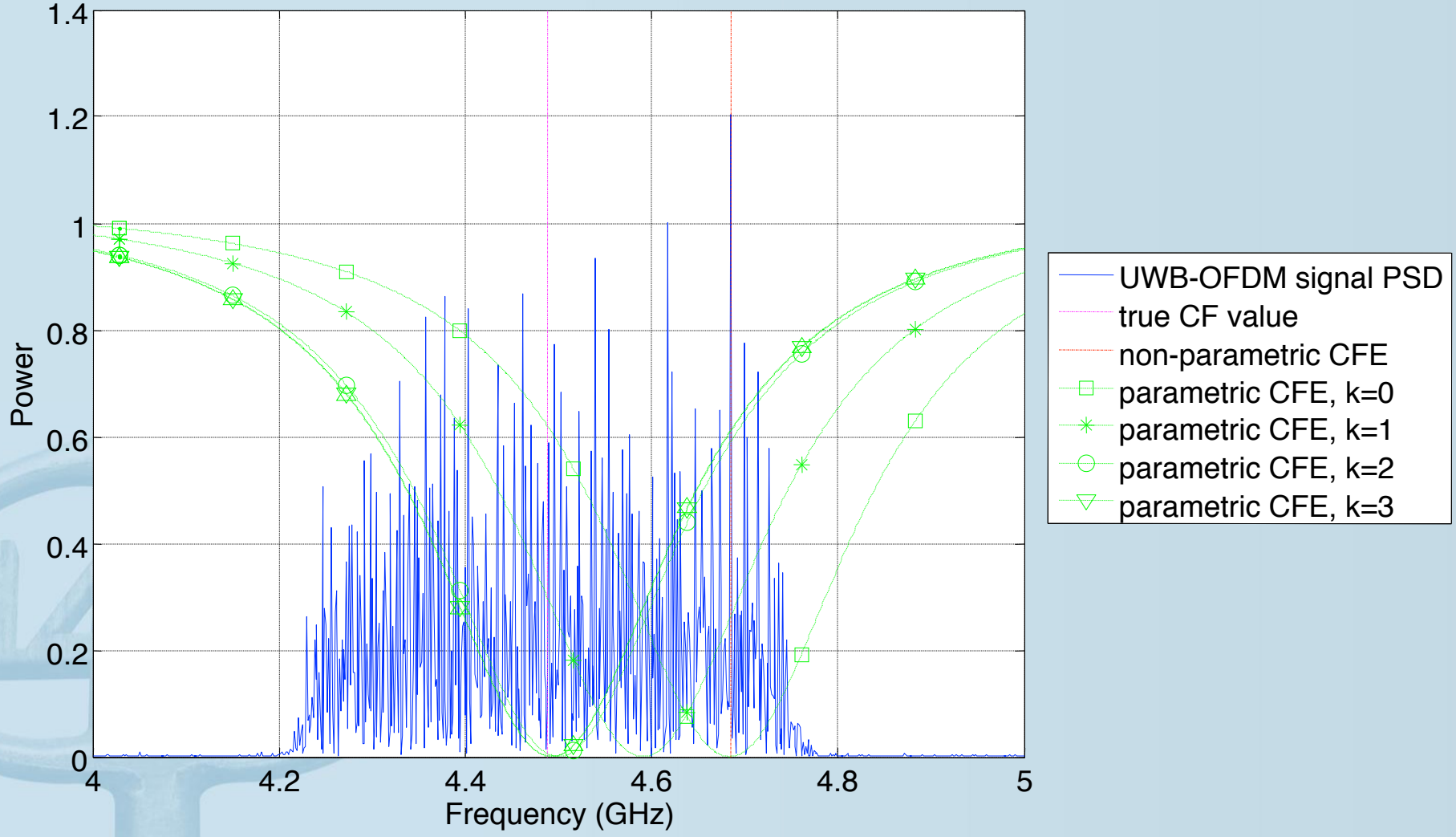
    based on IIR filtering residual signal from previous subproblem

    [van Waterschoot & Moonen, 2008]

$$\begin{cases} e_p[n, \vartheta_p] = \frac{1-\nu e^{j\theta_p} z^{-1}}{1-\rho e^{j\theta_p} z^{-1}} e_{p-1}[n, \vartheta_{p-1}] \\ \frac{\partial}{\partial \theta_p} e_p[n, \vartheta_p] = \frac{j(\rho-\nu) e^{j\theta_p} z^{-1}}{(1-\rho e^{j\theta_p} z^{-1})^2} e_{p-1}[n, \vartheta_{p-1}] \end{cases}$$

Blind detection and spectral estimation: Carrier frequency estimation (4)

- Example: CFE of ECMA-368 UWB signal



Blind detection and spectral estimation: Carrier frequency estimation (5)

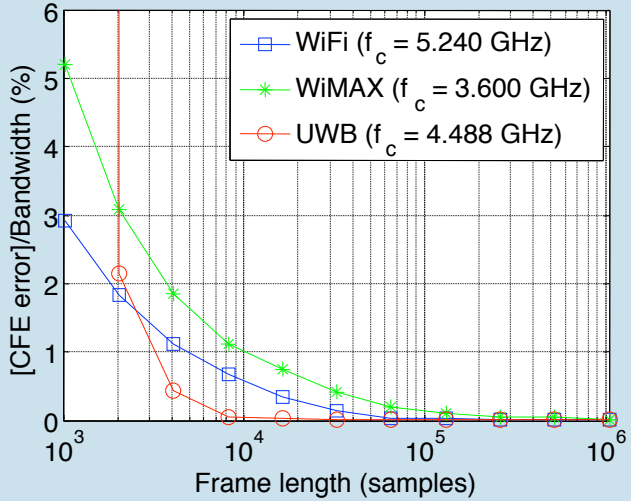
- Performance evaluation:
  - Monte Carlo simulations with 1000 trials per simulation
  - performance measure = absolute CFE error/bandwidth (%)
    vs. frame length (# observed samples) with fixed SNR = 10 dB
    vs. SNR with fixed frame length = 16384 samples
  - lack of prior knowledge on OFDM bandwidth and power values
    $\rightarrow$ fixed bilinear filter pole and zero radii: $\rho = 0.95,\ \nu = 1$
  - OFDM signals under consideration:
    - IEEE 802.11a WiFi signal (f_c = 5.24 GHz)
    - IEEE 802.16 WiMAX signal (f_c = 3.6 GHz)
    - ECMA-368 UWB signal (f_c = 4.488 GHz or 3.432 GHz)
  - observed signal scenarios:
    - single OFDM signal in observed signal
    - three OFDM signals in observed signal (no spectral overlap)
    - three OFDM signals in observed signal (WiMAX-UWB spectral overlap)

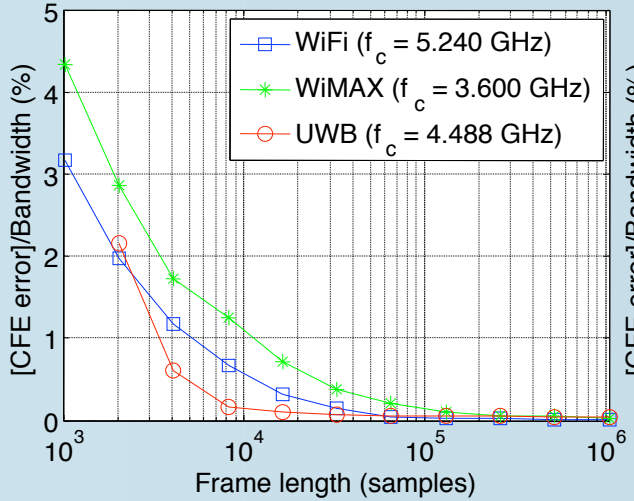
Blind detection and spectral estimation: Carrier frequency estimation (6)

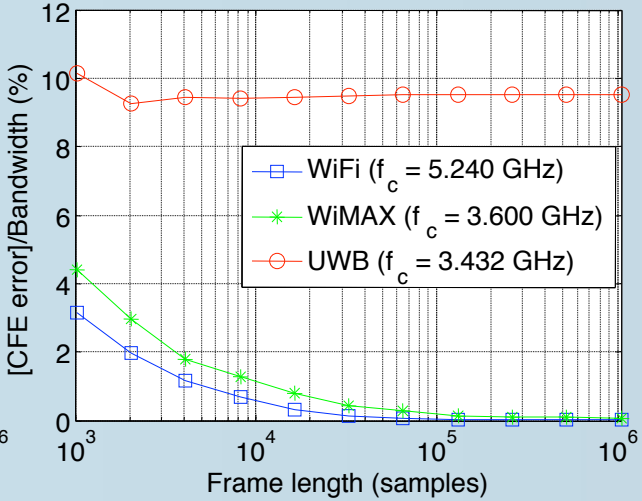
single OFDM signal

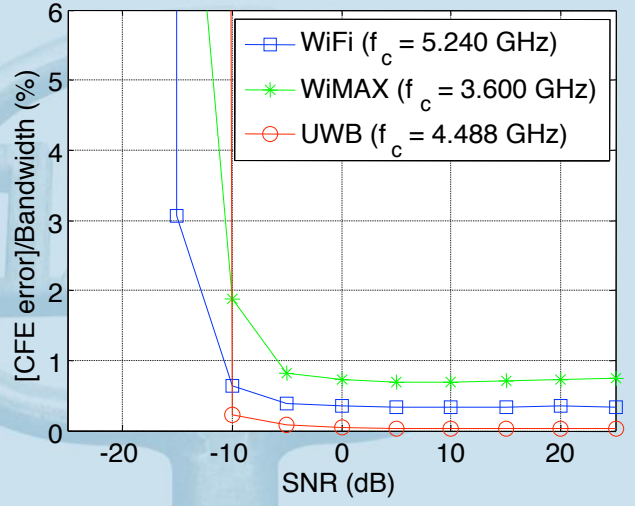
three OFDM signals (no spectral overlap)

