

Filter Bank Multicarrier (FBMC): An Integrated Solution to Spectrum Sensing and Data Transmission in Cognitive Radio Networks

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PHYDYAS Workshop
Sunday, June 21, 2009
Hotel Brufani, Perugia, Italy

Outline



Introduction

- Cognitive radios
- Primary and Secondary Users (PUs and SUs)
- Spectrum sensing and sharing
- Spectrum leakage



FFT-based OFDM

- Subcarrier leakage
 - Limitations and Solutions



Spectrum analysis methods

- Periodogram spectral estimator (PSE)
- Multitaper method (MTM)
- Filter bank spectral estimator (FBSE)



Filter bank as a multicarrier communication tool

- Filtered multitone (FMT)
- Offset QAM/Staggered modulated multitone (SMT)
- Cosine modulated multitone (CMT)

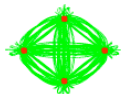


Implementation of Filterbank Multicarrier Systems

- Polyphase structures



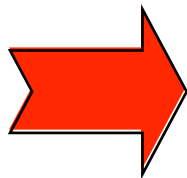
Conclusions



Introduction: What is Cognitive Radio?

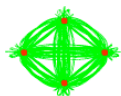
Cognitive Radios:

- Have the capability to be aware of their surrounding environment
- Can change PHY depending on environment
- Can change PHY depending on traffic needs
- Can alter higher layer behavior as needed
- Learn from past experiences



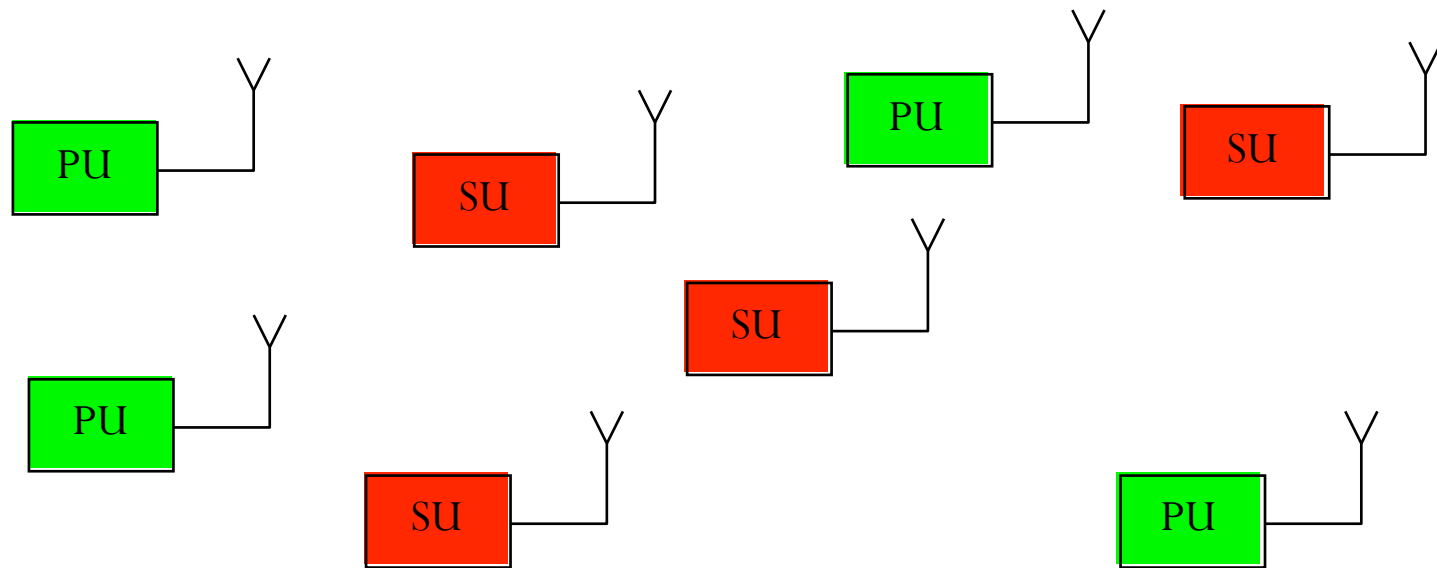
Capable of complex adaptation on lower layers

[Mit99] J. Mitola, III and G.Q. Maguire, Jr., "Cognitive Radio: Making Software Radios More Personal," *IEEE Personal Communications*, vol. 6, no. 4, 1999.



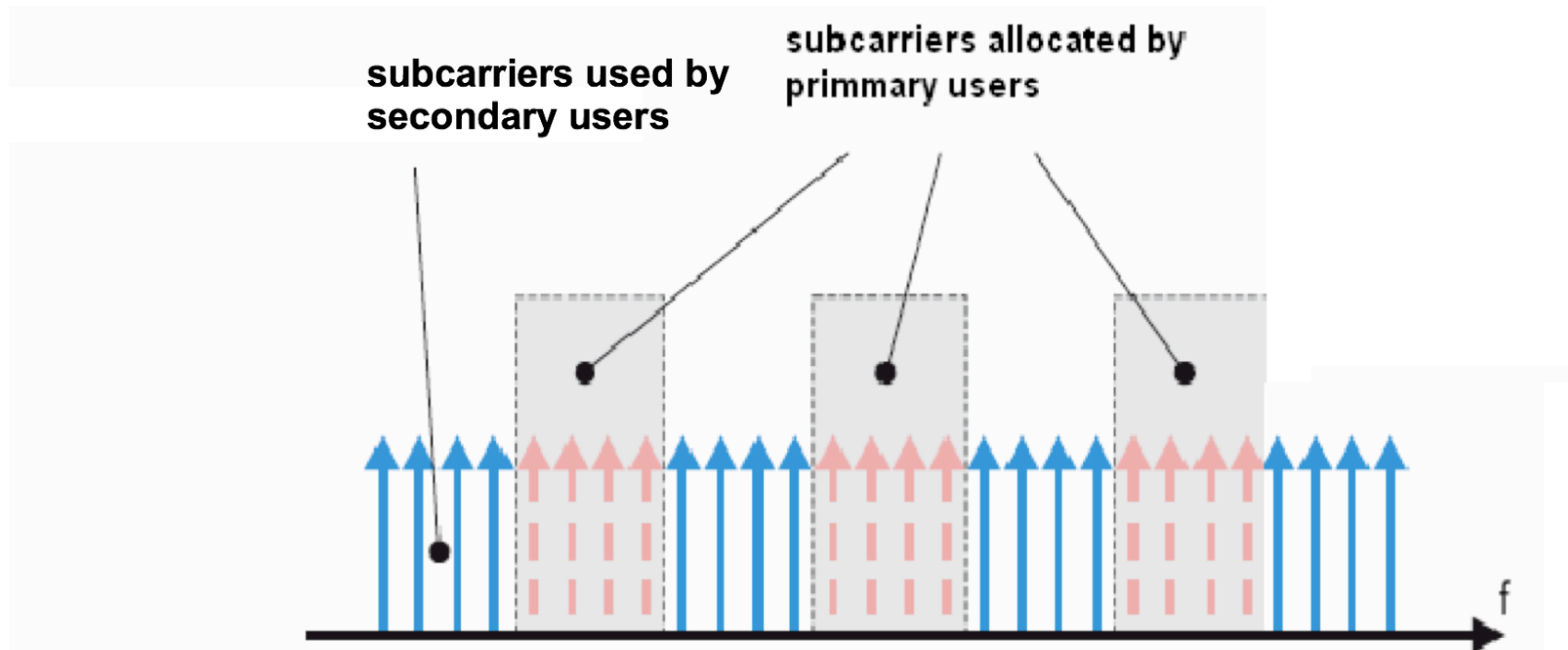
Introduction: Primary and secondary users

- ❑ Primary (licensed) and secondary (unlicensed) users coexist and share the same spectrum
- ❑ PUs have priority and thus SUs must back-off as soon as PUs begin a communication
 - This requires channel sensing

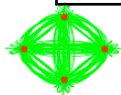


Introduction

Multicarrier has been proposed for channel sensing and co-existence of SUs with PUs

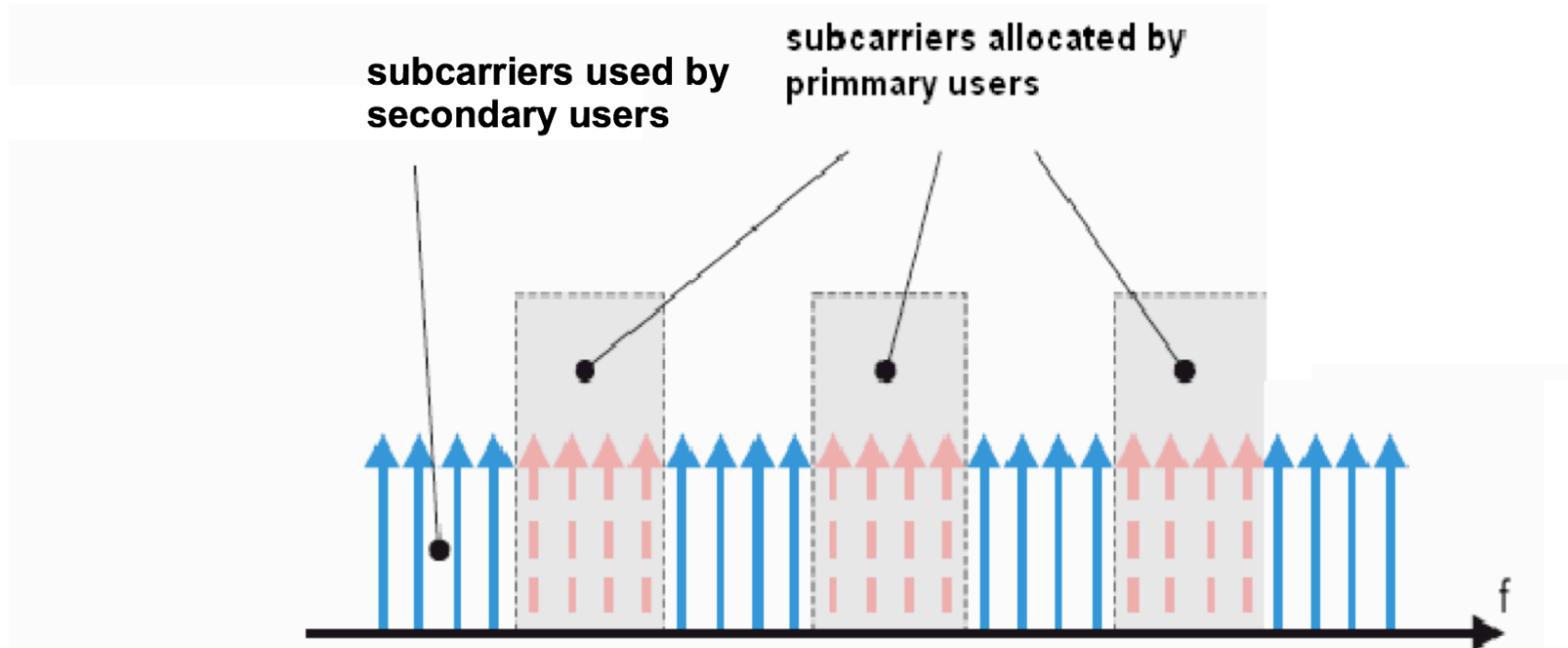


[Wei]on04] T.A. Weiss and F.K. Jondral, "Spectrum pooling: an innovative strategy for the enhancement of spectrum efficiency," *IEEE Commun. Magazine*, Vol. 42, No. 3, March 2004, pp. S8 - S14.



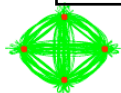
Introduction

Multicarrier has been proposed for channel sensing and co-existence of PUs and SUs



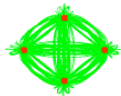
To avoid interference among primary and secondary users good separation/filtering of different subcarriers is necessary.

[Wei]on04] T.A. Weiss and F.K. Jondral, "Spectrum pooling: an innovative strategy for the enhancement of spectrum efficiency," *IEEE Commun. Magazine*, Vol. 42, No. 3, March 2004, pp. S8 - S14.

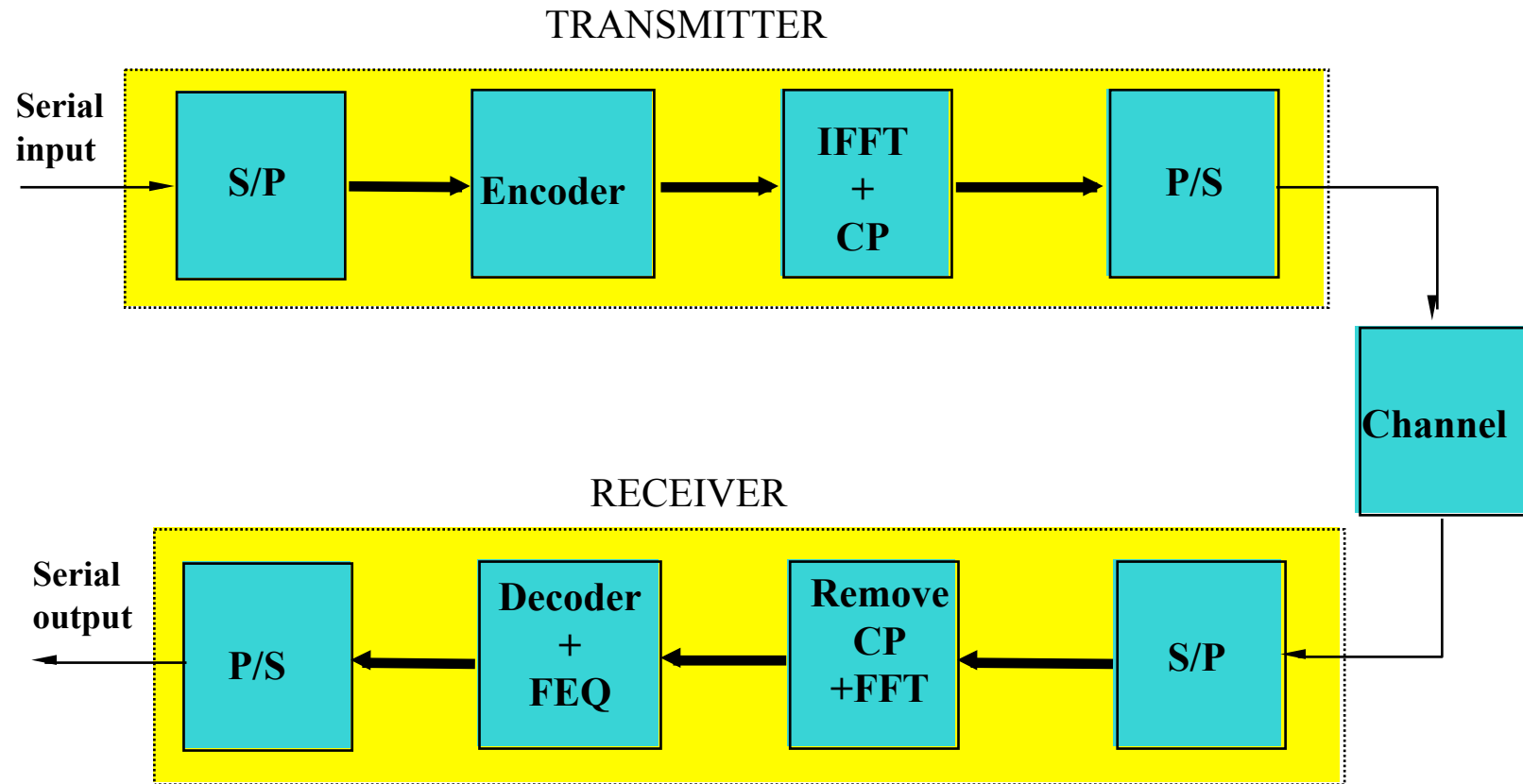


Multicarrier Methods:

Conventional OFDM



FFT-based OFDM: Transceiver Structure



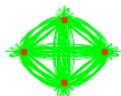
S/P: Serial-to-Parallel

P/S: Parallel-to-Serial

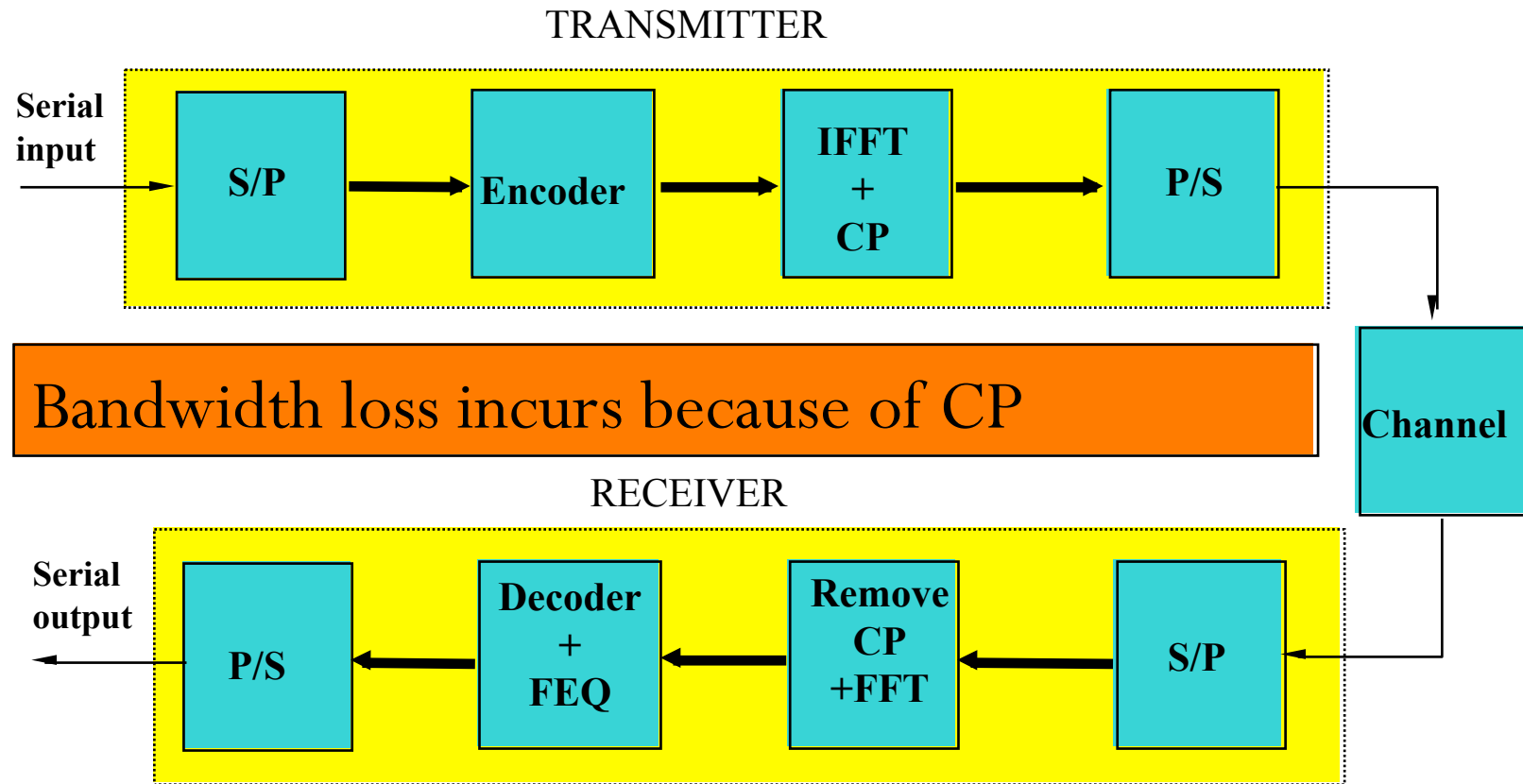
CP: Cyclic Prefix

TEQ: Time Domain Equalizer

FEQ: Frequency Domain Equalizer

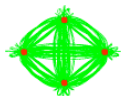


FFT-based OFDM: Transceiver Structure



S/P: Serial-to-Parallel
P/S: Parallel-to-Serial

CP: Cyclic Prefix
TEQ: Time Domain Equalizer
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FFT-based OFDM: Parameters definition

N : The maximum number of subcarriers / FFT length

C : The number of cyclic prefix samples

T_s : The sample interval (in sec.)