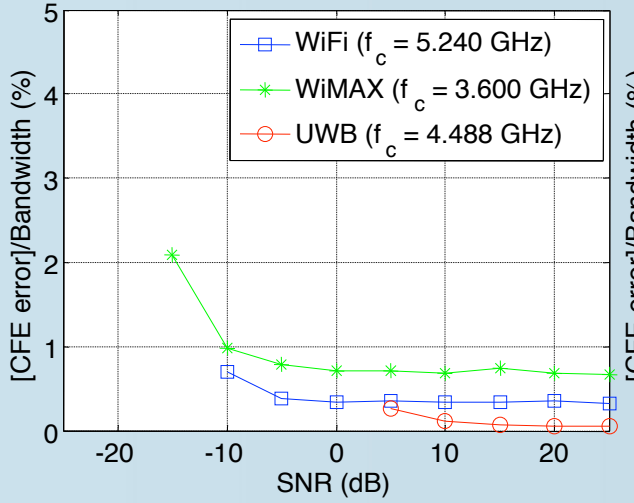
three OFDM signals (overlap WiMAX-UWB)

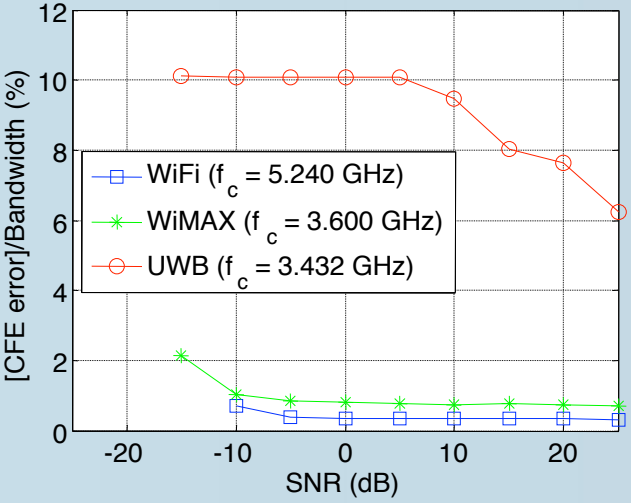












Blind detection and spectral estimation: Detection (1)

- Detection algorithm
  - binary hypothesis test for a given carrier frequency estimate

$$\begin{aligned} \mathcal{H}_0 &: \text{no OFDM signal at } f_c^{(p)} && \text{(Null hypothesis)} \\ \mathcal{H}_1 &: \text{OFDM signal at } f_c^{(p)} && \text{(Alternative hypothesis)} \end{aligned}$$

  - decision variable:

$$\delta^{(p)} = \frac{1}{L}\sum_{n=1}^{L}\left\{\left[\left(\prod_{l=1}^{p}\frac{1-\nu_1 e^{j\theta_l}z^{-1}}{1-\rho_1 e^{j\theta_l}z^{-1}}\right)y[n]\right]^2 - \left[\left(\prod_{l=1}^{p}\frac{1-\nu_2 e^{j\theta_l}z^{-1}}{1-\rho_2 e^{j\theta_l}z^{-1}}\right)y[n]\right]^2\right\}$$

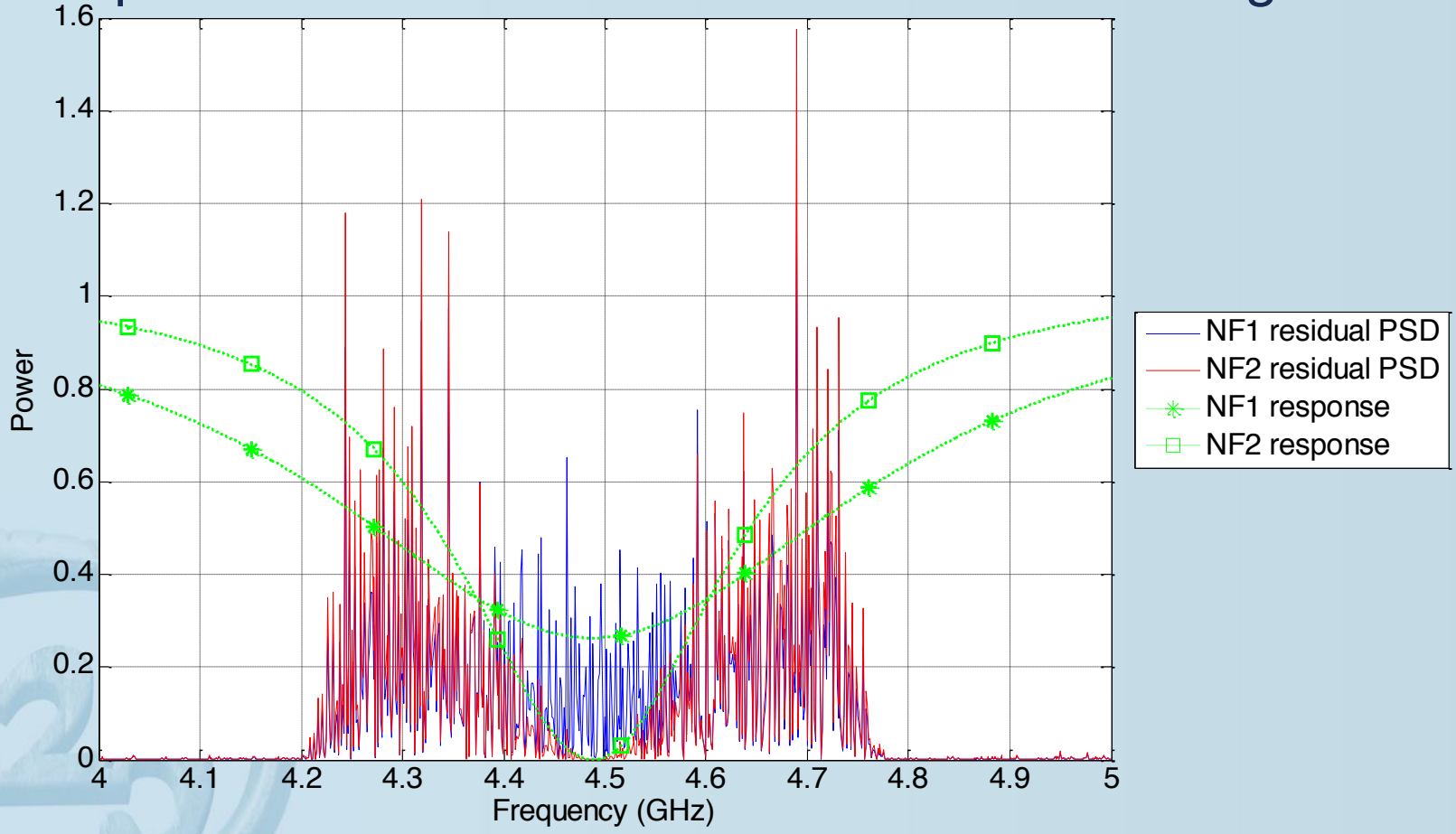
$$\text{with } \rho_1/\nu_1 = \rho_2/\nu_2 \text{ and } \nu_1 < \nu_2$$

  - detection criterion:

$$\delta^{(p)} > 0 \Rightarrow \text{reject } \mathcal{H}_0 \Rightarrow d^{(p)} = 1$$

Blind detection and spectral estimation: Detection (2)