$T = NT_s$: The duration of each FFT block

$T_G = CT_s$: The duration of each cyclic prefix / guard interval

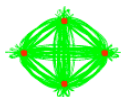
$T_S = T + T_G$: The duration of each OFDM symbol

f_i : The center frequency of the i th subcarrier

X_i : The data symbol at the i th subcarrier (i.e., in freq. domain)

$g(n)$: The symbol shaping window ($n = 0, 1, \dots, N+C-1$)

$x_i(n) = X_i g(n) e^{j2\pi(n-C-1)i/N} / N$: The i th subcarrier signal samples in time domain (before modulation to RF band); n is time index



FFT-based OFDM: Transmit signal and its spectrum

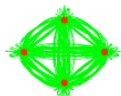
Power spectrum of the i th subcarrier:

$$\Phi_{x_i x_i}(f) = K |G(f - f_i)|^2$$

Adding up power spectra of all active subcarriers, the power spectrum of transmit signal $x(t)$ is obtained as:

$$\Phi_{xx}(f) = \sum_i \Phi_{x_i x_i}(f)$$

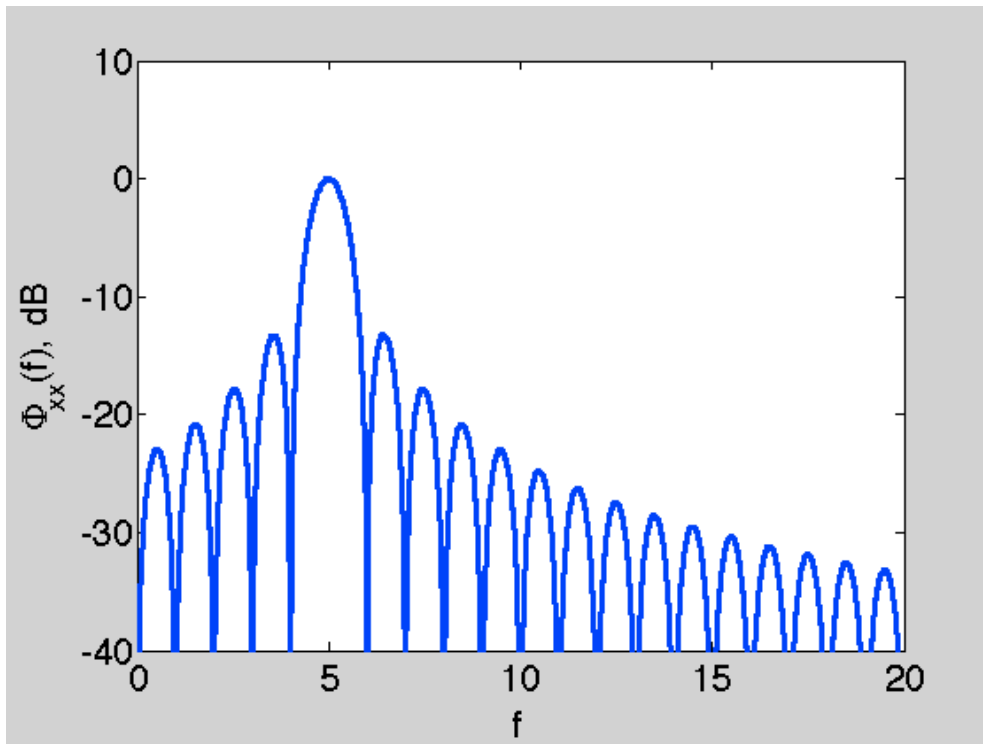
[FarKem08] B. Farhang-Boroujeny and R. Kempter, “**Multicarrier communication techniques for spectrum sensing and communication in cognitive radios**,” IEEE Commun. Magazine, April 2008.



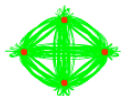
FFT-based OFDM: Transmit signal and its spectrum

In conventional OFDM, where $g(n) = 1$, for $n = 0, 1, \dots, N+C-1$,

$$\Phi_{x_i x_i}(f) = K |\text{sinc}((f - f_i)T_S)|^2, \text{ where } \text{sinc}(x) = \frac{\sin(\pi x)}{\pi x}$$

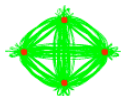
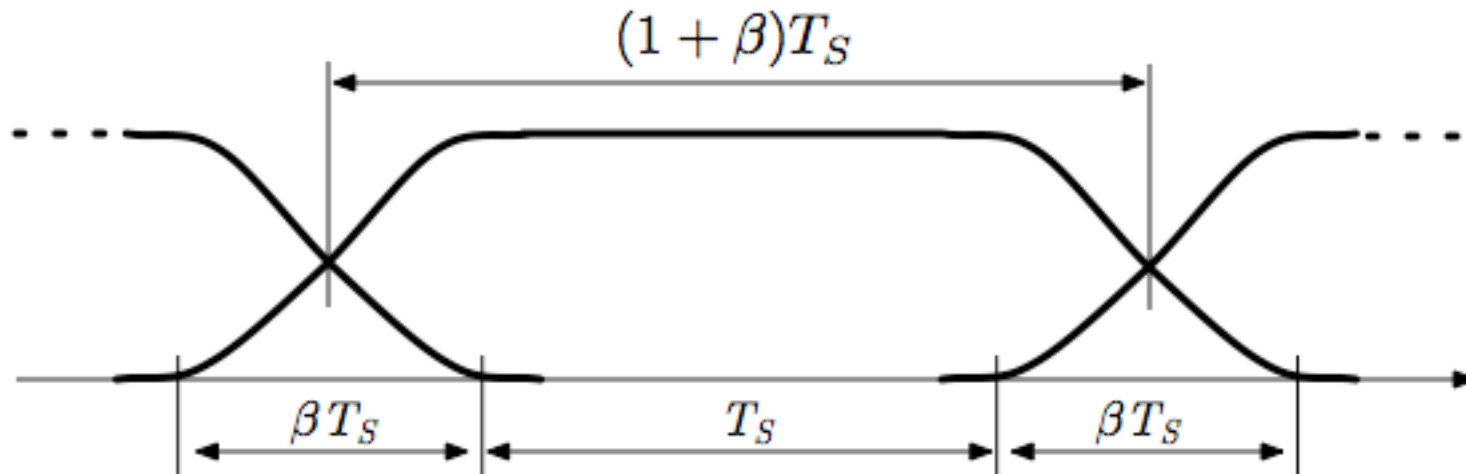


- Poor side lobes
 - interference with PUs
- First side lobe is at -13 dB!



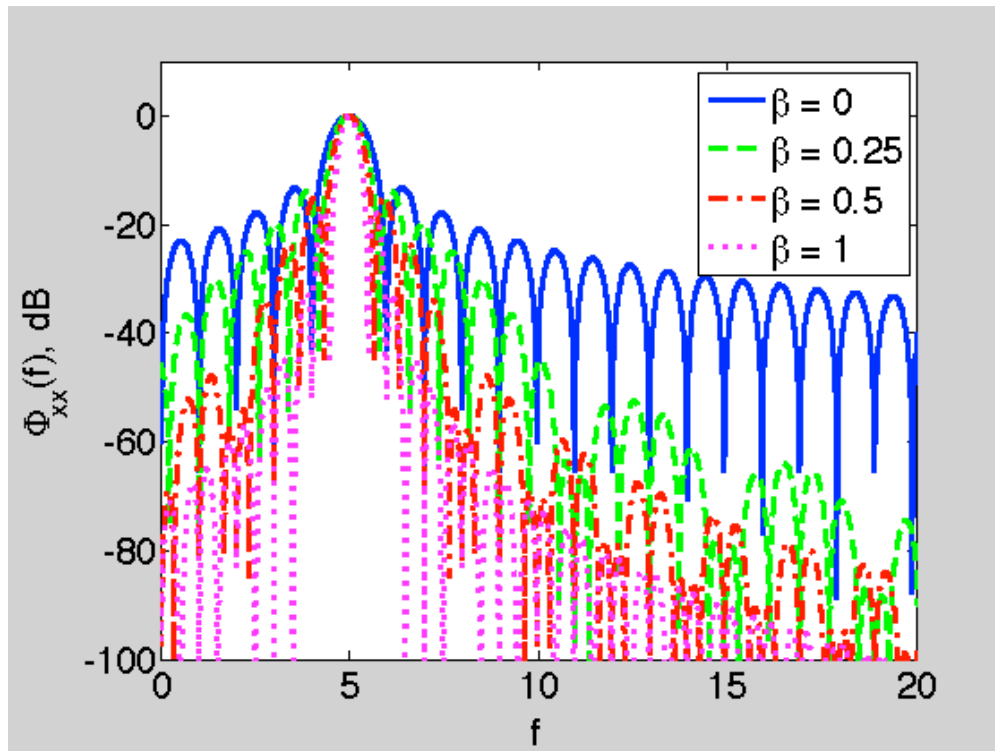
FFT-based OFDM: Improving the spectrum of transmit signal

- The large side lobes in the PSD of each subcarrier is a direct consequence of using a rectangular window
- The side lobes can be suppressed significantly by using a window that roles-off gently
- This increases the duration of each OFDM symbol from T_s to $(1+\beta)T_s$
 - Hence, further loss in bandwidth efficiency



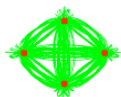
FFT-based OFDM: Improving the spectrum of transmit signal

- Side lobes are suppressed by increasing β
- To suppress side lobes sufficiently, β values of 0.5 or greater may be required (Wiess et al. (2004))
 - Hence, significant loss in bandwidth efficiency



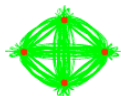
Notes:

- The side lobes adjacent to the main lobe remain significant, even for $\beta = 1$
- To solve this problem one may introduced guard bands between PU and SU bands
 - Hence, further loss in bandwidth efficiency

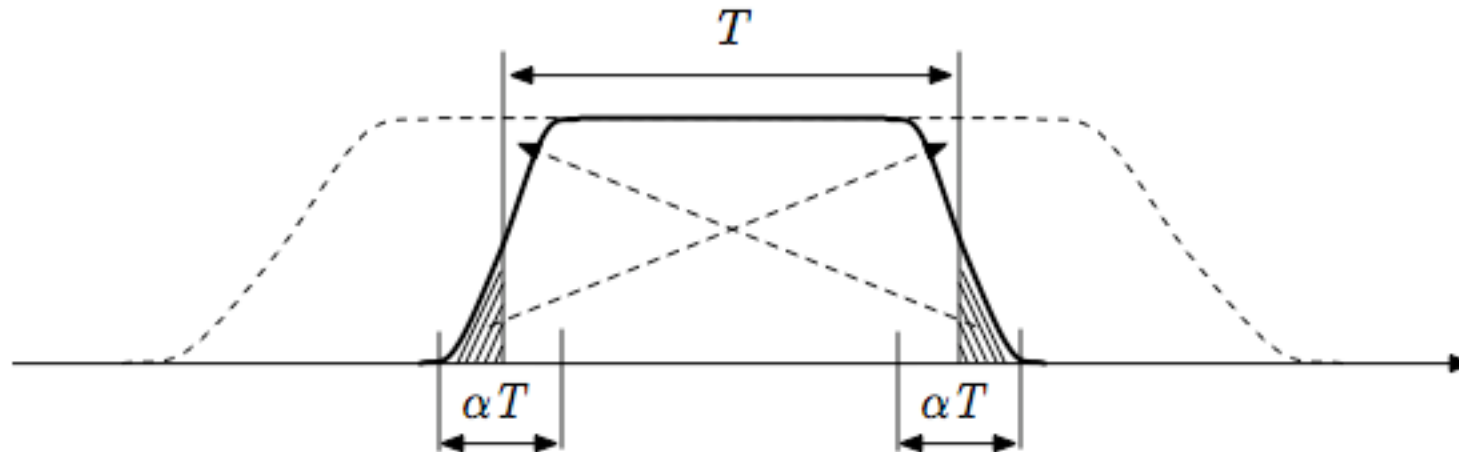


FFT-based OFDM: Interference introduced by PUs and other SUs on a SU

- Traditional OFDM applies FFT to a length N , CP-removed (rectangular) window, of the received signal.
 - ◆ This is equivalent to applying a bank of bandpass filters with modulated sinc frequency responses.
 - ◆ The large side-lobes of the sinc responses result in significant energy pick up from the bands that are unsynchronized with the intended bands.
 - Hence, significant interference will be picked up.
- **Solution:** apply a window function with gentle transition to zero, before applying FFT.
 - Further reduction in bandwidth efficiency

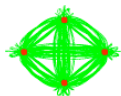


FFT-based OFDM: Reducing interference from PUs and other SUs on a SU



- Filtering is performed on $(1+\alpha)N$ samples and the result is decimated to N samples.
 - ◆ This is achieved by performing aliasing in time domain (as shown above) and then applying an N -point FFT.

[SilsNil99] F. Sjöberg, M. Isaksson, R. Nilsson, P. Odling, S.K. Wilson, and P.O. Borjesson, “**Zipper: A duplex method for VDSL based on DMT**,” *IEEE Transactions on Communications*, vol. 47, No. 8, pp. 1245-1252, Aug. 1999.



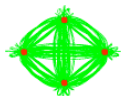
FFT-based OFDM: SUMMARY

Advantages

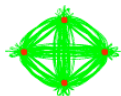
- OFDM is a well-studied method.
- OFDM chip-sets are already developed/available.
- Perfect cancellation of ISI and ICI is achieved, thanks to CP.

Disadvantages

- Hard to synchronize when subcarriers are shared among different transmitters.
- Small asynchronicity between different transmitters results in significant intercarrier interference.
- In cognitive radio, significant overhead should be added to avoid interference between primary and secondary users.



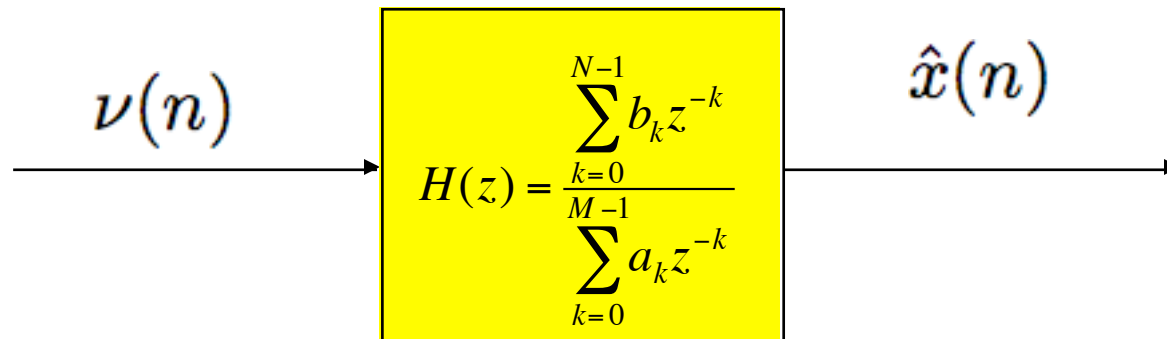
Spectral Estimation



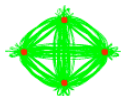
Spectral Estimation Methods: Parametric spectral estimation

The signal $x(n)$ whose spectrum is desired is treated as a random process and modeled as in the following figure.

- The input $\nu(n)$ is a white random process with variance of unity.
- The parameters a_k and b_k are optimized such that $x(n)$ and $\hat{x}(n)$ have the closet autocorrelation coefficients.



$$\begin{aligned}\Phi_{xx}(f) &= \Phi_{\nu\nu}(f) |H(e^{j2\pi f})|^2 \\ &= |H(e^{j2\pi f})|^2\end{aligned}$$



Spectral Estimation Methods: Non-parametric spectral estimation

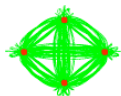
Different types of non-parametric spectral estimators:

- Periodogram Spectral Estimator (PSE)
- Blackman-Tukey Spectral Estimator (BTSE)
- Minimum Variance Spectral Estimator (MVSE)
- Multitaper Method (MTM)
- Filter Bank Spectral Estimator (FBSE)

[LimOpp88] J.S. Lim and A.V. Oppenheim (1988), **Advanced Topics in Signal Processing**. Prentice Hall, Englewood Cliffs, New Jersey, 1988.

[Thom82] D.J. Thomson (1982), “**Spectrum estimation and harmonic analysis**,” *Proceedings of the IEEE*, vol. 70, no. 9, pp. 1055-1096, Sept. 1982.

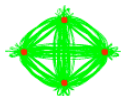
[Hay05] S. Haykin (2005), “**Cognitive radio: brain-empowered wireless communications**,” *IEEE Journal Selected Areas in Communications*, vol. 23, no. 3, pp. 201-220, Feb. 2005.



Spectral Estimation Methods: Non-parametric spectral estimation

Periodogram Spectral Estimator (PSE): obtains the spectrum of a random process $x(n)$, based on the observed samples $\{x(n), x(n-1), x(n-2), \dots, x(n-N+1)\}$, by evaluating the amplitude of the DFT of the observed vector

$$\mathbf{x}(n) = \begin{bmatrix} x(n) \\ x(n-1) \\ x(n-2) \\ \vdots \\ x(n-N+1) \end{bmatrix}$$



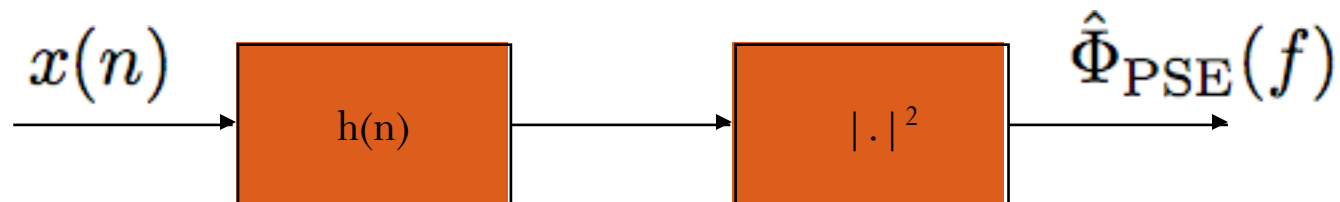
Spectral Estimation Methods: Non-parametric spectral estimation

PSE as a filterbank: The process of applying IDFT to the observed vector $\mathbf{x}(n)$, may be also formulated as

$$\hat{\Phi}_{\text{PSE}}(f) = \left| \sum_{k=0}^{N-1} h(k)x(n-k) \right|^2$$

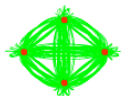
where

$$h(k) = \frac{1}{\sqrt{N}} e^{j2\pi f k}$$



For IDFT we have a bank of filters centered at

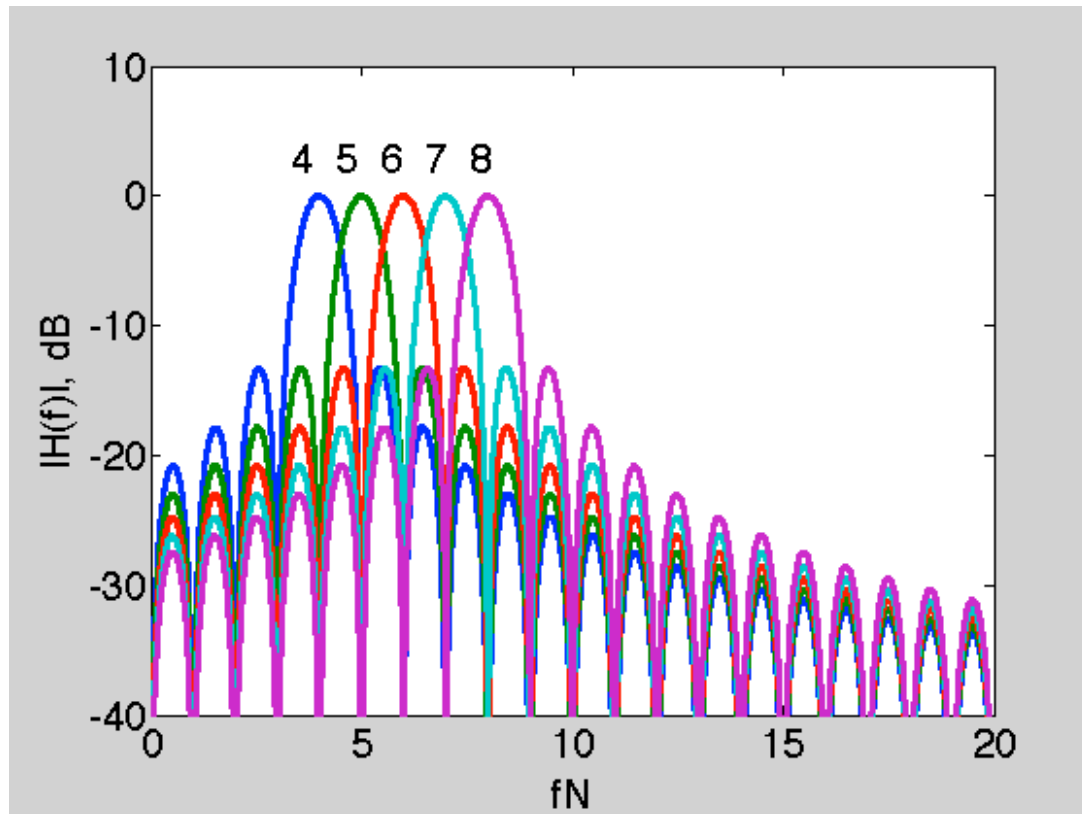
$$f = 0, \frac{1}{N}, \frac{2}{N}, \dots, \frac{N-1}{N}$$



Spectral Estimation Methods: Non-parametric spectral estimation

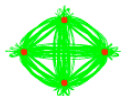
DFT filters:

$$|H(f)| = \frac{1}{\sqrt{N}} \left| \frac{\sin(N\pi(f - f_i))}{\sin(\pi(f - f_i))} \right| \approx \sqrt{N} |\text{sinc}(N(f - f_i))|$$



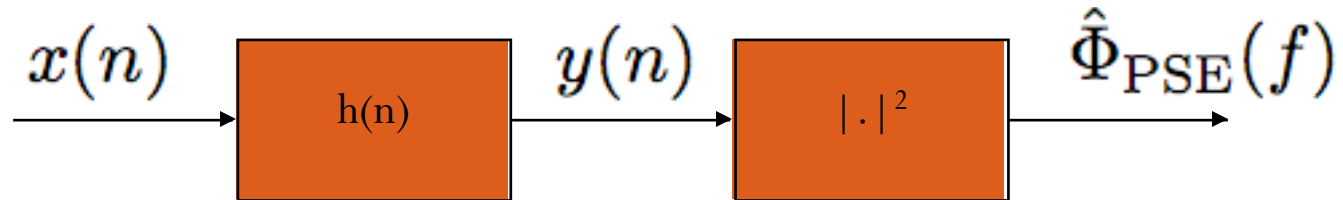
Relatively large side-lobes result in significant leakage of spectral power among different bands.

- Hence, reduces the spectral dynamic range.



Spectral Estimation Methods: Non-parametric spectral estimation

Resolution of the estimates of PSD in PSE:



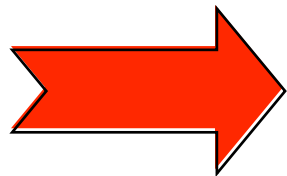
Since $x(n)$ is a random process, the filter output $y(n)$ is a random variable.

Moreover, $y(n)$, in general, can be approximated by a Gaussian, since it is constructed by linearly combining (a large) set of samples of $x(n)$. Accordingly, $|y(n)|^2$ has a chi-square distribution with 2 degree of freedom, viz.,

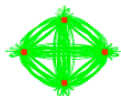
$$\hat{\Phi}_{\text{PSE}}(f) \sim \chi_2^2$$

Thus,

$$\text{VAR}[\hat{\Phi}_{\text{PSE}}(f)] = 2E^2[\hat{\Phi}_{\text{PSE}}(f)]$$



Because of their large variance,
the PSD estimates are **not reliable**.



Spectral Estimation Methods: Non-parametric spectral estimation

Prototype filter of the DFT filterbank:

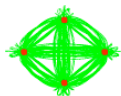
- The prototype filter in DFT has the coefficient vector

$$h_0(n) = \frac{1}{\sqrt{N}}, \quad \text{for } n = 0, 1, \dots, N-1$$

- The i th-band filter in DFT has the coefficient vector

$$h_i(n) = h_0(n)e^{j2\pi in/N} = \frac{1}{\sqrt{N}}e^{j2\pi in/N}$$

- The prototype filter is a lowpass filter and the i th-band filter is obtained by modulating the prototype filter.
- The prototype filter is also the 0th-band filter in the filterbank.



Spectral Estimation Methods: Non-parametric spectral estimation

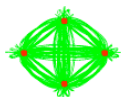
Matrix formulation the DFT filterbank:

$$\begin{bmatrix} X_n(0) \\ X_n(1) \\ X_n(2) \\ \vdots \\ X_n(N-1) \end{bmatrix} = \begin{bmatrix} 1 & 1 & 1 & \dots & 1 \\ 1 & e^{j2\pi/N} & e^{j4\pi/N} & \dots & e^{j2\pi(N-1)/N} \\ 1 & e^{j4\pi/N} & e^{j8\pi/N} & \dots & e^{j4\pi(N-1)/N} \\ \vdots & \vdots & \vdots & \dots & \vdots \\ 1 & e^{j2\pi(N-1)/N} & e^{j4\pi(N-1)/N} & \dots & e^{j2\pi(N-1)^2/N} \end{bmatrix}$$

IDFT matrix

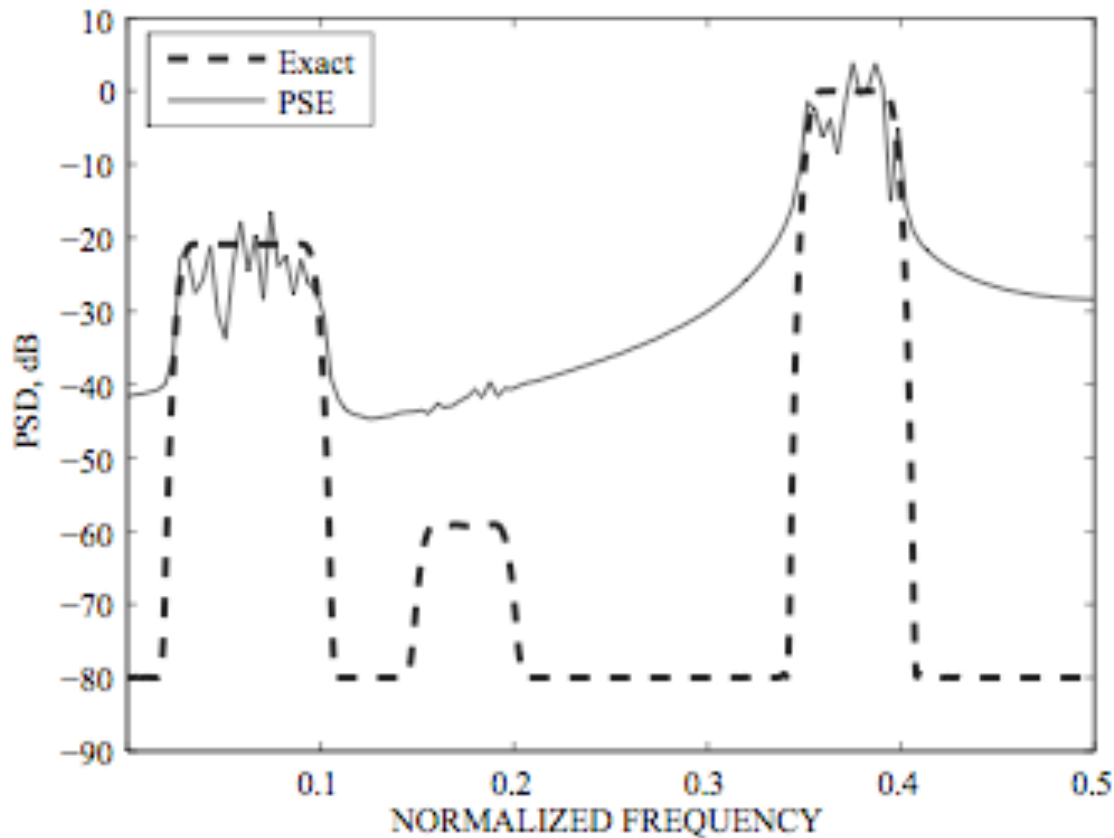
$$\times \left(\underbrace{\begin{bmatrix} \frac{1}{\sqrt{N}} \\ \frac{1}{\sqrt{N}} \\ \frac{1}{\sqrt{N}} \\ \vdots \\ \frac{1}{\sqrt{N}} \end{bmatrix}}_{\text{rectangular window / prototype filter coefficients}} \cdot \begin{bmatrix} x(n) \\ x(n-1) \\ x(n-2) \\ \vdots \\ x(n-N+1) \end{bmatrix} \right)$$

May be viewed as a rectangular window / prototype filter coefficients that are applied to input samples before taking the Fourier transform.



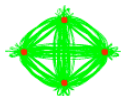
Spectral Estimation Methods: Non-parametric spectral estimation

A snapshot of PSE:



Notes:

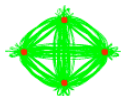
- DFT size $N = 256$.
- The expected large variance of the estimates is seen.
- Small spectral dynamic range of PSE is a direct consequence of large side lobes in DFT filters.



Spectral Estimation Methods: Non-parametric spectral estimation

Blackman-Tukey Spectral Estimator (BTSE):

- The large side lobes in PSE arises because of the use of a poor prototype filter.
 - The problem can be resolved by improving the prototype filter, equivalent to applying a window function before filtering.
 - Window function can be directly applied to the signal samples or to the auto-correlation coefficients of the signal samples.
 - When window function is applied to the auto-correlation coefficients of the signal samples, the resulting method called BTSE.



Spectral Estimation Methods: Non-parametric spectral estimation

Blackman-Tukey Spectral Estimator (BTSE): Examples of common windows (length input signal sample = $M+1$, $N = 2M+1$)

- Rectangular:

$$w(k) = \begin{cases} \frac{1}{\sqrt{2M+1}}, & |k| \leq M \\ 0, & |k| > M \end{cases} \quad W(\omega) = W_{\text{rect}}(\omega) = \frac{\sin \frac{\omega}{2}(2M+1)}{\sqrt{2M+1} \sin \frac{\omega}{2}}$$

- Hanning:

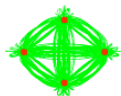
$$w(k) = \begin{cases} \frac{1}{2\sqrt{2M+1}}(1 + \cos \frac{\pi k}{M}), & |k| \leq M \\ 0, & |k| > M \end{cases}$$

- Hamming:

$$W(\omega) = \frac{1}{4}W_{\text{rect}}(\omega - \frac{\pi}{M}) + \frac{1}{2}W_{\text{rect}}(\omega) + \frac{1}{4}W_{\text{rect}}(\omega + \frac{\pi}{M})$$

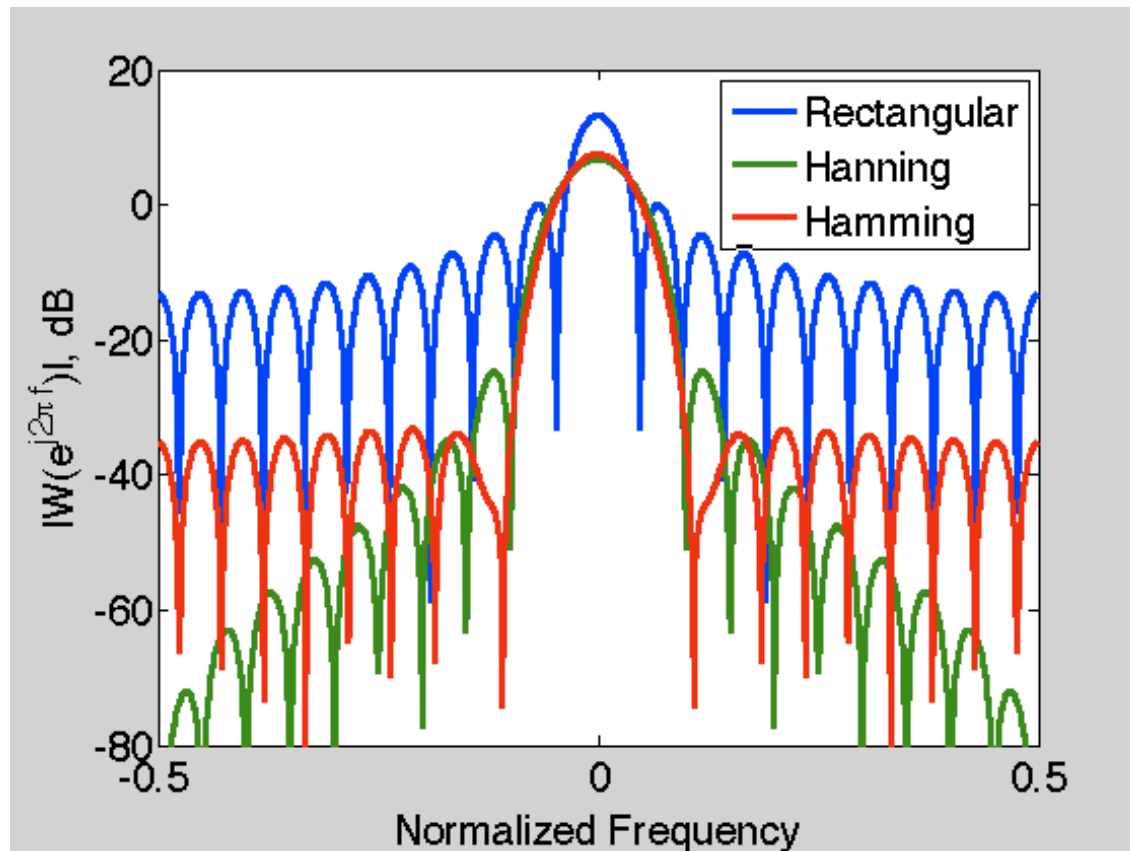
$$w(k) = \begin{cases} \frac{1}{\sqrt{2M+1}}(0.54 + 0.46 \cos \frac{\pi k}{M}), & |k| \leq M \\ 0, & |k| > M \end{cases}$$

$$W(\omega) = 0.23W_{\text{rect}}(\omega - \frac{\pi}{M}) + 0.54W_{\text{rect}}(\omega) + 0.23W_{\text{rect}}(\omega + \frac{\pi}{M})$$



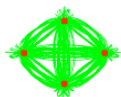
Spectral Estimation Methods: Non-parametric spectral estimation

Blackman-Tukey Spectral Estimator (BTSE): Examples of common windows



Note: side lobes have decreased at the cost of wider main lobe.