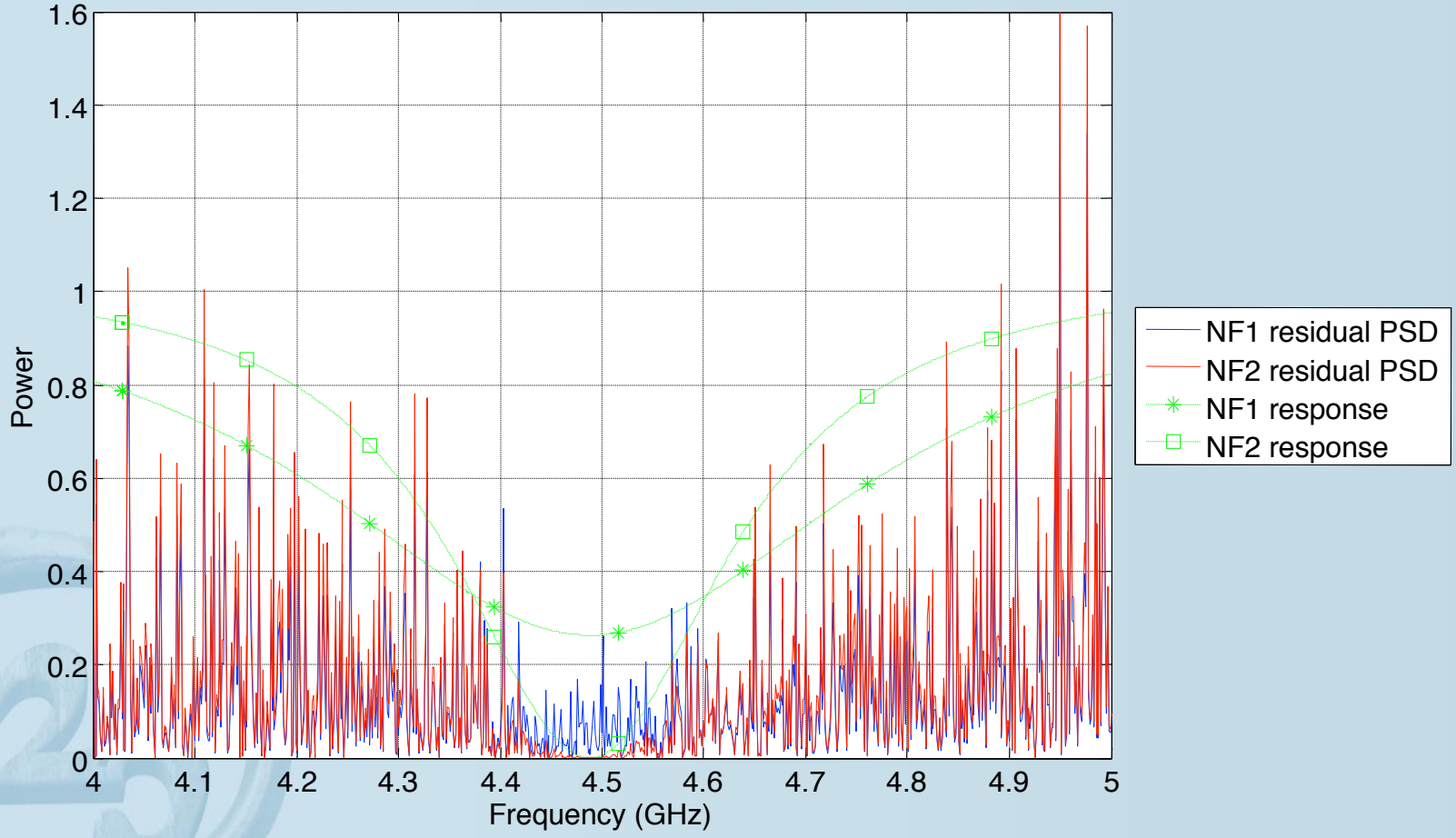
- Example: decision variable for ECMA-368 UWB signal



NF1 output power > NF2 output power $\Rightarrow \delta^{(p)} > 0$

Blind detection and spectral estimation: Detection (3)

- Example: decision variable for noise



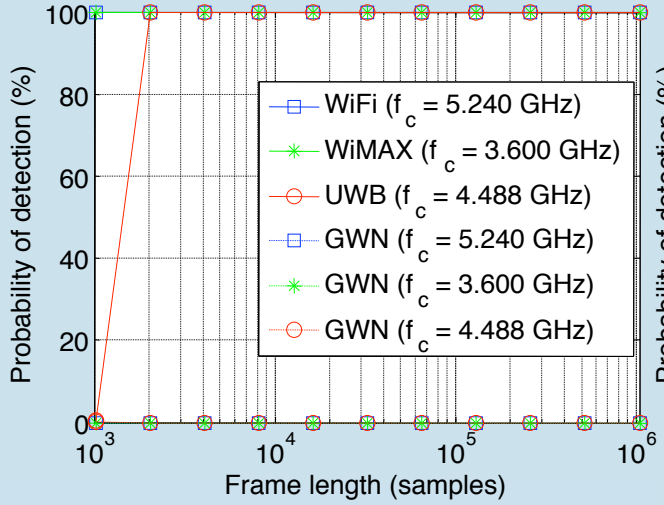
NF1 output power < NF2 output power $\Rightarrow \delta^{(p)} < 0$

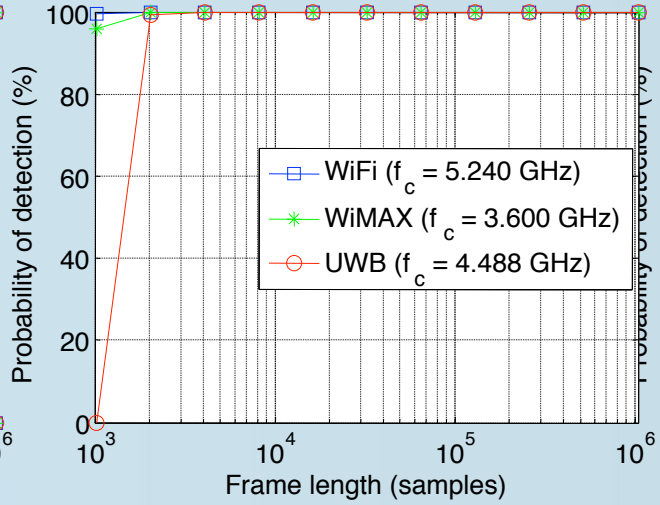
Blind detection and spectral estimation: Detection (4)

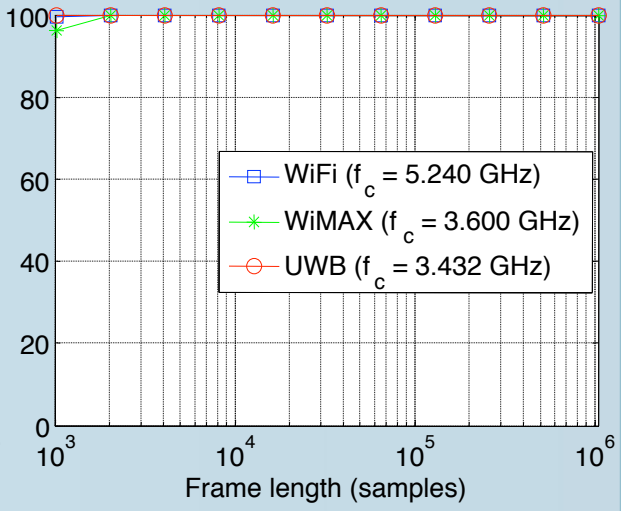
single OFDM signal

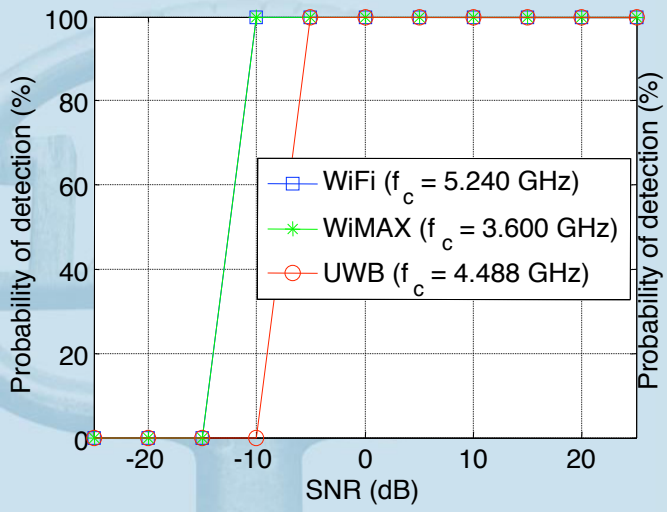
three OFDM signals (no spectral overlap)

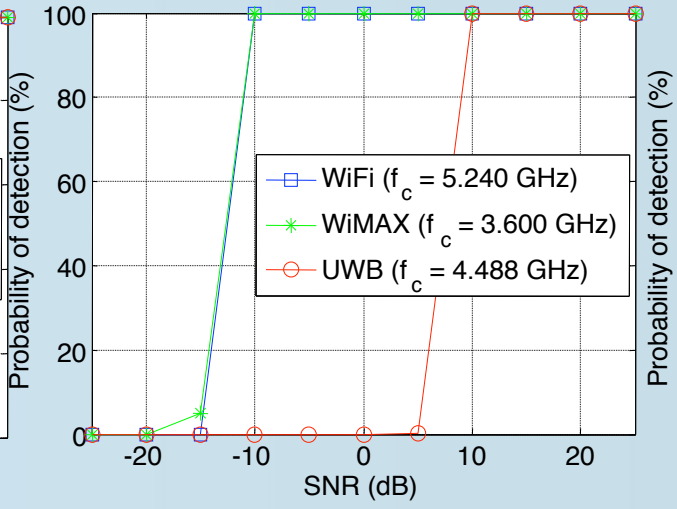
three OFDM signals (overlap WiMAX-UWB)

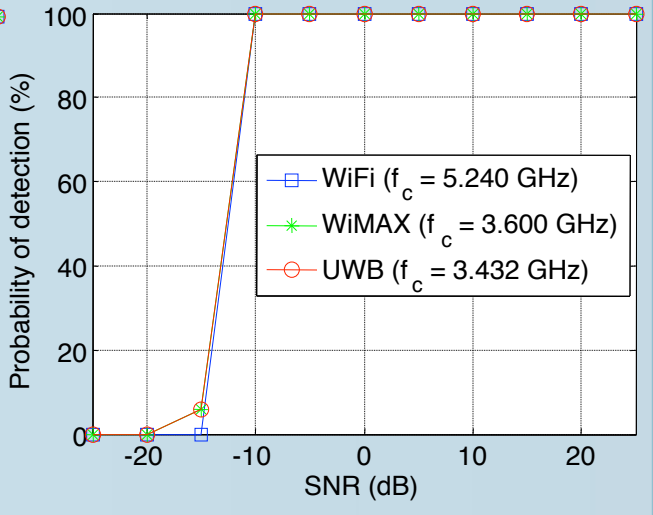












Blind detection and spectral estimation: Bandwidth-power-SNR estimation (1)