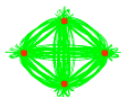
➤ Reduction of Leakage among different bands, i.e., increased spectral dynamic range, is traded at the cost of a lower resolution in frequency.



Spectral Estimation Methods: Non-parametric spectral estimation

Minimum Variance Spectral Estimator (MVSE):

- Each point of PSD is estimated using a different filter.
- These filters are adopted to the spectrum whose estimate is desired.
- Each filter is selected to have a gain of unity at the center of the passband, while the side lobes are optimized for minimum leakage of energy from other bands.
- **We do not explore this method as a good candidate for spectrum sensing in CR, mostly because of its computational complexity.**

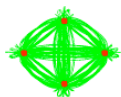


Spectral Estimation Methods: Non-parametric spectral estimation

Multitaper Method (MTM):

- This method replaces the single (prototype) filter in the previous methods by a few filters for measurement of each point of PSD.
- All filters have the same passband. However, by design they are orthogonal, hence, their outputs are a set of uncorrelated random variables.
- The output power of the filters are averaged to reduce the variance of the estimates.
- A set of prototype filters are used for all the bands and polyphase architecture is used for efficient implementation.
- The prototype filters are a set of **prolate filters** that satisfy some desirable properties/optimality conditions, as discussed in the next slide.

[Thom82] D.J. Thomson (1982), “**Spectrum estimation and harmonic analysis**,” *Proceedings of the IEEE*, vol. 70, no. 9, pp. 1055-1096, Sept. 1982.



Spectral Estimation Methods: Non-parametric spectral estimation

Origin of the Prolate Filters: Slepian Sequences

- The Slepian sequences $\mathbf{s}_k = [s_k(1), s_k(2), \dots, s_k(N)]^T$, $k=1, 2, \dots, K$ constitute a set of K unit length orthogonal bases vectors which are used to obtain an optimal expansion of the time sequence

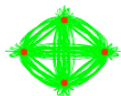
$$\mathbf{x}(n) = [x(n-N+1), x(n-N+2), \dots, x(n-1), x(n)]^T$$

over the frequency band $[f_i - \Delta f/2, f_i + \Delta f/2]$.

- The expansion of $\mathbf{x}(n)$ has the form of

$$\mathbf{x}(n) = \mathbf{K}_1 \mathbf{s}_1 + \mathbf{K}_2 \mathbf{s}_2 + \dots + \mathbf{K}_K \mathbf{s}_K, \quad \mathbf{K}_k = \mathbf{s}_k^H \mathbf{x}(n)$$

- To maximize the accuracy of the estimate, the Slepian sequences are chosen such that their spectrum is maximally concentrated over the desired band $[f_i - \Delta f/2, f_i + \Delta f/2]$. This can be related to the *minimax theorem*.



Spectral Estimation Methods: Non-parametric spectral estimation

Minimax Theorem:

The distinct eigenvalues $\lambda_1 > \lambda_2 > \dots > \lambda_N$ of the correlation matrix \mathbf{R} of an observation vector $\mathbf{x}(n)$, and their corresponding eigenvectors, $\mathbf{q}_1, \mathbf{q}_2, \dots, \mathbf{q}_N$, may be obtained through the following optimization procedure:

$$\lambda_1 = \max E[|\mathbf{q}_1^H \mathbf{x}(n)|^2], \text{ subject to } \mathbf{q}_1^H \mathbf{q}_1 = 1$$

and for $k = 2, 3, \dots, N$

$$\lambda_k = \max E[|\mathbf{q}_k^H \mathbf{x}(n)|^2], \text{ subject to } \mathbf{q}_k^H \mathbf{q}_k = 1 \text{ and } \mathbf{q}_k^H \mathbf{q}_i = 0, \text{ for } i=1, 2, \dots, k-1$$

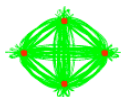
Alternatively, the following procedure may also be used:

$$\lambda_N = \min E[|\mathbf{q}_N^H \mathbf{x}(n)|^2], \text{ subject to } \mathbf{q}_N^H \mathbf{q}_N = 1$$

and for $k = N-1, \dots, 2, 1$

$$\lambda_k = \min E[|\mathbf{q}_k^H \mathbf{x}(n)|^2], \text{ subject to } \mathbf{q}_k^H \mathbf{q}_k = 1 \text{ and } \mathbf{q}_k^H \mathbf{q}_i = 0, \text{ for } i=N, N-1, \dots, k+1$$

[Far98] B. Farhang-Boroujeny, **Adaptive Filters: Theory and Applications**. John Wiley & Sons, 1998.



Spectral Estimation Methods: Non-parametric spectral estimation

Prolate Filters Design:

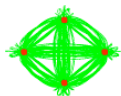
1. Construct the correlation matrix \mathbf{R} of a random process with the power spectral density

$$\Phi_{xx}(f) = \begin{cases} 1, & |f| < \frac{\Delta f}{2} \\ 0, & \text{otherwise} \end{cases}$$

\mathbf{R} is a Toeplitz matrix with the first row of

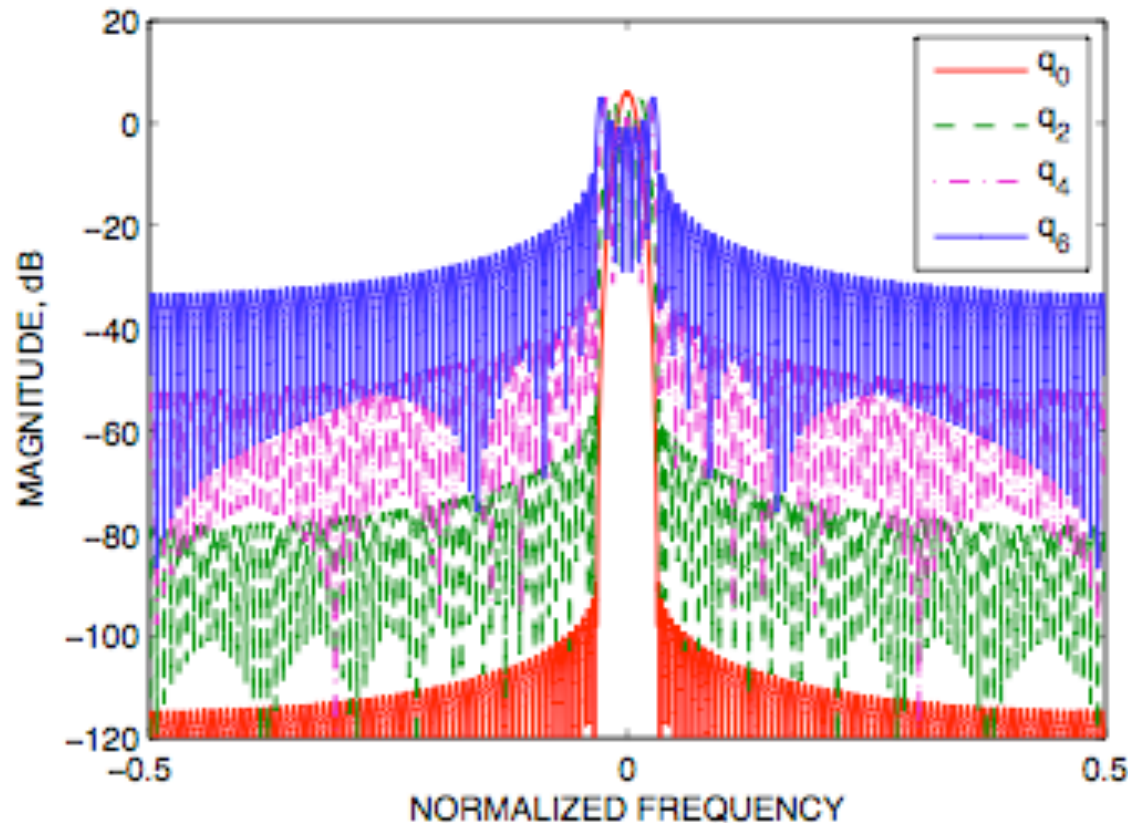
$$[\phi(0) \ \phi(1) \ \cdots, \ \phi(N-1)], \text{ with } \phi(n) = \Delta f \text{sinc}(\Delta f n)$$

2. The first K eigenfilters corresponding to the largest eigenvalues of \mathbf{R} are the coefficient vectors of the prolate filters.



Spectral Estimation Methods: Non-parametric spectral estimation

An Example of Prolate Filters:

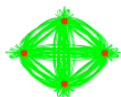


Filter parameters:

- Number of subbands: $N=16$
- Filter length: $L=8N=128$

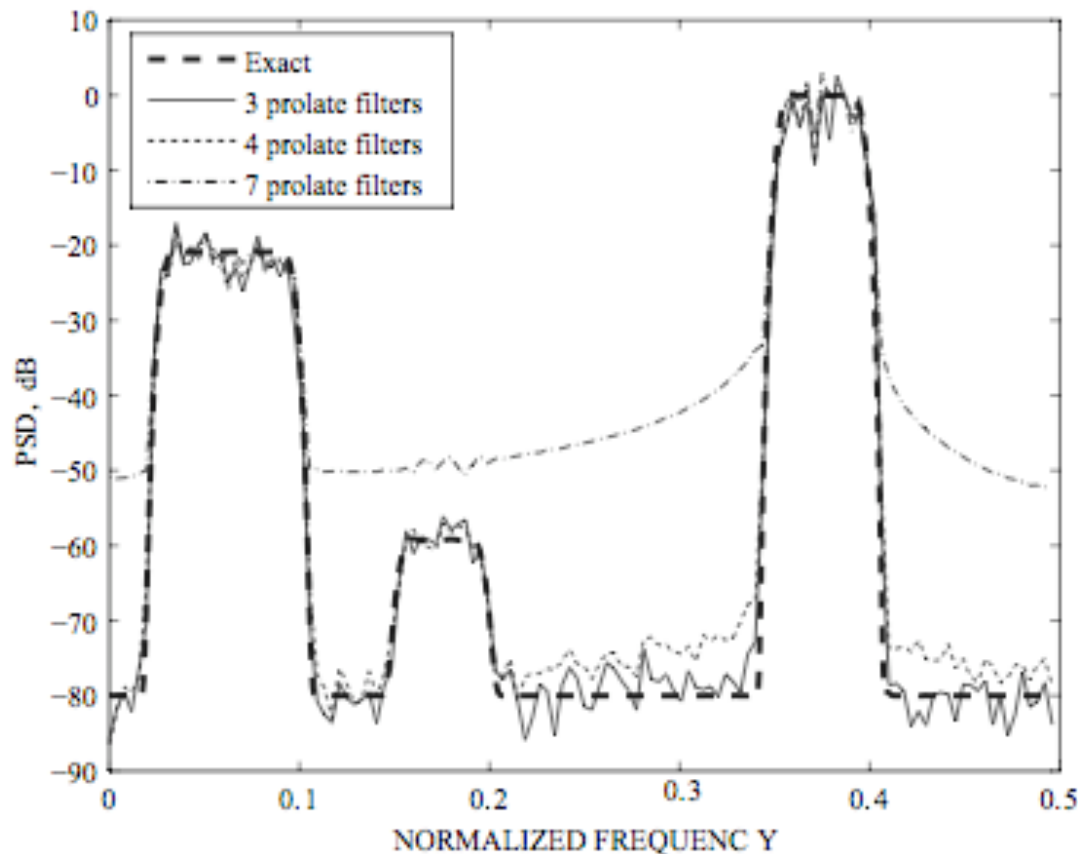
Observations:

- Only the first few filters have good stopband attenuation.
 - Hence, to have a good spectral dynamic range, only the outputs of the first few prolate filters can be used.



Spectral Estimation Methods: Non-parametric spectral estimation

An Example of Spectrum Sensing Using Prolate Filters:

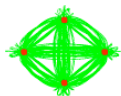


Filter parameters:

- Number of subbands: $N=128$
- Filter length: $L=8N=1024$

Observation:

- As predicted, to have a good spectral dynamic range (say, 60 dB or better), not more than three prolate filters should be used.



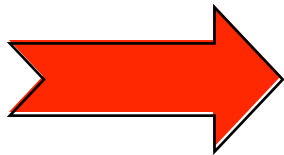
Spectral Estimation Methods: Non-parametric spectral estimation

Adaptive MTM: Thompson proposed the following formulae in order to reduce the impact of poor side lobes of higher numbered prolate filters:

$$\hat{S}(f) = \frac{\sum_{k=0}^{K-1} |d_k(f)|^2 \hat{S}_k(f)}{\sum_{k=0}^{K-1} |d_k(f)|^2} \quad \text{and} \quad d_k(f) = \frac{\sqrt{\lambda_k} S(f)}{\lambda_k S(f) + B_k(f)}$$

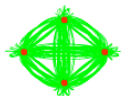
Leakage power from other
bands

Starting with coarse estimate of $S(f)$, $d_k(f)$ is estimated and used to improve the estimate of $S(f)$. Iterations continue until $S(f)$ converges.



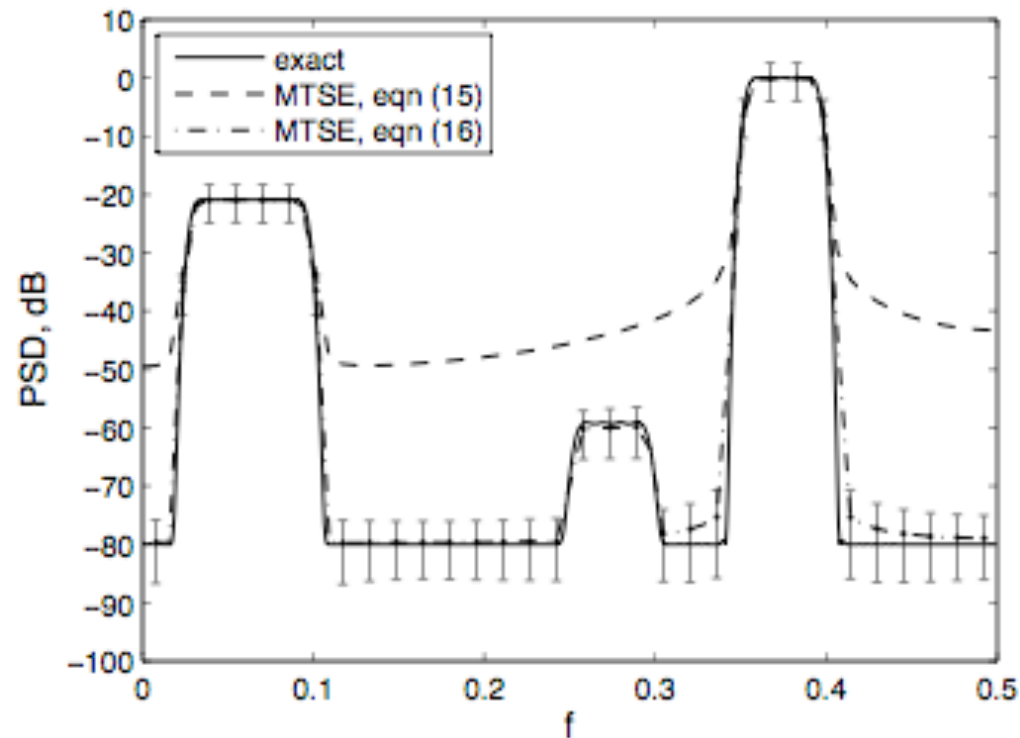
A computationally intensive procedure!

[Thom82] D.J. Thomson (1982), “**Spectrum estimation and harmonic analysis**,” *Proceedings of the IEEE*, vol. 70, no. 9, pp. 1055-1096, Sept. 1982.

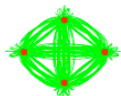


Spectral Estimation Methods: Non-parametric spectral estimation

An Sample result of adaptive MTM:

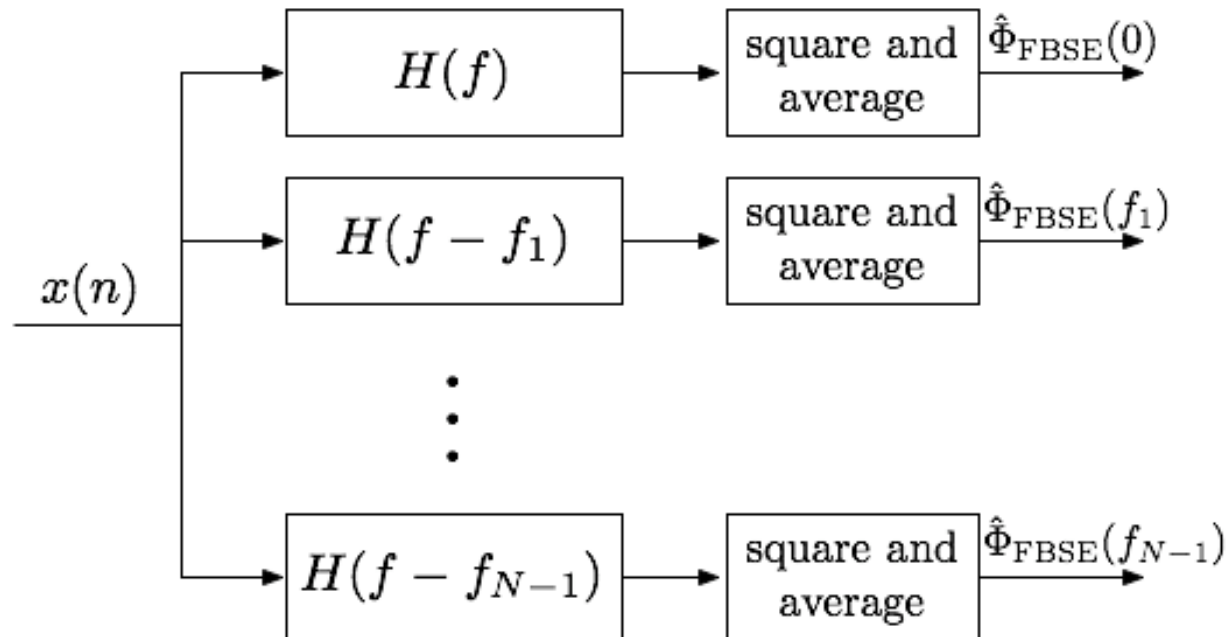


- Results are average of 10,000 snapshots.
- Vertical lines indicate 95% confidence intervals.
- Variances are larger at lower levels of PSD.



Spectral Estimation Methods: Non-parametric spectral estimation

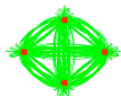
Filter Bank Spectral Estimator (FBSE):



Notes:

- $H(f)$, known as prototype filter, is centered around $f = 0$.
- The rest of the filters are frequency-shifted/modulated copies of $H(f)$.

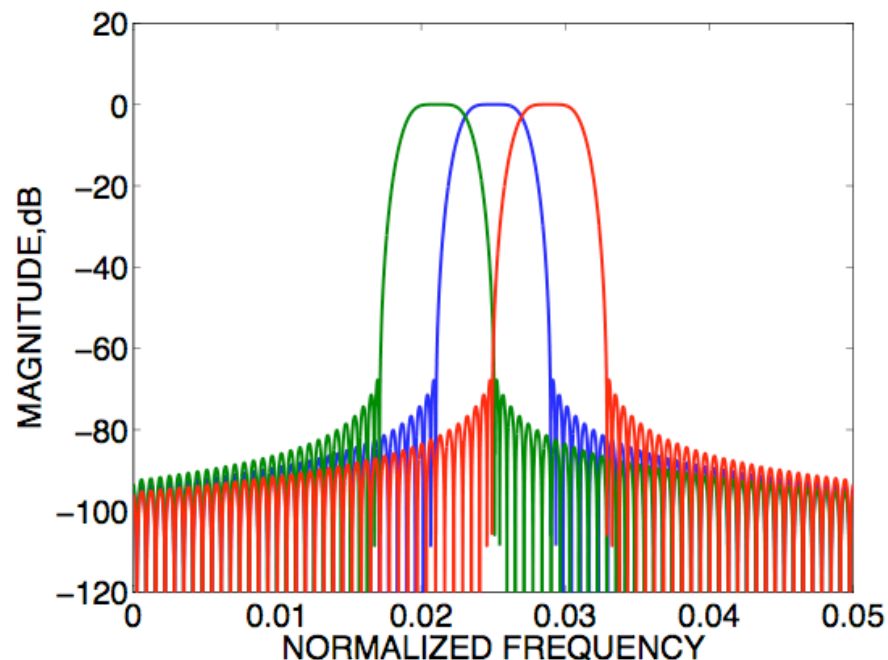
[Far08] B. Farhang-Boroujeny, "Filter bank spectrum sensing for cognitive radios," IEEE Trans. On Signal Processing, May 2008.



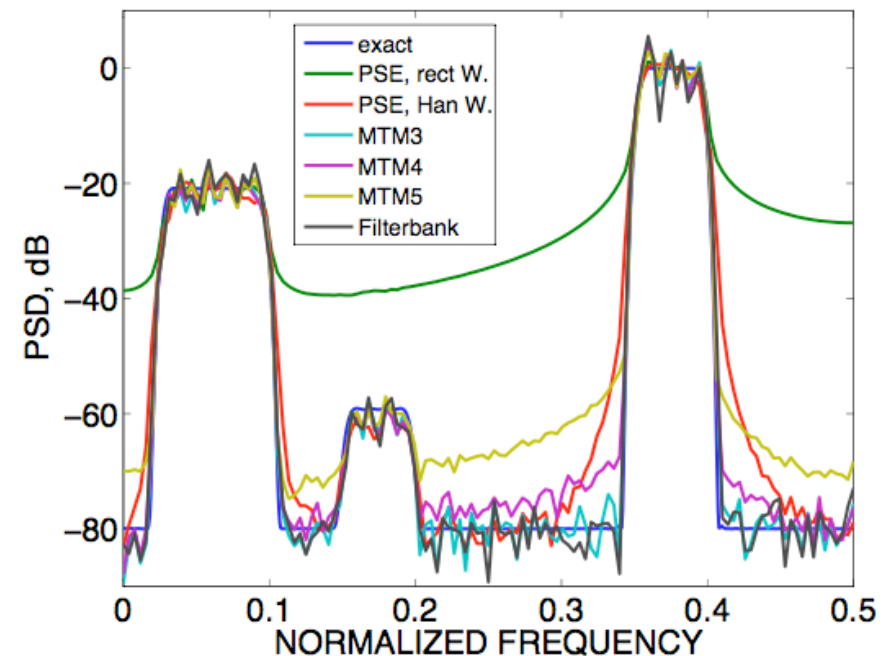
Spectral Estimation Methods: Non-parametric spectral estimation

An Example of Spectrum Sensing Using FBSE:

Magnitude responses of three bandpass filters in a filter bank

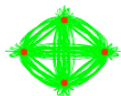


Snapshots of FBSE and other techniques



Filter parameters:

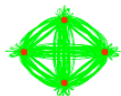
- Number of subbands: $N=256$; Filter length: $L=6N=1536$;
- The data length for FBSE is $8N = 2048$; Output power of each subband is measured by averaging over three samples at spacing 256.
- The data lengths for PSE and MTM are 256 and 1024 samples, respectively.



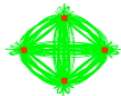
Spectral Estimation Methods: SUMMARY

CONCLUSIONS:

- Periodogram spectral estimation may be insufficient to achieve the required spectral dynamic range.
- The multitaper method gives excellent results, with limited length of data. However, it is too complex to implement.
- Filter bank spectral estimation gives excellent result. Yet, it may cost very little if filter bank multicarrier is adopted for signal modulation.

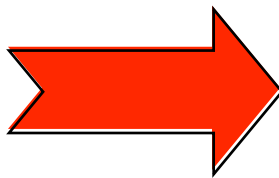


Filter Bank Multicarrier (FBMC) Methods

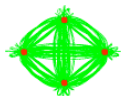


Filter Bank Multicarrier (FBMC) Methods

- **Filtered Multitone (FMT):** Uses subcarrier bands with no overlap. Data symbols are quadrature amplitude modulated (QAM).
 - Guard bands are used to separate subcarrier bands. **This results in some loss in bandwidth efficiency**
- **Multicarrier with Offset QAM/Staggered Modulated Multitone (SMT):** Subcarrier bands are maximally overlapped/minimally spaced.
 - **Carrier spacing = symbol rate**
- **Cosine Modulated Multitone (CMT):** Uses pulse amplitude modulated (PAM) symbols with vestigial sideband modulation. Subcarrier bands are maximally overlapped / minimally spaced.
 - **Carrier spacing = one half of symbol rate**



Both SMT and CMT achieve
maximum bandwidth efficiency



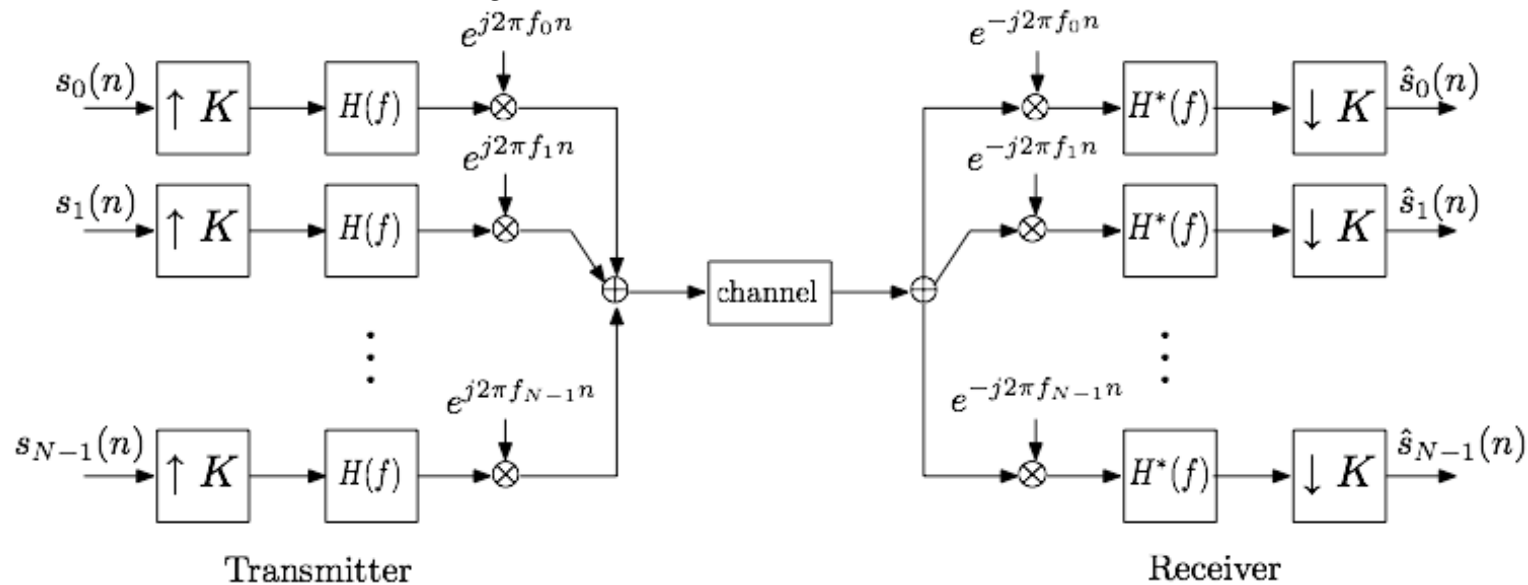
Filter Bank Multicarrier (FBMC) Methods: FMT (Summary)

- FMT follows the simple principles of the conventional frequency division multiplexing (FDM).

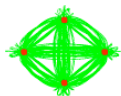
➤ Subcarrier bands have no overlap.



- To allow an efficient implementation based on polyphase structures, (i) some restrictions on the position of subcarrier bands are imposed; (ii) a prototype filter is used for all subcarrier bands.
- Transmit symbols, in general, are QAM and $H(f)$ and $H^*(f)$ are a pair of root-Nyquist filters.
- Equalizers are needed after decimators at the receiver.
- A choice of $K > N$ allows addition of guard bands between subcarrier bands.

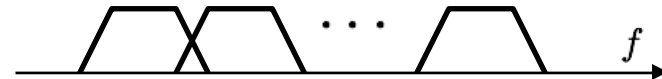


[CherEle99] G. Cherubini, E. Eleftheriou, S. Olcer (1999), "Filtered multitone modulation for VDSL," in *Proc. IEEE Globecom 99*, vol. 2, pp. 1139-1144, 1999.

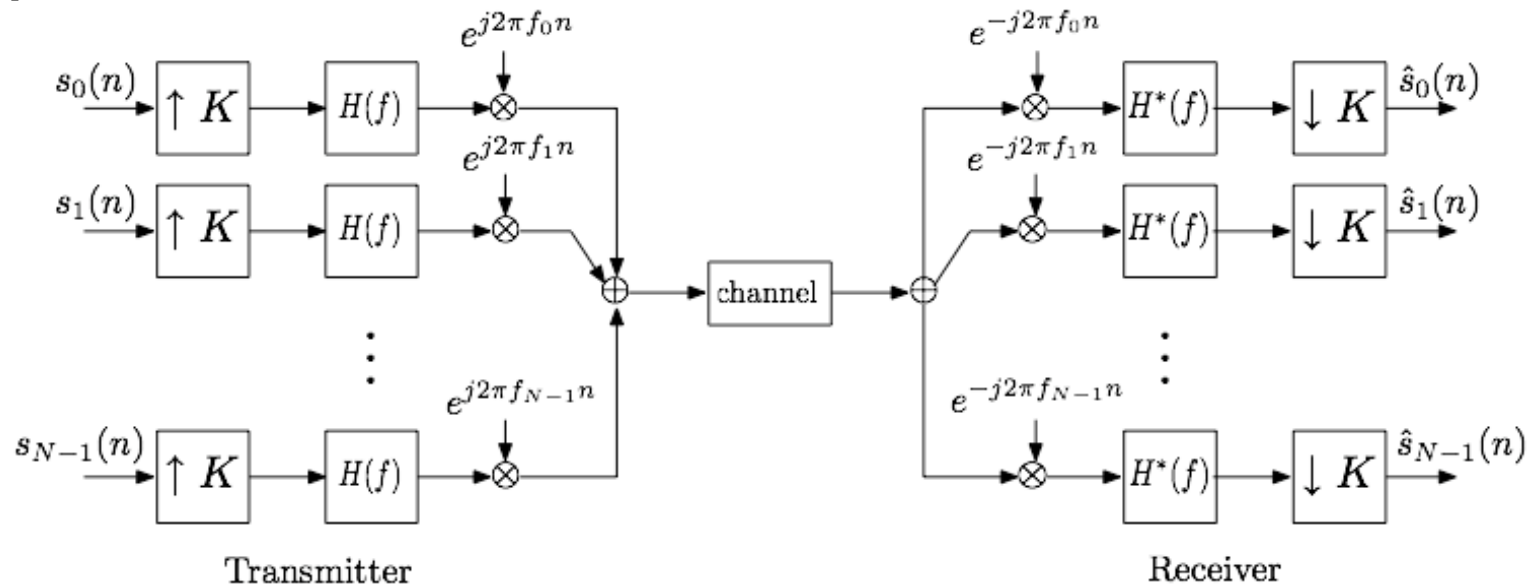


Filter Bank Multicarrier (FBMC) Methods: SMT (Summary)

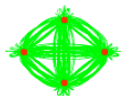
- SMT allows overlap of adjacent bands.
 - Maximizes bandwidth efficiency ($K = N$)



- Transmit symbols are offset QAM: in-phase and quadrature components have a time offset of half symbol interval (not shown below), i.e., time staggered.
- If the overlaps are limited to adjacent bands and $H(f)$ and $H^*(f)$ are a pair of root-Nyquist filters, separation of data symbols at the receiver output is guaranteed.
- Equalizers are needed after decimators at the receiver.

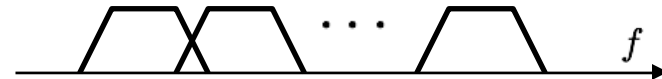


[Sal67] B.R. Saltzberg (1967), "Performance of an efficient parallel data transmission system," *IEEE Trans. Comm. Tech.*, vol. 15, no. 6, pp. 805-811, Dec. 1967.

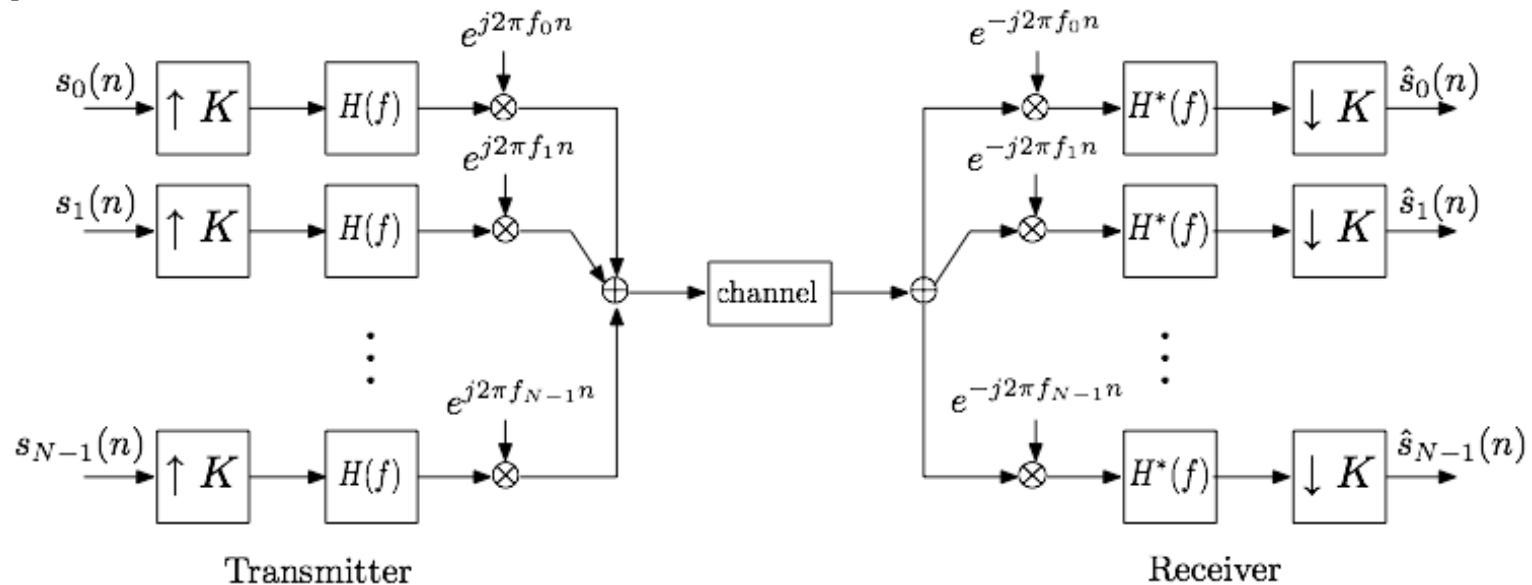


Filter Bank Multicarrier (FBMC) Methods: CMT (Summary)

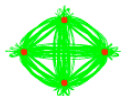
- CMT allows overlap of adjacent bands.
 - Maximizes bandwidth efficiency ($K = N$)



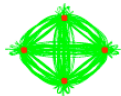
- Transmit symbols are PAM (pulse amplitude modulated). To allow maximum bandwidth efficiency, vestigial sideband modulation is adopted.
- The overlaps are limited to adjacent bands to simplify filter designs. Here, also, selection of root-Nyquist filters for $H(f)$ and $H^*(f)$ guarantees separation of data symbols at the receiver.
- Equalizers are needed after decimators at the receiver.



[Far03] B. Farhang-Boroujeny (2003), "Multicarrier modulation with blind detection capability using cosine modulated filter banks," *IEEE Trans. Commun.*, vol. 51, pp. 2057-2070, Dec. 2003.