- Bandwidth estimation algorithm
  - spectral "isolation" of *p*th OFDM signal:
    1. downconversion of (*p-1*)th CFE algorithm residual signal
    2. downsampling of downconverted signal using fixed $B_{\max}$
  - DFT-based estimation of isolated signal PSD
  - definition of spectral mask:

$$S(f) = \begin{cases} P_L, & f < -B/2 \\ P_H, & -B/2 \le f \le B/2 \\ P_L, & f > B/2 \end{cases}$$

  - bandwidth estimation criterion:

$$\min V(B, P_L, P_H) = \min \int_{-B_{\max}/2}^{B_{\max}/2} [P(f) - S(f)]^2 df$$

    1. (closed-form) minimization of $V(B, P_L, P_H)$ w.r.t. P_L, P_H
    2. (numerical) minimization of $V(B)$ w.r.t. B

Blind detection and spectral estimation: Bandwidth-power-SNR estimation (2)

- Signal power and SNR estimation
  - estimation of signal+noise PSD and noise PSD:

$$\hat{P}_{xn}^{(p)} = \frac{\hat{B}^{(p)}}{B_{\max}} P_H(\hat{B}^{(p)})$$

$$\hat{P}_{n}^{(p)} = \left(1 - \frac{\hat{B}^{(p)}}{B_{\max}}\right) P_L(\hat{B}^{(p)}) \frac{f_s}{2B_{\max}}$$

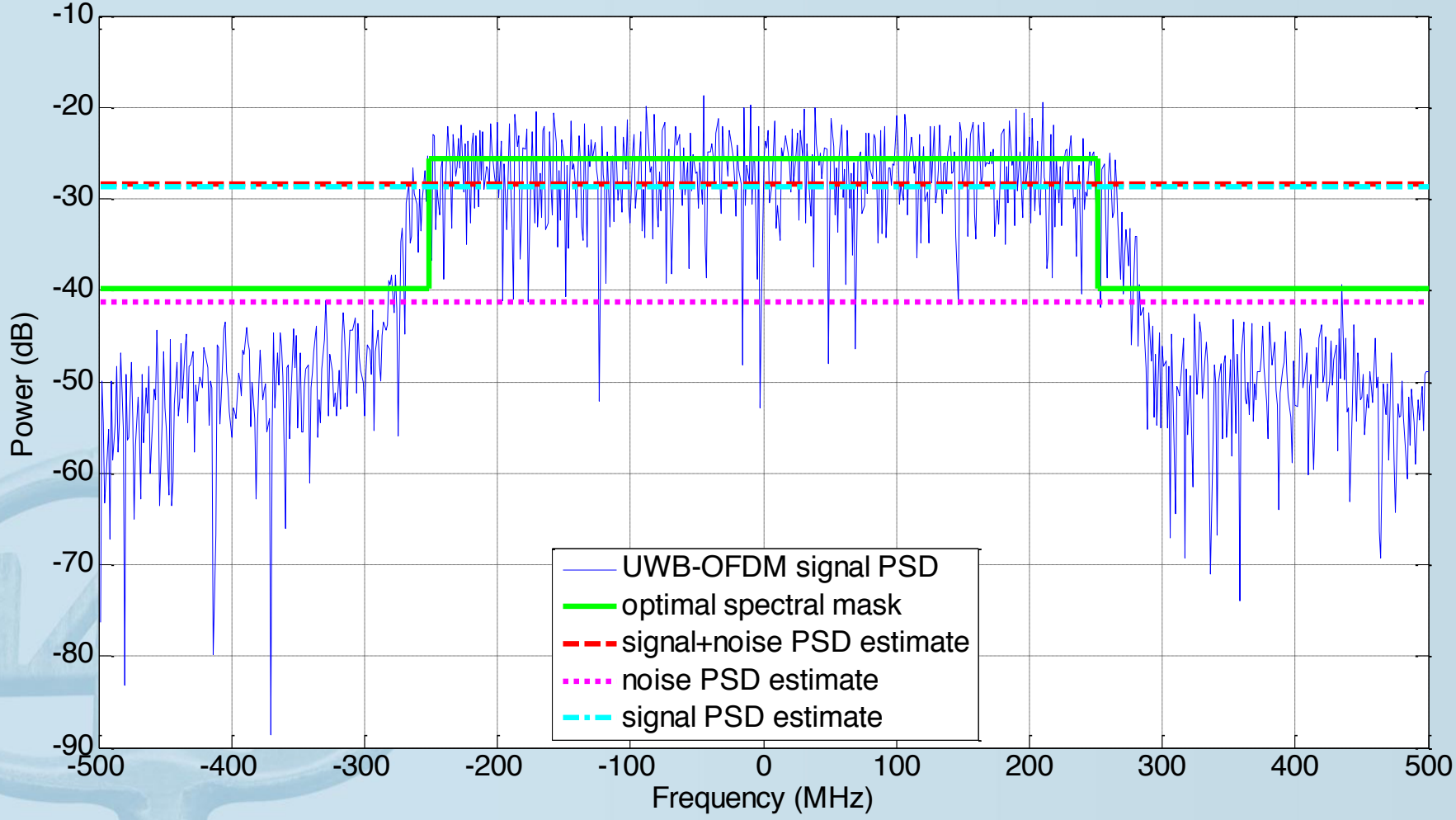
  - estimation of OFDM signal power and SNR:

$$\hat{\Phi}^{(p)} = \hat{B}^{(p)}[\hat{P}_{xn}^{(p)} - \hat{P}_{n}^{(p)}]$$

$$\hat{\mathrm{SNR}}^{(p)} = \frac{\hat{P}_{xn}^{(p)} - \hat{P}_{n}^{(p)}}{\hat{P}_{n}^{(p)}}$$

Blind detection and spectral estimation: Bandwidth-power-SNR estimation (3)

- Example: B-P-SNR estimation of ECMA-368 UWB signal



Blind detection and spectral estimation: Bandwidth estimation

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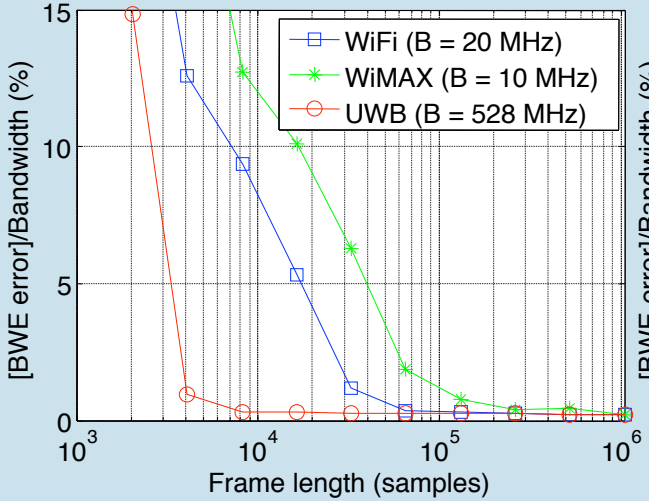
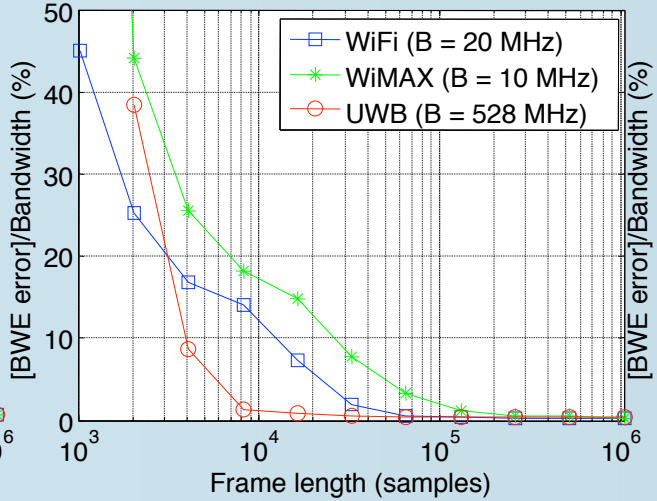
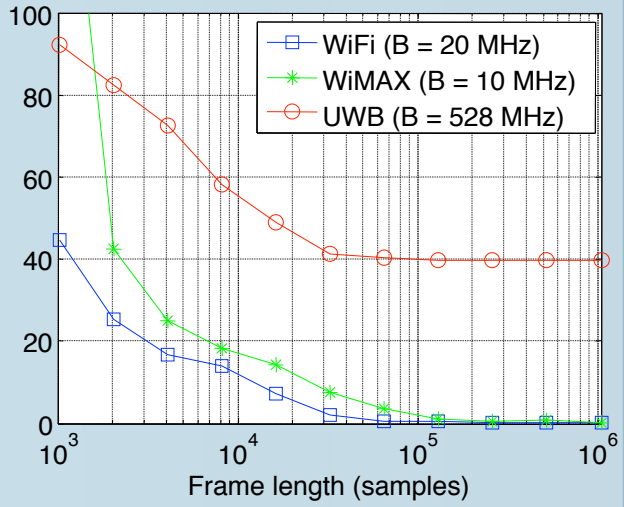
single OFDM signal