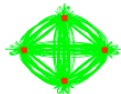
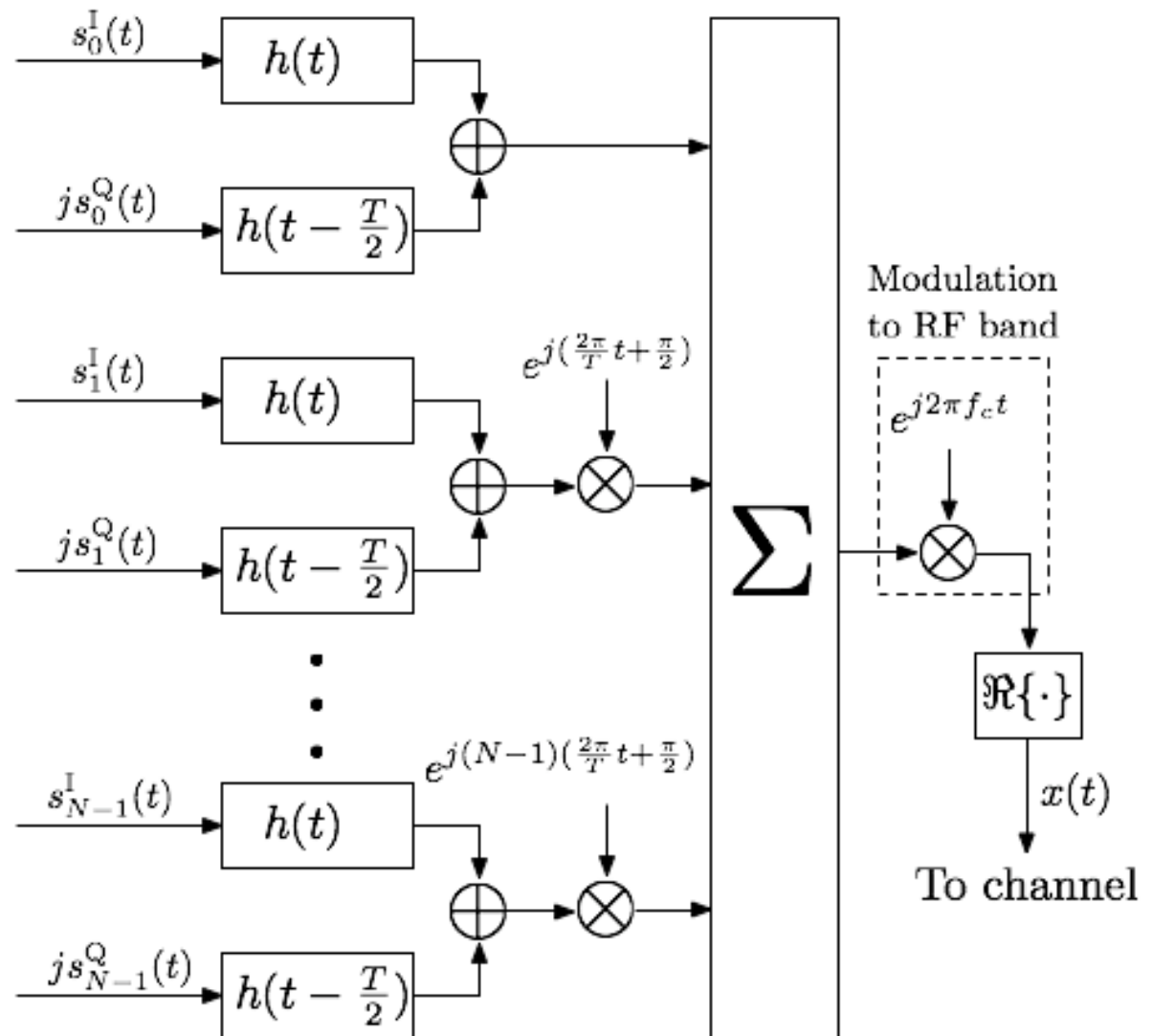
Staggered Modulated Multitone (SMT) Details



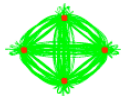
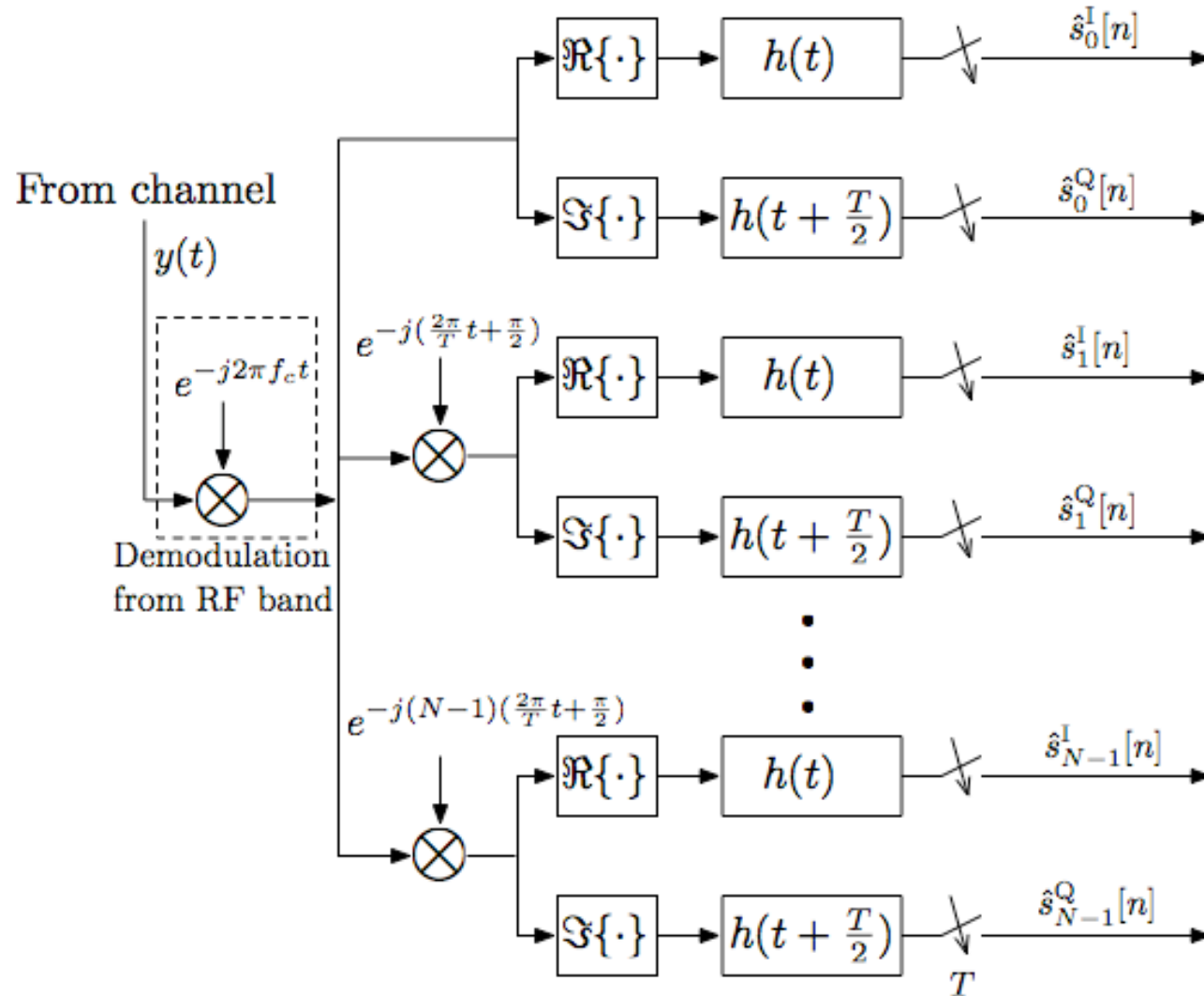
Filter Bank Multicarrier (FBMC) Methods: SMT (Synthesis/TX)

$$s_k(t) = \sum_{n=-\infty}^{\infty} s_k[n] \delta(t - nT)$$

$$s_k[n] = s_k^I[n] + js_k^Q[n]$$

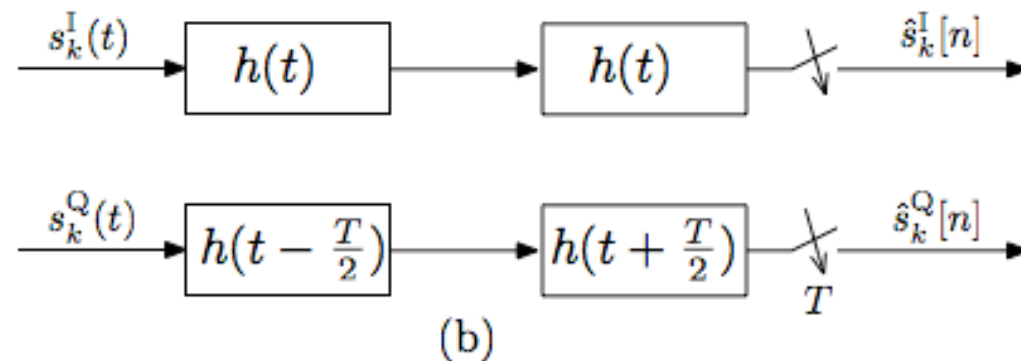
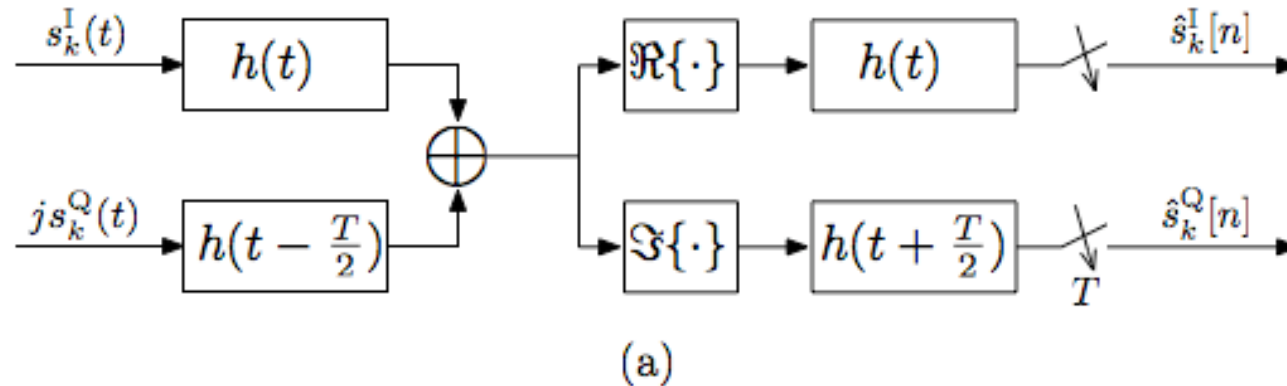


Filter Bank Multicarrier (FBMC) Methods: SMT (Analysis/RX)

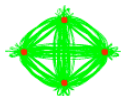


Filter Bank Multicarrier (FBMC) Methods: SMT (Details)

Intersymbol Interference (ISI):



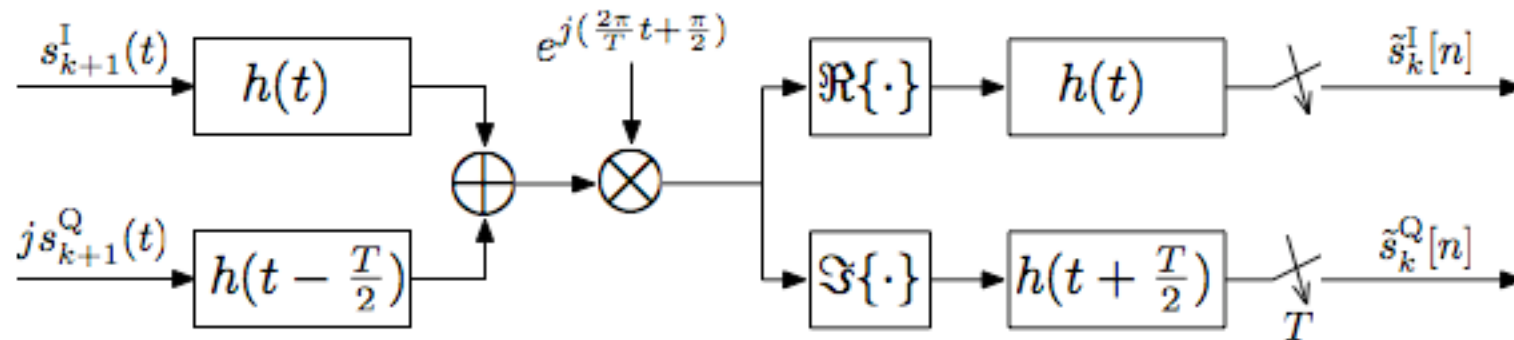
- (b) follows from (a) because $h(t)$ is a real-valued function.
- To achieve ISI free transmission $h(t)$ must be a square-root Nyquist and symmetric pulse shape.



Filter Bank Multicarrier (FBMC) Methods: SMT (Details)

Inter-carrier Interference (ICI):

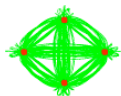
The path between the $k+1$ th and k th subcarrier:



Here, the outputs are ICI terms from $k+1$ th to the k th subcarrier. Thus, for ICI free transmission the output samples must be zero.

Equations that show this are presented in the next slide.

One of the four different cases is presented. The rest are similar.



Filter Bank Multicarrier (FBMC) Methods: SMT (Details)

The impulse response between the input $s_{k+1}^l(t)$ and the output before the sampler in the upper-right branch of the figure in the previous slide is given by

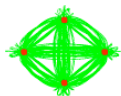
$$\begin{aligned} g_1(t) &= \Re \left\{ h(t) e^{j\left(\frac{2\pi}{T}t + \frac{\pi}{2}\right)} \right\} \star h(t) = \left(h(t) \cos \left(\frac{2\pi}{T}t + \frac{\pi}{2} \right) \right) \star h(t) \\ &= - \left(h(t) \sin \left(\frac{2\pi}{T}t \right) \right) \star h(t) = - \int_{-\infty}^{\infty} h(\tau) \sin \left(\frac{2\pi}{T}\tau \right) h(t - \tau) d\tau. \end{aligned}$$

Substituting $t=nT$, we get

$$g_1(nT) = - \int_{-\infty}^{\infty} h(\tau) \sin \left(\frac{2\pi}{T}\tau \right) h(nT - \tau) d\tau.$$

Applying the change of variable τ to $\frac{nT}{2} + \tau$

$$\begin{aligned} g_1(nT) &= - \int_{-\infty}^{\infty} h \left(\frac{nT}{2} + \tau \right) \sin \left(\frac{2\pi}{T}\tau + n\pi \right) h \left(\frac{nT}{2} - \tau \right) d\tau \\ &= (-1)^{n+1} \underbrace{\int_{-\infty}^{\infty} h \left(\frac{nT}{2} + \tau \right) h \left(\frac{nT}{2} - \tau \right) d\tau}_{\text{even}} \underbrace{\sin \left(\frac{2\pi}{T}\tau \right)}_{\text{odd}} = 0 \end{aligned}$$

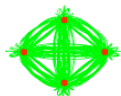


Filter Bank Multicarrier (FBMC) Methods: SMT (Details)

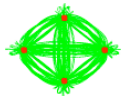
Channel Impact:

- In general, channel introduces a gain that varies across the channel.
- However, if the number of subcarriers is large, the gain over each subcarrier band may be approximated by a complex-valued constant, say h_k .
- When this approximation holds, the channel effect can be compensated for each subcarrier by using a single tap equalizer whose gain is set equal to $1/h_k$.
- Hirosaki (1981), has explored the problem for the more general case where channel gain varies across each subcarrier band and suggested a fractionally-spaced transversal equalizer for each subcarrier.

[Hir81] B. Hirosaki, “**An orthogonally multiplexed QAM system using the discrete Fourier transform**”, *IEEE Trans. Commun.*, Volume 29, Issue 7, Jul 1981, pp. 982 – 989.



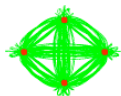
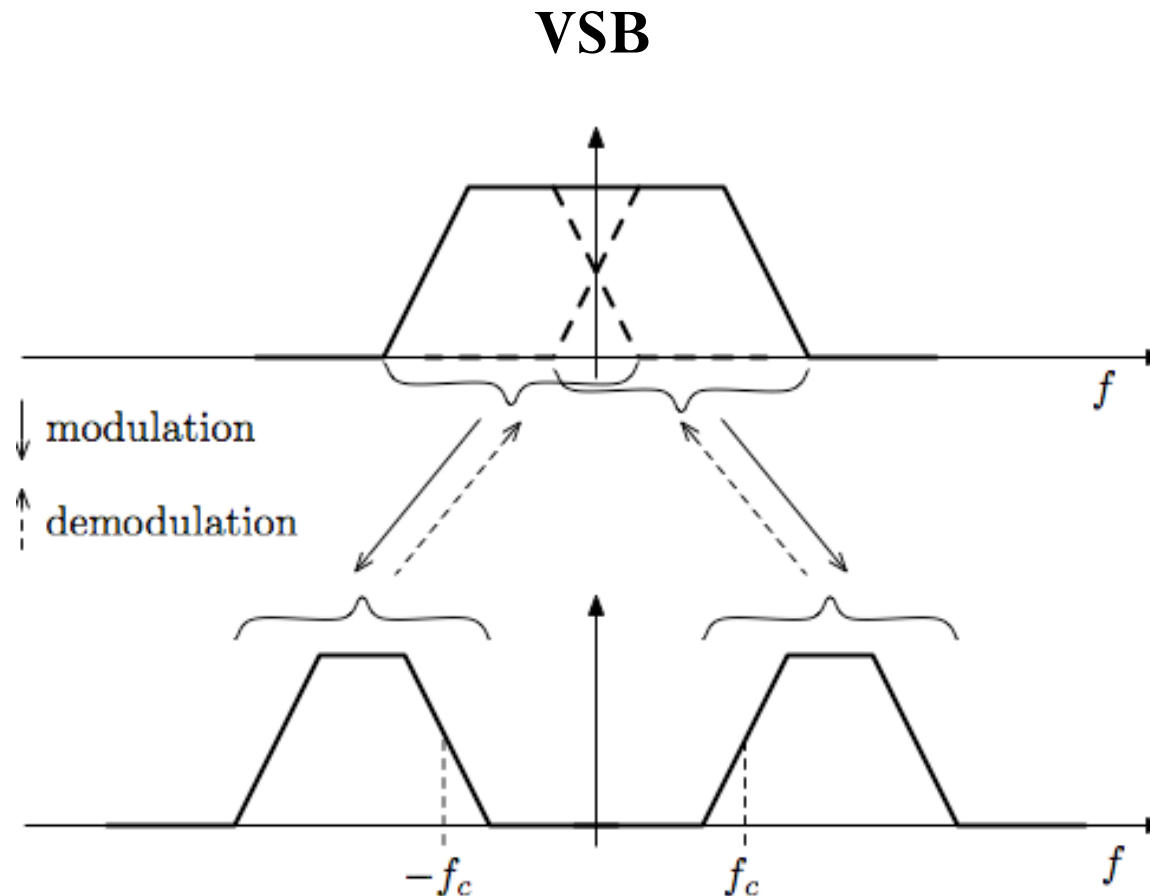
Cosine Modulated Multitone (CMT) Details



Filter Bank Multicarrier (FBMC) Methods: CMT (Details)

Modulation type: Vestigial Side-Band (VSB)

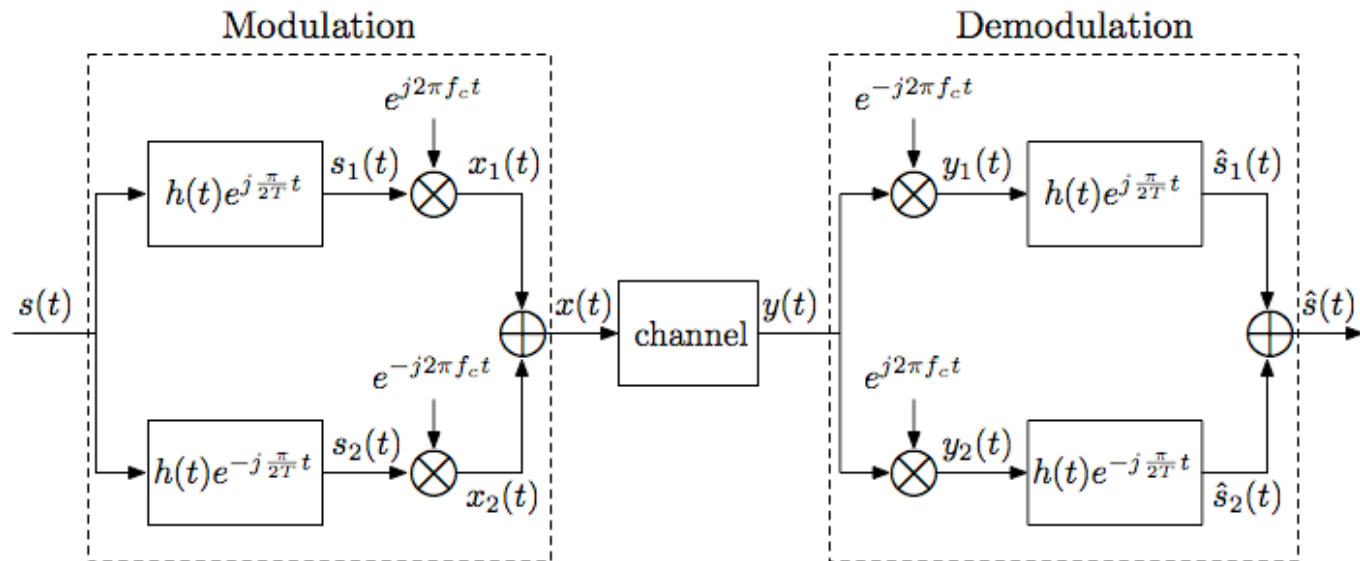
Data symbols: PAM (Pulse Amplitude Modulation)



Filter Bank Multicarrier (FBMC) Methods: CMT (Details)

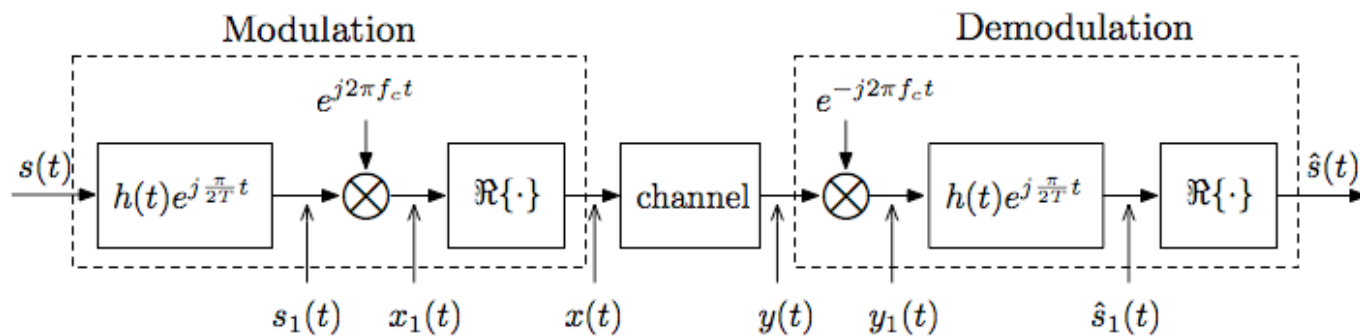
Block diagram of a VSB transceiver.

Detailed:

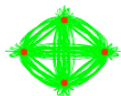


(a)

Simplified:



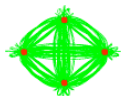
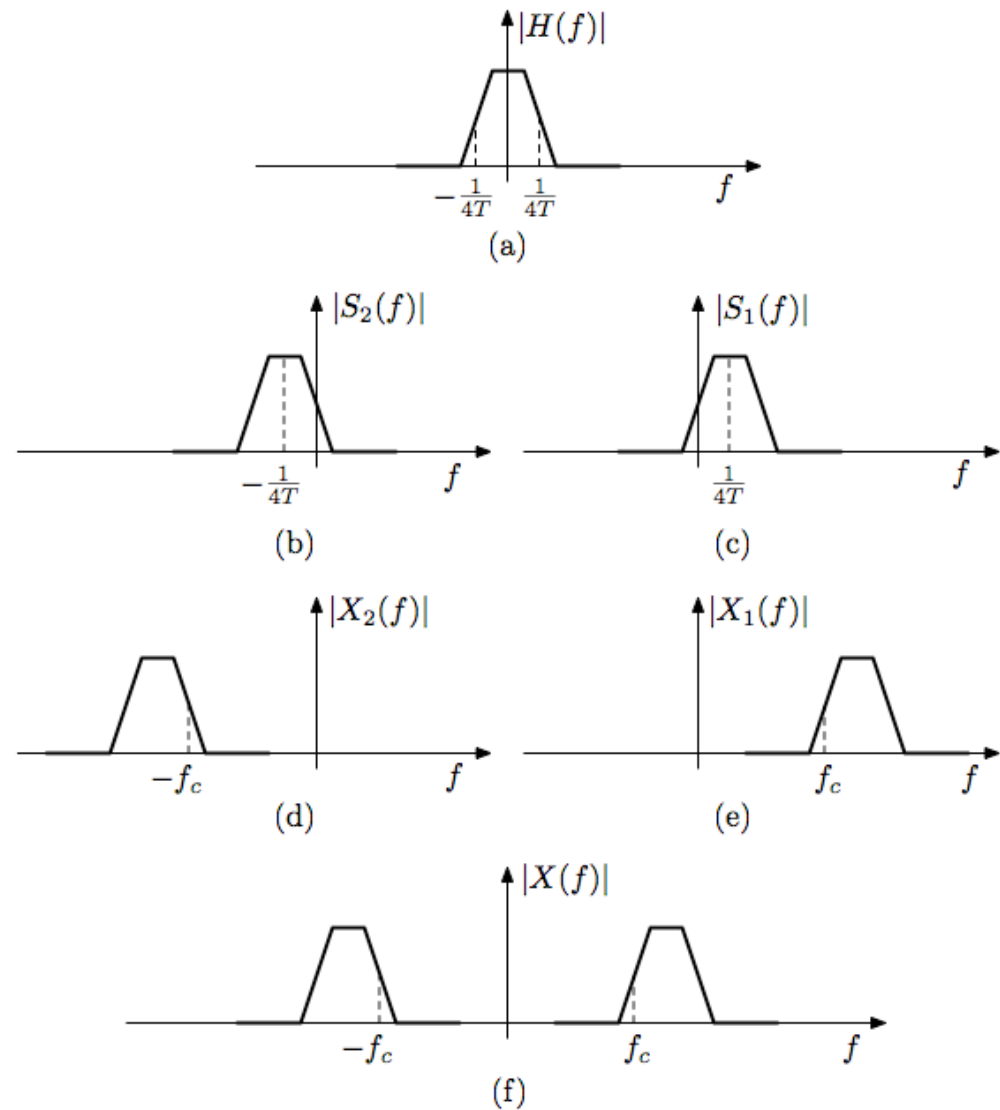
(b)



Filter Bank Multicarrier (FBMC) Methods: CMT (Details)

Spectra of various signals in
Figure (a) of the previous slide.

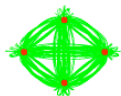
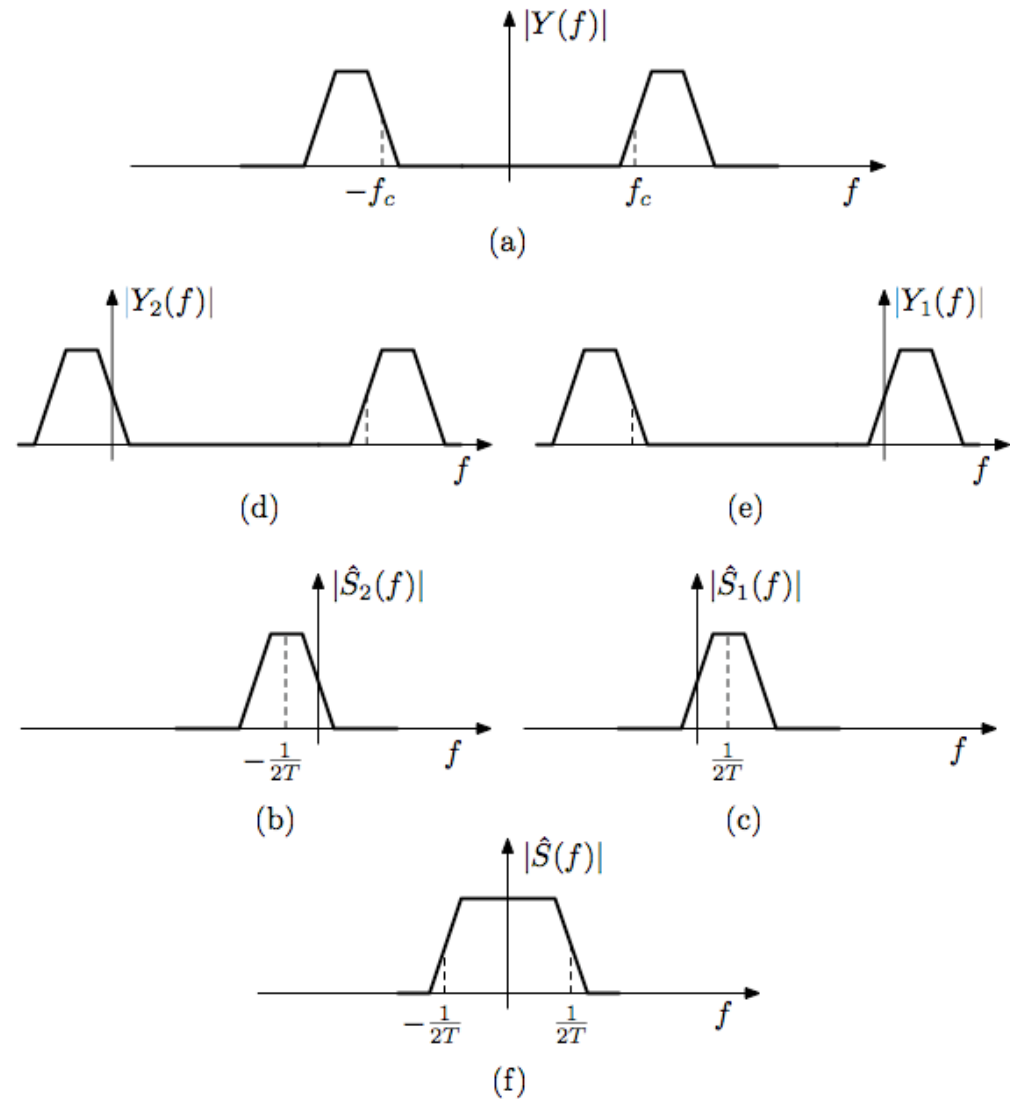
Set 1:



Filter Bank Multicarrier (FBMC) Methods: CMT (Details)

Spectra of various signals in
Figure (a) of the previous slide.

Set 2:



Filter Bank Multicarrier (FBMC) Methods: CMT (Details)

Intersymbol interference (ISI): Ignoring the channel distortion, the impulse response across a VSB channel is obtained as

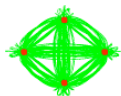
$$\begin{aligned} g(t) &= \Re\{h(t)e^{j\frac{\pi}{2T}t} \star h(t)e^{j\frac{\pi}{2T}t}\} \\ &= \Re\left\{\int_{-\infty}^{\infty} h(\tau)e^{j\frac{\pi}{2T}\tau} h(t-\tau)e^{j\frac{\pi}{2T}(t-\tau)} d\tau\right\} \\ &= \Re\left\{e^{j\frac{\pi}{2T}t} \int_{-\infty}^{\infty} h(\tau)h(t-\tau)d\tau\right\}. \end{aligned}$$

Letting $\int_{-\infty}^{\infty} h(\tau)h(t-\tau)d\tau = p(t)$, this simplifies to

$$g(t) = p(t) \cos\left(\frac{\pi}{2T}t\right).$$

Thus,

$$g(nT) = p(nT) \cos\left(\frac{\pi}{2T}nT\right) = p(nT) \cos\left(\frac{n\pi}{2}\right).$$



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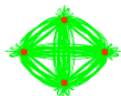
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Equal to zero for n non-zero even integer

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Filter Bank Multicarrier (FBMC) Methods: CMT (Details)

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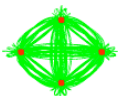
Thus,

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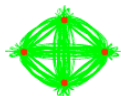
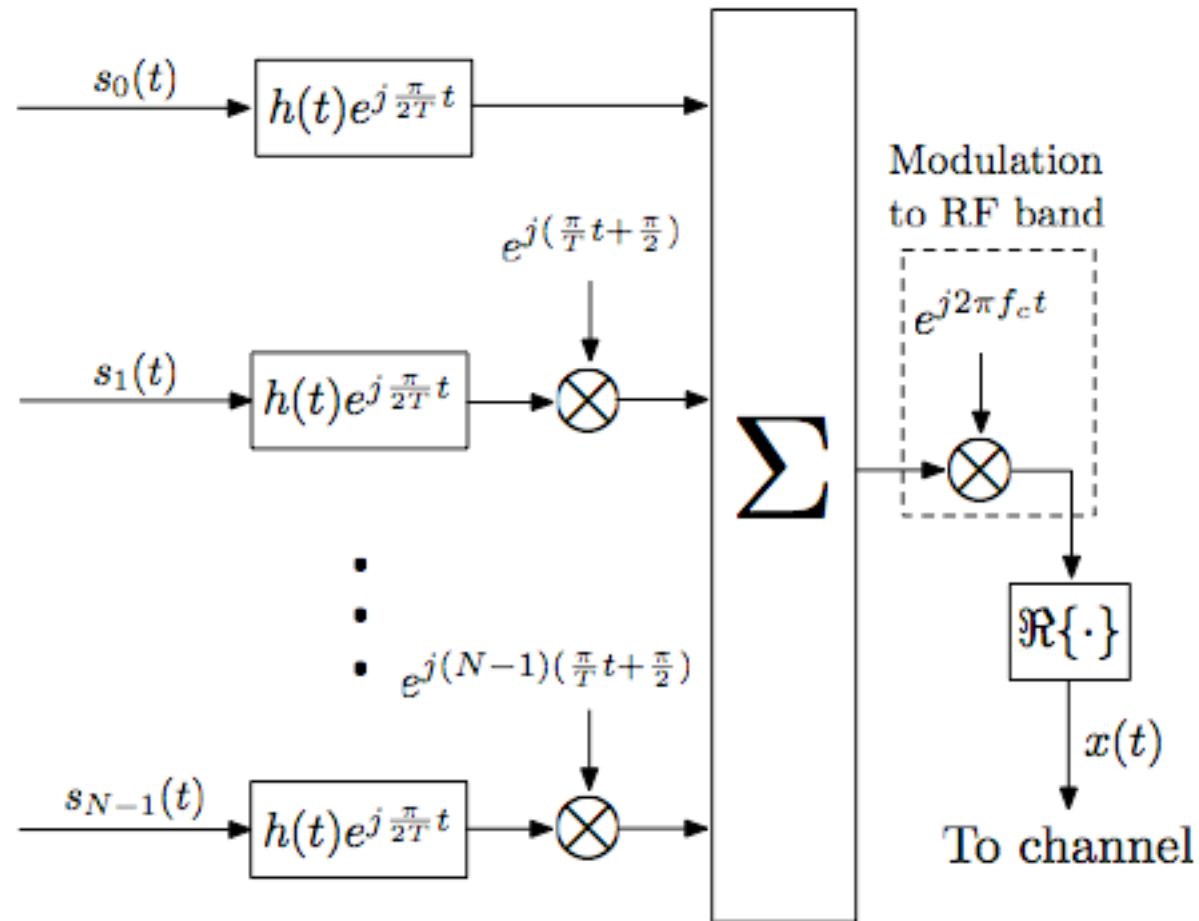
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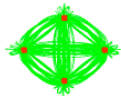
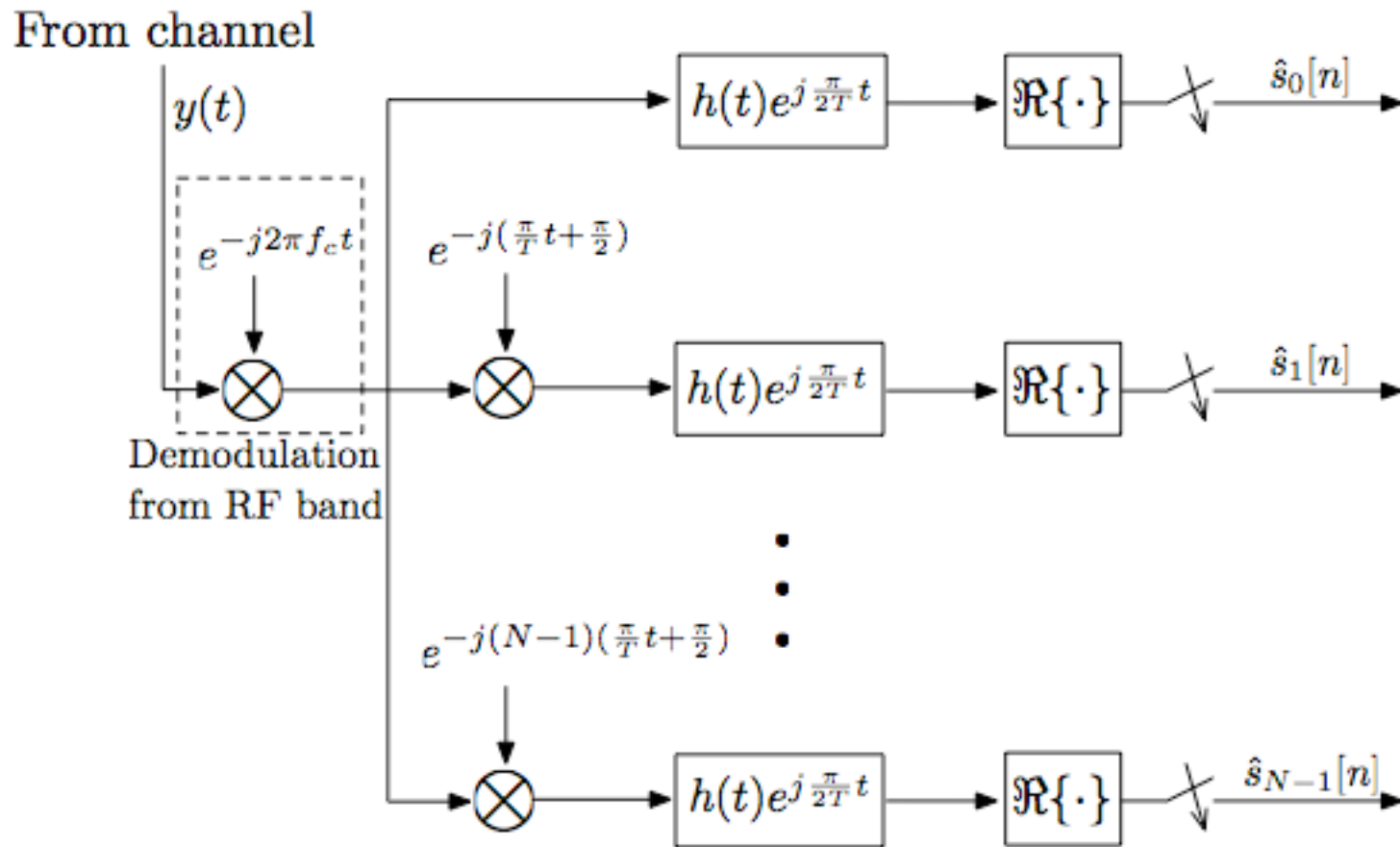
(i.e., $p(t)$ is Nyquist pulse with zero crossing at the interval $2T$)