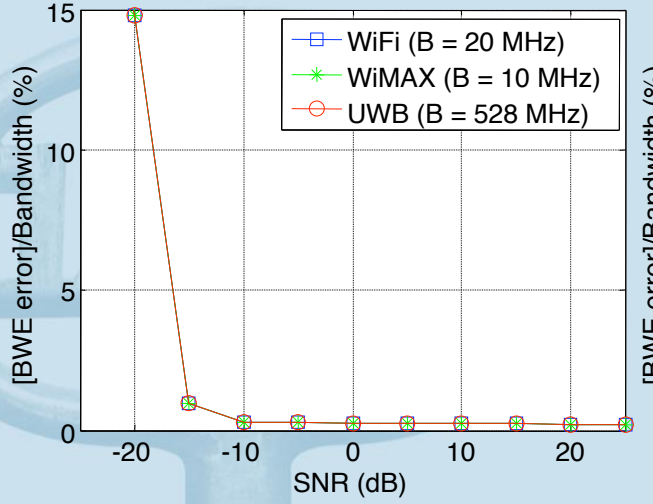
three OFDM signals (no spectral overlap)

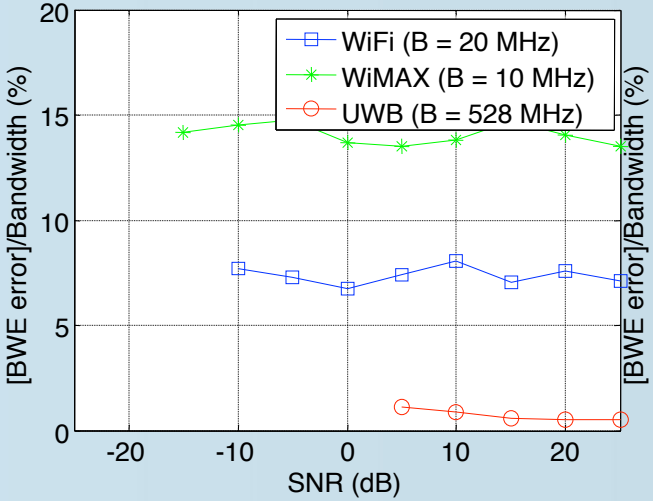
three OFDM signals (overlap WiMAX-UWB)

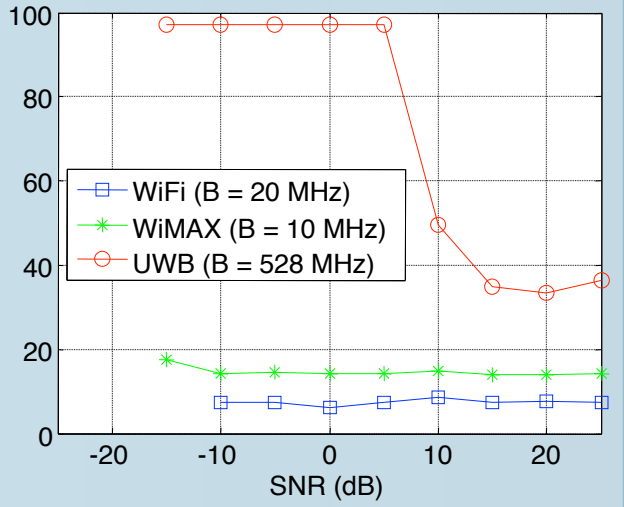












Blind detection and spectral estimation: Power estimation

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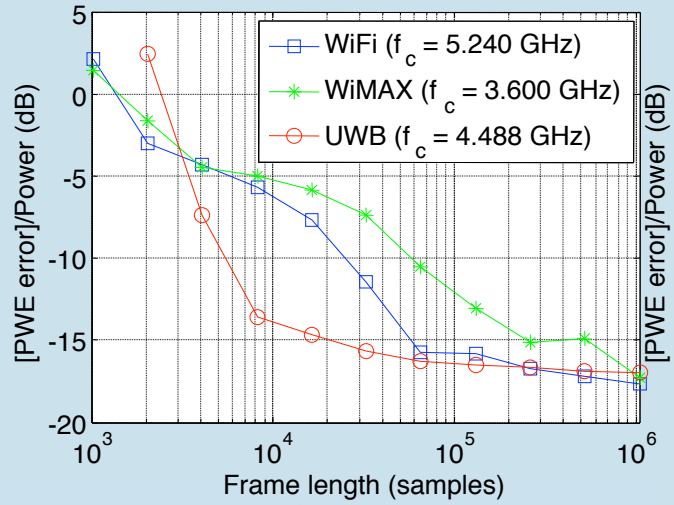
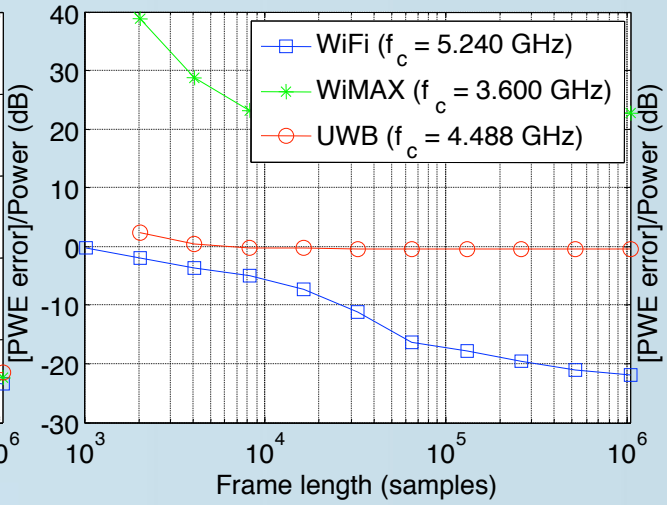
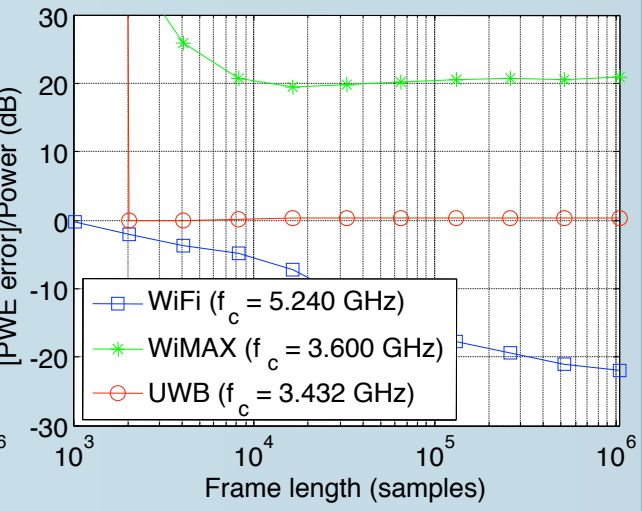
single OFDM signal

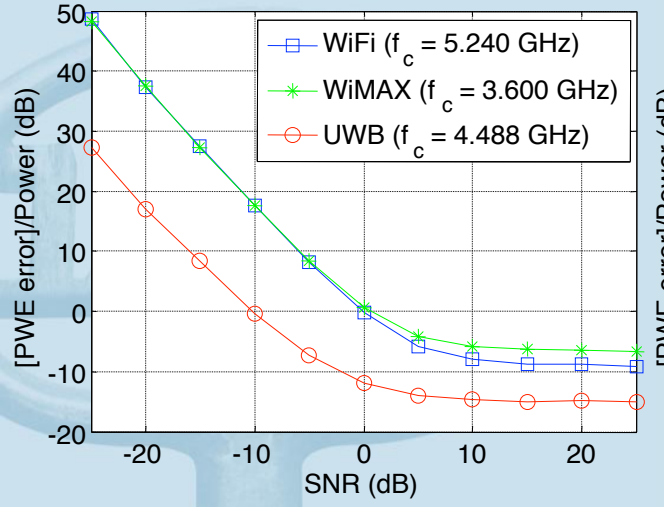
three OFDM signals (no spectral overlap)

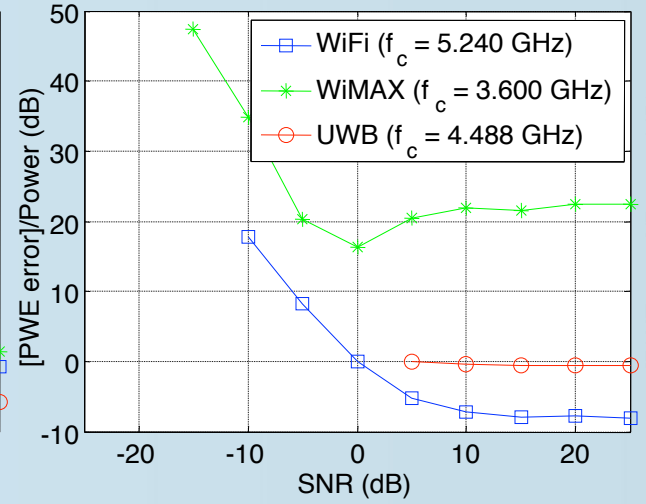
three OFDM signals (overlap WiMAX-UWB)

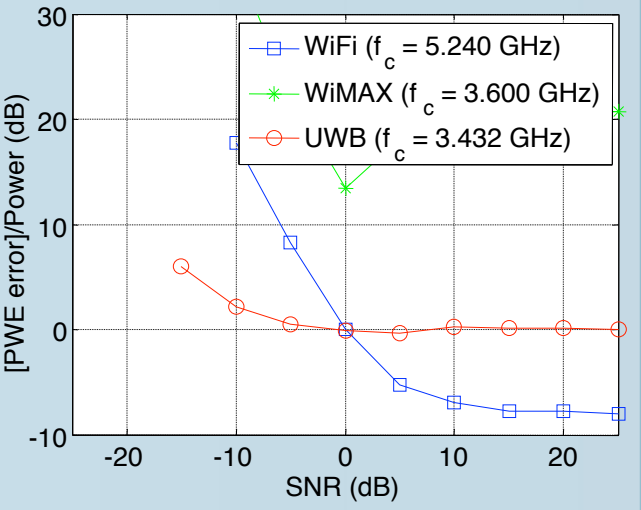












Blind detection and spectral estimation: SNR estimation

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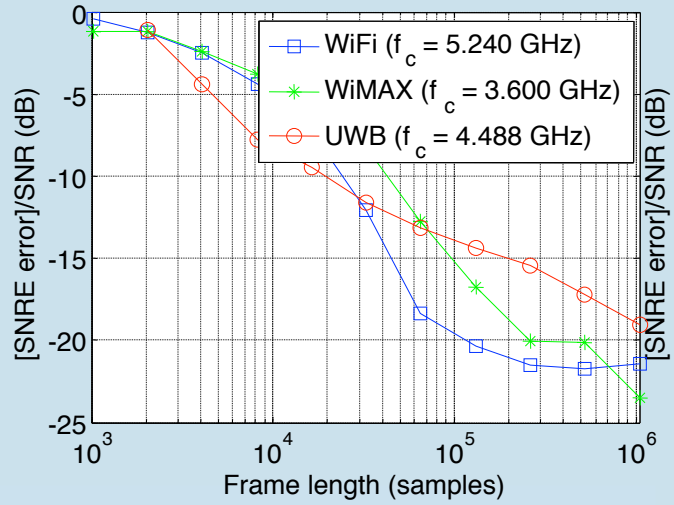
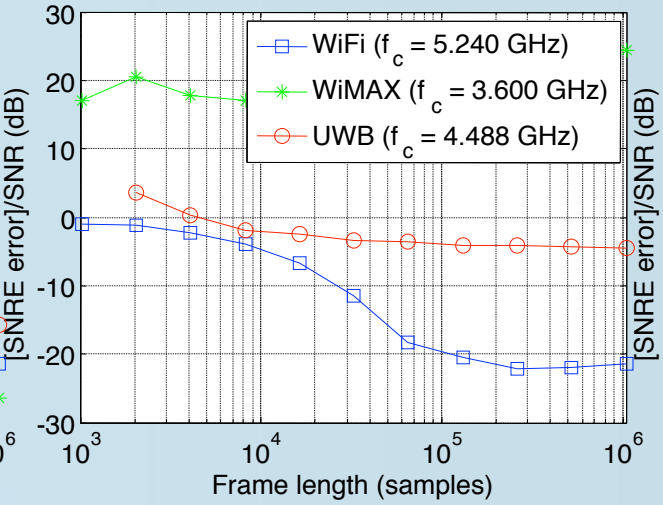
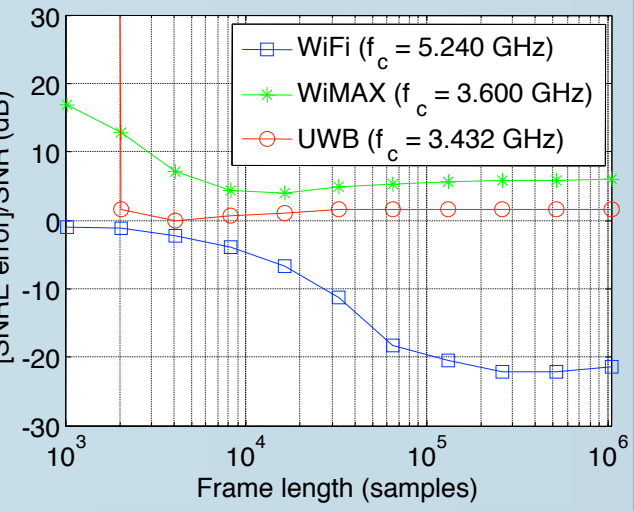
single OFDM signal

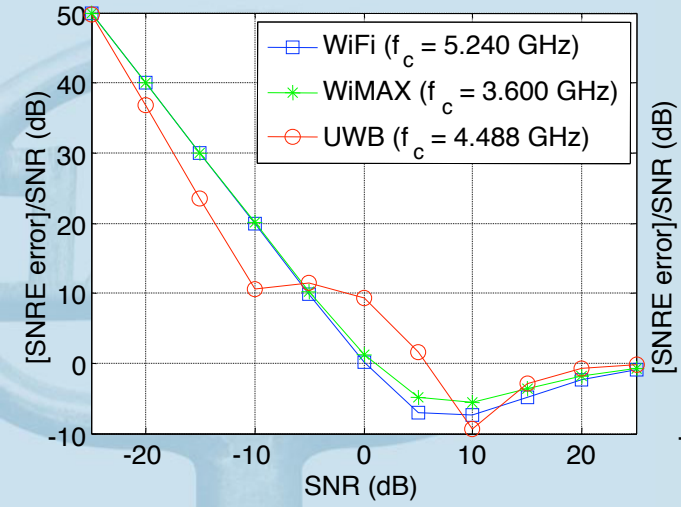
three OFDM signals (no spectral overlap)

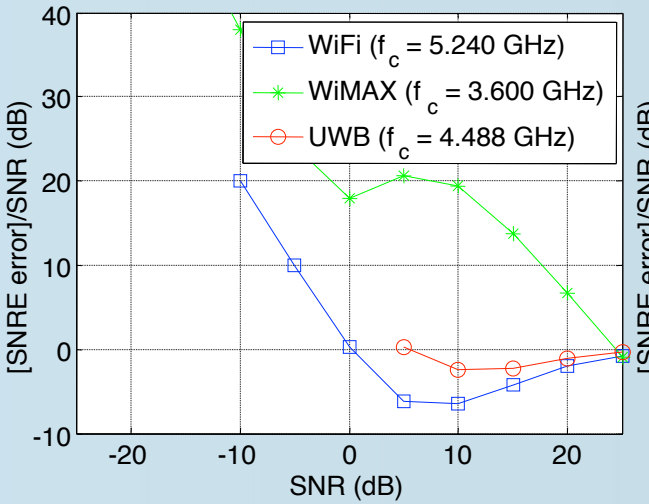
three OFDM signals (overlap WiMAX-UWB)

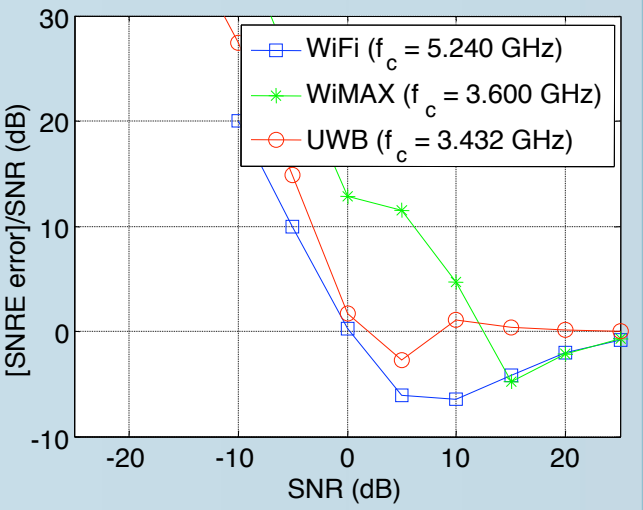












Outline

- Introduction
  - ultra-wideband cognitive radio applications
  - observed signal model
  - power spectral density of OFDM signals
- Blind detection and spectral estimation of OFDM signals
  - blind detection and spectral estimation block scheme
  - joint carrier frequency estimation and detection algorithm
  - joint bandwidth-power-SNR estimation algorithm
- Blind demodulation of OFDM signals
  - blind synchronization by time offset estimation
  - blind OFDM demodulation by modulation parameter estimation
  - blind equalization by phase offset and channel estimation
  - blind subcarrier demodulation by constellation estimation
- Conclusion

Blind demodulation: Requirements

- first two steps based on spectral estimation algorithms
  1. blind RF demodulation using carrier frequency estimate
  2. blind downsampling using bandwidth estimate
- four more steps to blind data recovery
  1. blind synchronization
     - estimation of time offset
  2. blind OFDM demodulation:
     - estimation of time guard interval length
     - estimation of number of subcarriers
  3. blind equalization
     - estimation of phase offset
     - estimation of channel response
  4. blind subcarrier demodulation
     - estimation of subcarrier constellation

Blind demodulation: synchronization (1)

- OFDM symbol synchronization required for time guard interval removal and OFDM demodulation
- existing CP-OFDM time offset estimation criterion:

  [Speth *et al.*, 2001]

$$\hat{\tau} = \arg\max_{\tau} \frac{1}{(L/M-1)(M-N)} \sum_{l=1}^{L/M-1} \left| \sum_{(l-1)M+N+\tau}^{lM+\tau-1} y[n]y^*[n-N] \right|$$

  → estimates first arrival path rather than predominant arrival path

  → cannot be used for ZP-OFDM signals

Blind demodulation: synchronization (2)

- OFDM symbol synchronization required for time guard interval removal and OFDM demodulation
- existing CP-OFDM time offset estimation criterion:

  [Speth *et al.*, 2001]

$$\hat{\tau} = \arg\max_{\tau} \frac{1}{(L/M-1)(M-N)} \sum_{l=1}^{L/M-1} \left| \sum_{(l-1)M+N+\tau}^{lM+\tau-1} y[n]y^*[n-N] \right|$$

- novel time offset estimation criteria:
  - CP-OFDM:

$$\hat{\tau} = \arg\max_{\tau} \left. \frac{\frac{1}{(L/M-1)} \sum_{l=1}^{L/M-1} |y[n+1]-y[n+1-N]|^2}{\frac{1}{(L/M-1)} \sum_{l=1}^{L/M-1} |y[n]-y[n-N]|^2} \right|_{n=lM+\tau-1}$$

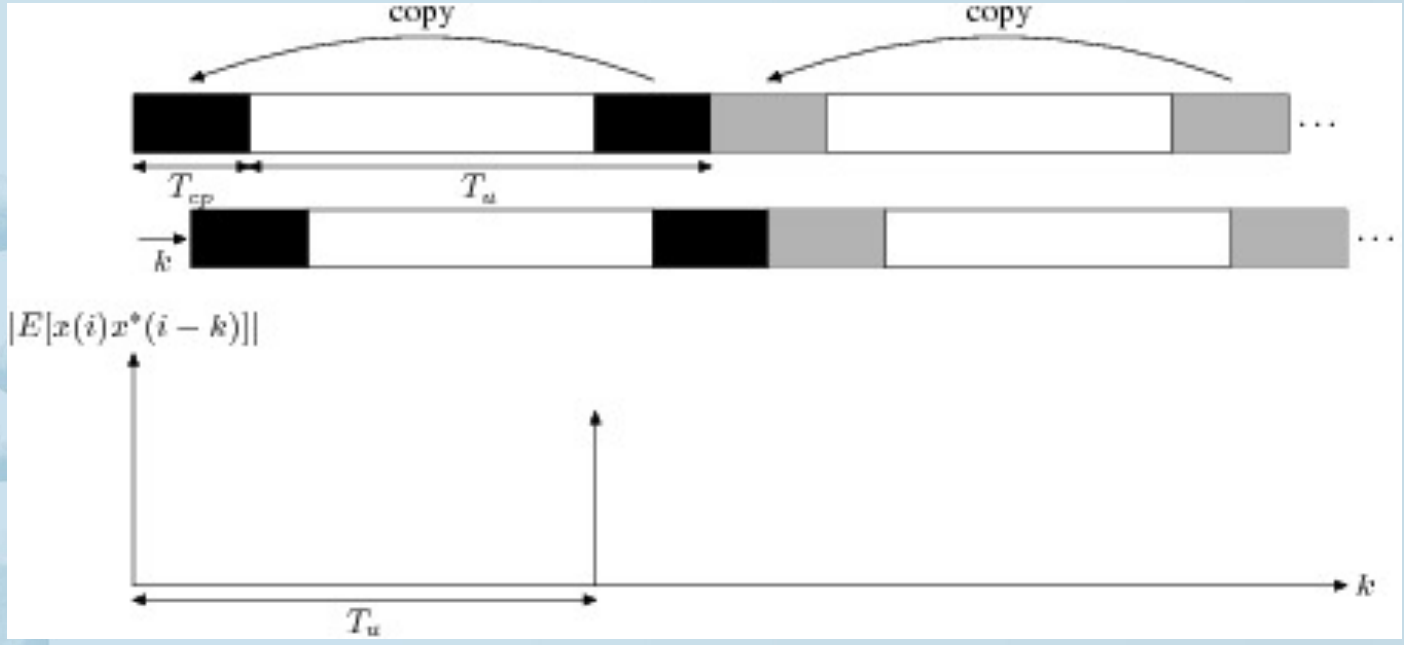
  - ZP-OFDM:

$$\hat{\tau} = \arg\max_{\tau} \left. \frac{\frac{1}{(L/M-1)} \sum_{l=1}^{L/M-1} |y[n+1]|^2}{\frac{1}{(L/M-1)} \sum_{l=1}^{L/M-1} |y[n]|^2} \right|_{n=lM+\tau-1}$$

[V. Le Nir, T. van Waterschoot, J. Duplicy, and M. Moonen, “Blind coarse timing offset estimation for CP-OFDM and ZP-OFDM transmission over frequency selective channels", *EURASIP J. Wireless. Commun. Networking*, vol. 2009.

Blind demodulation: OFDM demodulation (1)

- OFDM symbol synchronization and OFDM demodulation rely on time guard interval length and number of subcarriers
- CP-OFDM parameter estimation:
  - CP induces non-zero autocorrelation at lag *N*
    → number of subcarriers *N* estimated by autocorrelation peak picking
    [Ishii & Wornell, 2005] [Li *et al.*, 2007] [Shi *et al.*, 2007]



Blind demodulation: OFDM demodulation (2)

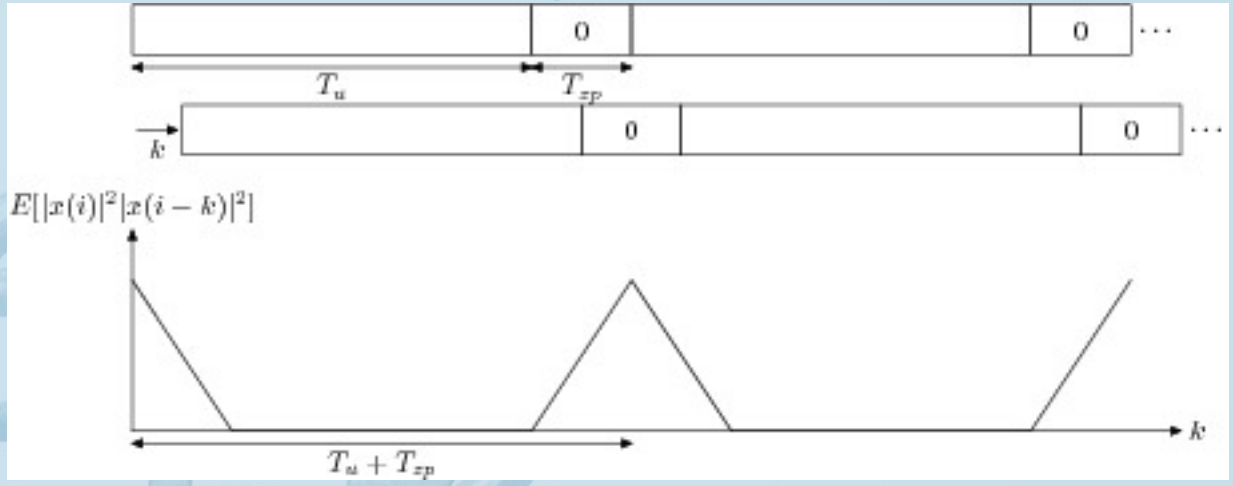
- OFDM symbol synchronization and OFDM demodulation rely on time guard interval length and number of subcarriers
- CP-OFDM parameter estimation:
  - CP induces non-zero autocorrelation at lag *N*
    → number of subcarriers *N* estimated by autocorrelation peak picking
    [Ishii & Wornell, 2005] [Li *et al.*, 2007] [Shi *et al.*, 2007]
  - CP-OFDM signal is cyclostationary with cyclic period *M* [Bolcskei, 2001]
    → symbol length estimated by cyclic correlation peak picking (lag *M*)

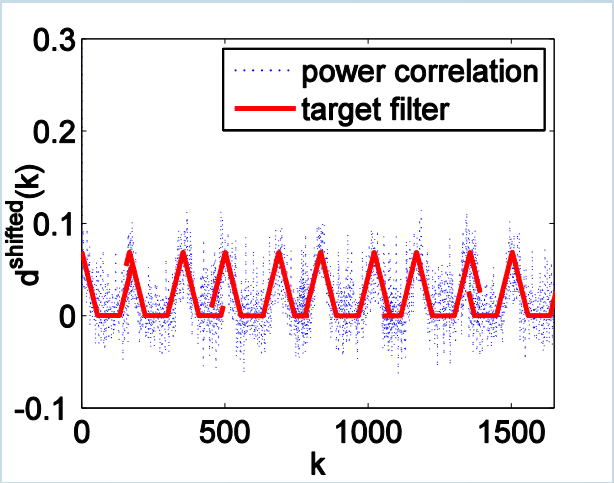


Blind demodulation: OFDM demodulation (3)

- OFDM symbol synchronization and OFDM demodulation rely on time guard interval length and number of subcarriers
- ZP-OFDM parameter estimation:
  - ZP induces non-zero "power autocorrelation" between lags *N* and *N* + *2(M-N)*

  → number of subcarriers *N* and symbol length *M* estimated by fitting a periodic triangular "mask" to power autocorrelation function

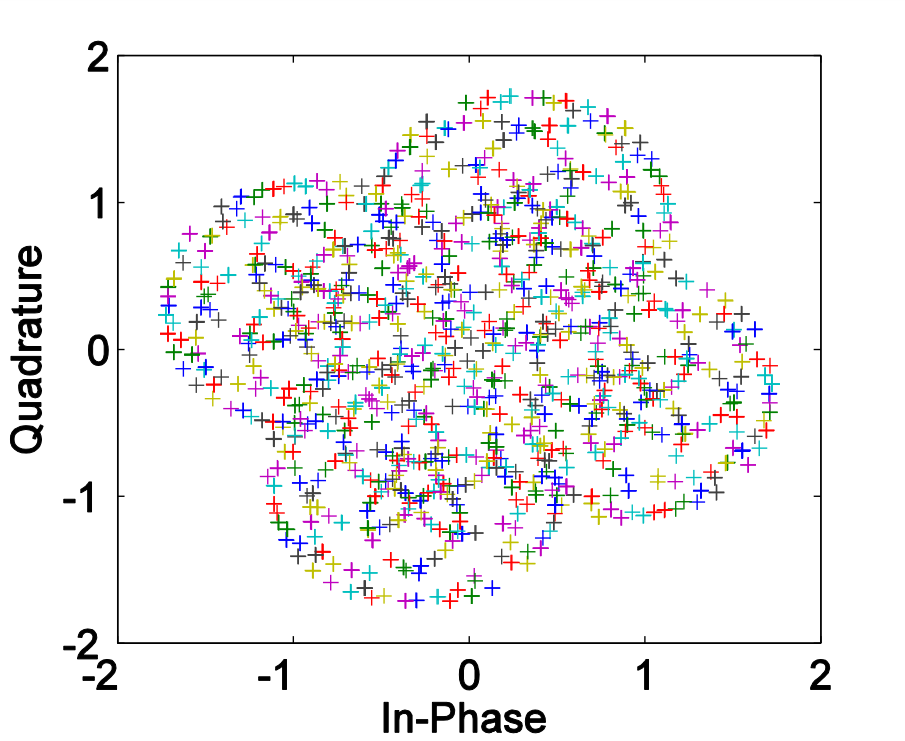


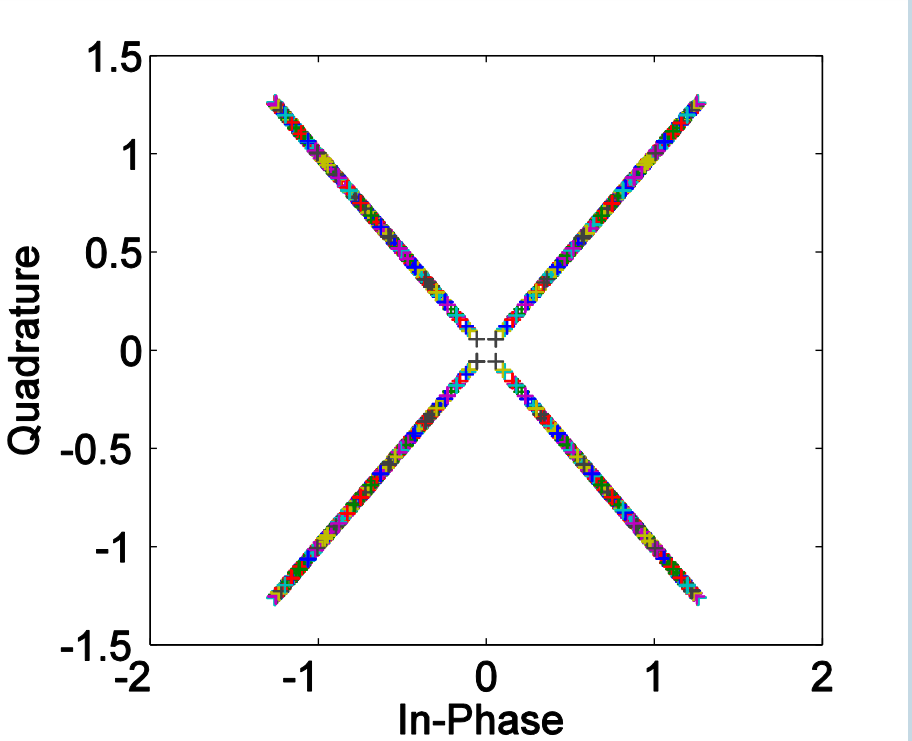


[V. Le Nir, T. van Waterschoot, J. Duplicy, and M. Moonen, "Blind CP-OFDM and ZP-OFDM parameter estimation in frequency selective channels", *EURASIP J. Wireless. Commun. Networking*, vol. 2009.

Blind demodulation: Equalization

- per-tone phase offset estimation by modulation stripping
  [Cowley, 1996] [Rezki *et al.*, 2003]





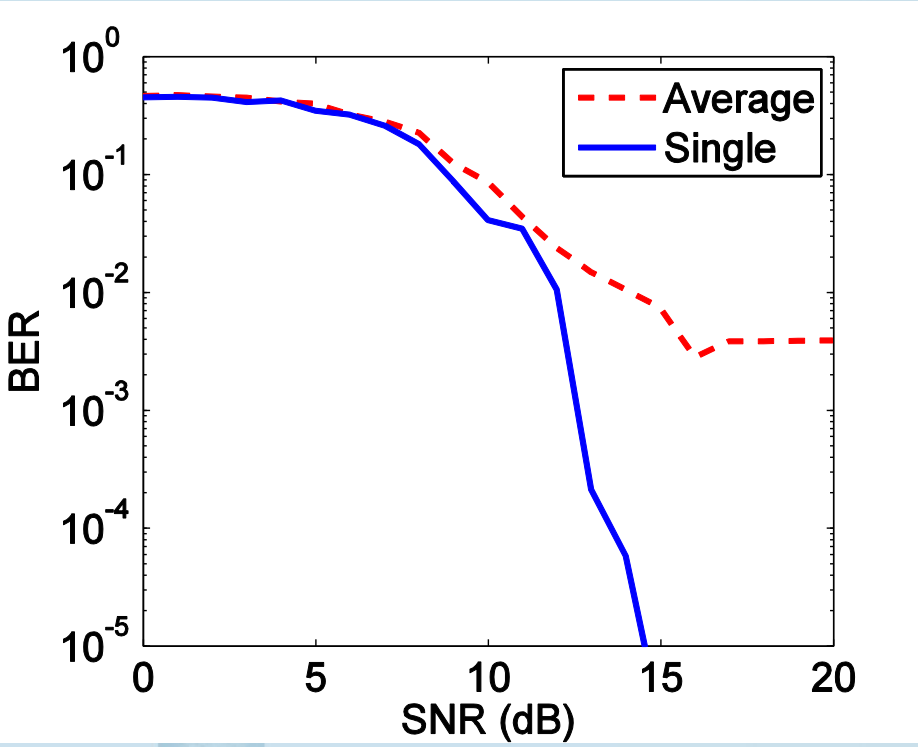
- blind channel magnitude response estimation by assuming i.i.d. of subcarrier powers at transmitter side

Blind demodulation: Subcarrier demodulation

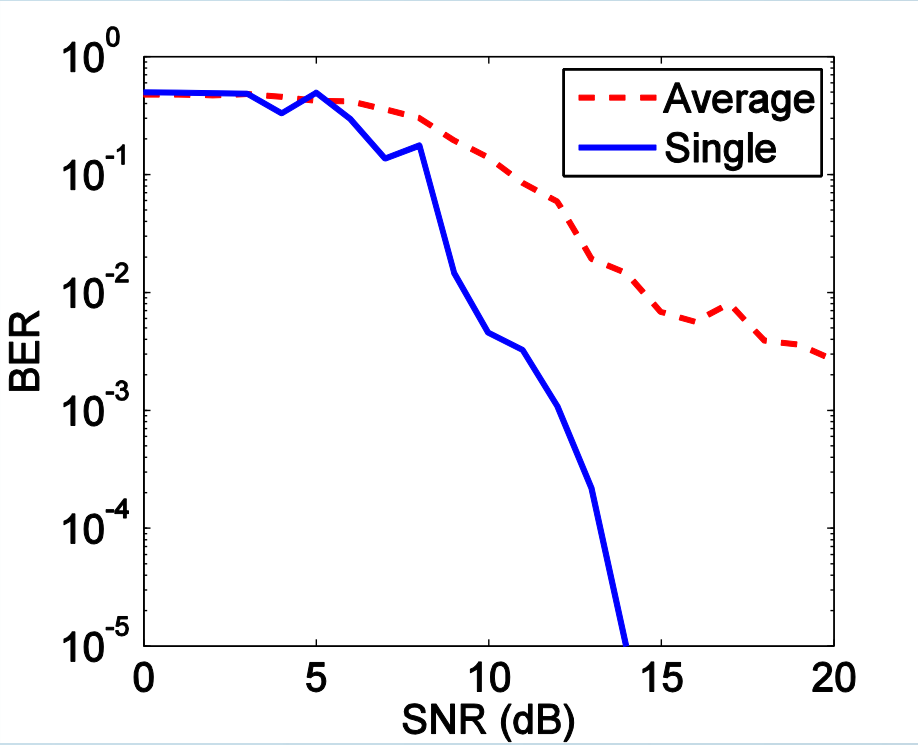
- estimation of subcarrier constellation by calculating normalized fourth-order cumulant
- example of data recovery performance:

  “WiMAX” signal on SUI-1 channel model

CP-OFDM



ZP-OFDM



Conclusion

- simple yet accurate closed-form PSD expressions for CP-OFDM and ZP-OFDM signals
- blind algorithms for OFDM detection, spectral estimation, and demodulation
- applications ranging from UWB management to blind reception and data recovery
- promising estimation and detection performance even in difficult conditions (spectral overlap, low SNR, …)
- moderate computational complexity ~ FFT complexity
- future work includes:
  - theoretical analysis of spectral estimation and detection algorithms
  - getting rid of Nyquist sampling
  - extension to other types of bandpass signals (e.g., CDMA)
  - …