Filter Bank Multicarrier (FBMC) Methods: CMT (Synthesis/TX)



Filter Bank Multicarrier (FBMC) Methods: CMT (Receiver/RX)



Filter Bank Multicarrier (FBMC) Methods: CMT (Details)

The impulse response between the input $s_{k+1}^l(t)$ at the transmitter and the output before the sampler at the k th subcarrier of the receiver is

$$\begin{aligned} g_1(t) &= \Re\{h(t)e^{j(\frac{3\pi}{2T}t + \frac{\pi}{2})} \star h(t)e^{j\frac{\pi}{2T}t}\} = \Re\left\{\int_{-\infty}^{\infty} h(\tau)e^{j(\frac{3\pi}{2T}\tau + \frac{\pi}{2})} h(t-\tau)e^{j\frac{\pi}{2T}(t-\tau)} d\tau\right\} \\ &= \Re\left\{e^{j(\frac{\pi}{2T}t + \frac{\pi}{2})} \int_{-\infty}^{\infty} h(\tau)e^{j\frac{\pi}{T}\tau} h(t-\tau) d\tau\right\}. \end{aligned}$$

Noting that

$$e^{j(\frac{\pi}{2T}t + \frac{\pi}{2})} = -\sin\left(\frac{\pi}{2T}t\right) + j\cos\left(\frac{\pi}{2T}t\right)$$

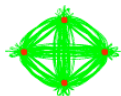
and, since $h(t)$ is a real function,

$$\int_{-\infty}^{\infty} h(\tau)e^{j\frac{\pi}{T}\tau} h(t-\tau) d\tau = \int_{-\infty}^{\infty} h(\tau)h(t-\tau) \cos\left(\frac{\pi}{T}\tau\right) d\tau + j \int_{-\infty}^{\infty} h(\tau)h(t-\tau) \sin\left(\frac{\pi}{T}\tau\right) d\tau.$$

Hence,

$$g_1(t) = -\sin\left(\frac{\pi}{2T}t\right) \int_{-\infty}^{\infty} h(\tau)h(t-\tau) \cos\left(\frac{\pi}{T}\tau\right) d\tau - \cos\left(\frac{\pi}{2T}t\right) \int_{-\infty}^{\infty} h(\tau)h(t-\tau) \sin\left(\frac{\pi}{T}\tau\right) d\tau.$$

We are interested in the sample values of $g_1(t)$ at the time instants nT , for integer values of n .



Filter Bank Multicarrier (FBMC) Methods: CMT (Details)

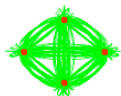
For an even value of $n = 2k$, one finds that $\sin\left(\frac{\pi}{2T}2kT\right) = \sin(k\pi) = 0$ and, thus

$$g_1(2kT) = (-1)^{k+1} \int_{-\infty}^{\infty} h(\tau)h(2kT - \tau) \sin\left(\frac{\pi}{T}\tau\right) d\tau$$

where we have noted that $\cos(k\pi) = (-1)^k$. Applying a change of variable τ to $kT + \tau$, we get

$$g_1(2kT) = - \int_{-\infty}^{\infty} h(kT + \tau)h(kT - \tau) \sin\left(\frac{\pi}{T}\tau\right) d\tau = 0.$$

where the second identity follows since the expression under the integral is an odd function of τ .



Filter Bank Multicarrier (FBMC) Methods: CMT (Details)

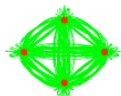
For an odd values of $n = 2k + 1$, one finds that $\cos\left(\frac{\pi}{2T}(2k + 1)T\right) = \cos(k\pi + \pi/2) = 0$ and, thus,

$$g_1((2k + 1)T) = (-1)^{k+1} \int_{-\infty}^{\infty} h(\tau)h((2k + 1)T - \tau) \cos\left(\frac{\pi}{T}\tau\right) d\tau.$$

where we have noted that $\sin(k\pi + \pi/2) = (-1)^k$. Applying a change of variable τ to $\frac{(2k+1)T}{2} + \tau$, we get

$$g_1\left(\frac{(2k+1)T}{2}\right) = \int_{-\infty}^{\infty} h\left(\frac{(2k+1)T}{2} + \tau\right)h\left(\frac{(2k+1)T}{2} - \tau\right) \sin\left(\frac{\pi}{T}\tau\right) d\tau = 0.$$

where the second identity follows since the expression under the integral is an odd function of τ .



Filter Bank Multicarrier (FBMC) Methods: CMT (Details)

For an odd values of $n = 2k + 1$, one finds that $\cos\left(\frac{\pi}{2T}(2k + 1)T\right) = \cos(k\pi + \pi/2) = 0$ and, thus,

$$g_1((2k + 1)T) = (-1)^{k+1} \int_{-\infty}^{\infty} h(\tau)h((2k + 1)T - \tau) \cos\left(\frac{\pi}{T}\tau\right) d\tau.$$

where we have noted that $\sin(k\pi + \pi/2) = (-1)^k$. Applying a change of variable τ to $\frac{(2k+1)T}{2} + \tau$, we get

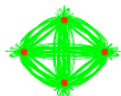
$$g_1\left(\frac{(2k+1)T}{2}\right) = \int_{-\infty}^{\infty} h\left(\frac{(2k+1)T}{2} + \tau\right)h\left(\frac{(2k+1)T}{2} - \tau\right) \sin\left(\frac{\pi}{T}\tau\right) d\tau = 0.$$

where the second identity follows since the expression under the integral is an odd function of τ .

Combining the above results, one finds that

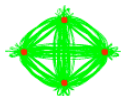
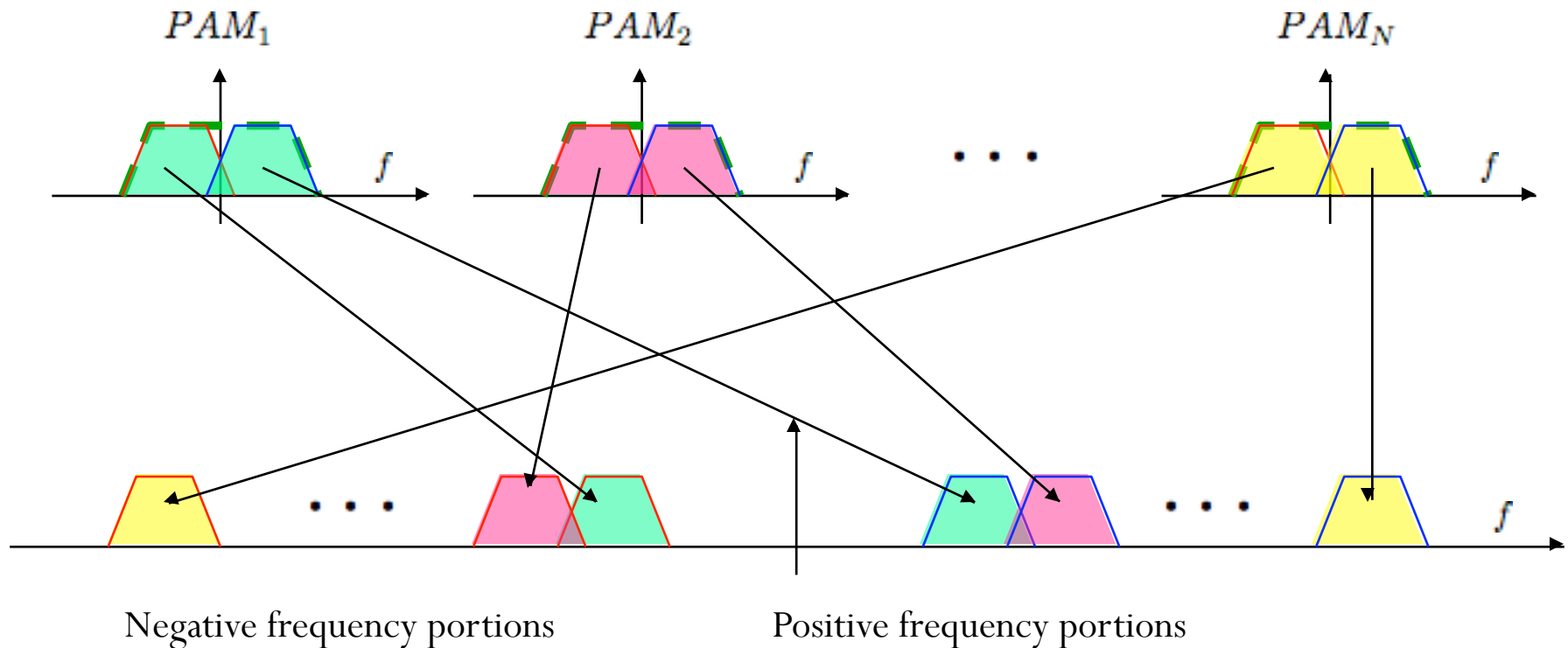
$$g_1(nT) = 0$$

for all integer values of n , which is the condition we required for ICI free transmission.



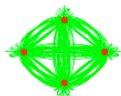
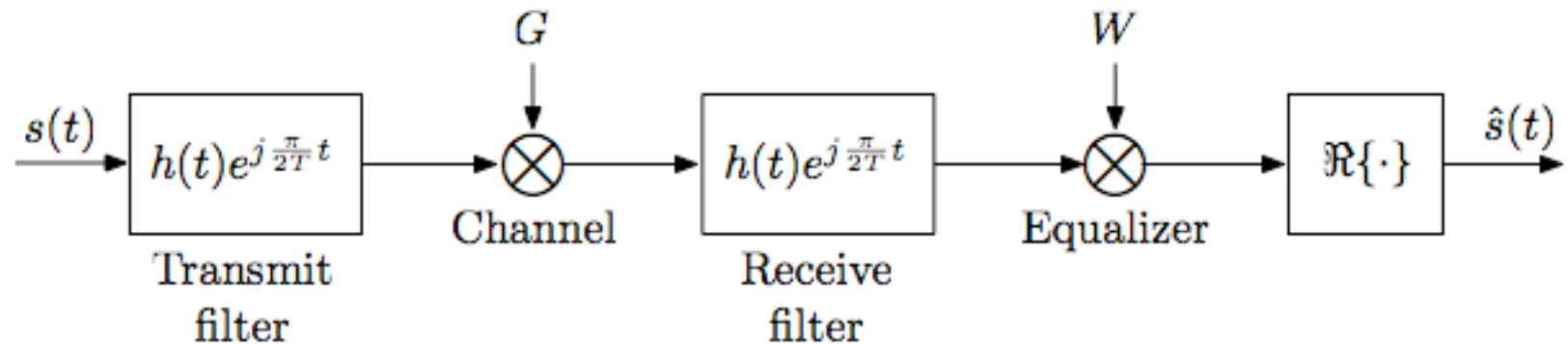
Filter Bank Multicarrier (FBMC) Methods: CMT (Details)

In CMT data symbols are PAM and modulated into a set of vestigial sideband subcarrier channels.



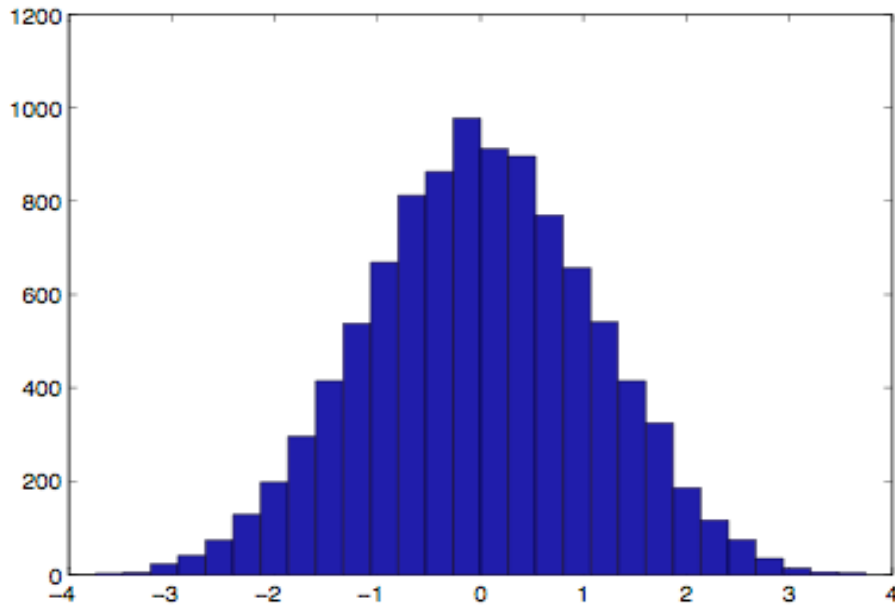
Filter Bank Multicarrier (FBMC) Methods: CMT (Details)

Equalization:

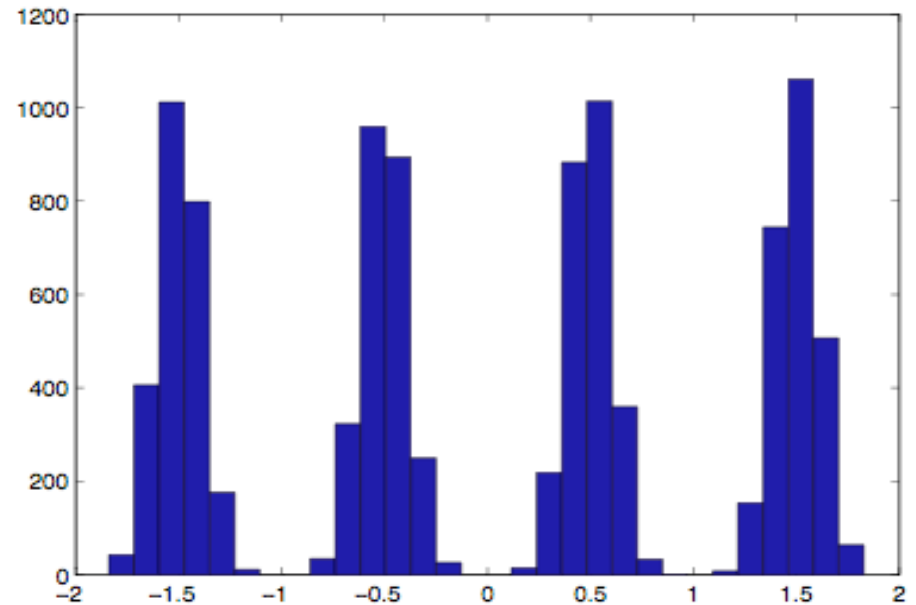


Filter Bank Multicarrier (FBMC) Methods: CMT (Details)

Blind equalization

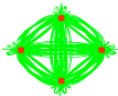


Before equalization



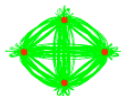
After equalization

[Far03] B. Farhang-Boroujeny (2003), “**Multicarrier modulation with blind detection capability using cosine modulated filter banks**,” *IEEE Trans. Commun.*, vol. 51, pp. 2057-2070, Dec. 2003.

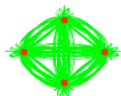


Recent results show that:

- SMT and CMT are equally sensitive to carrier and timing offset.
- However, in terms of tracking, SMT receives less gradient noise.

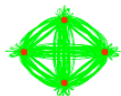
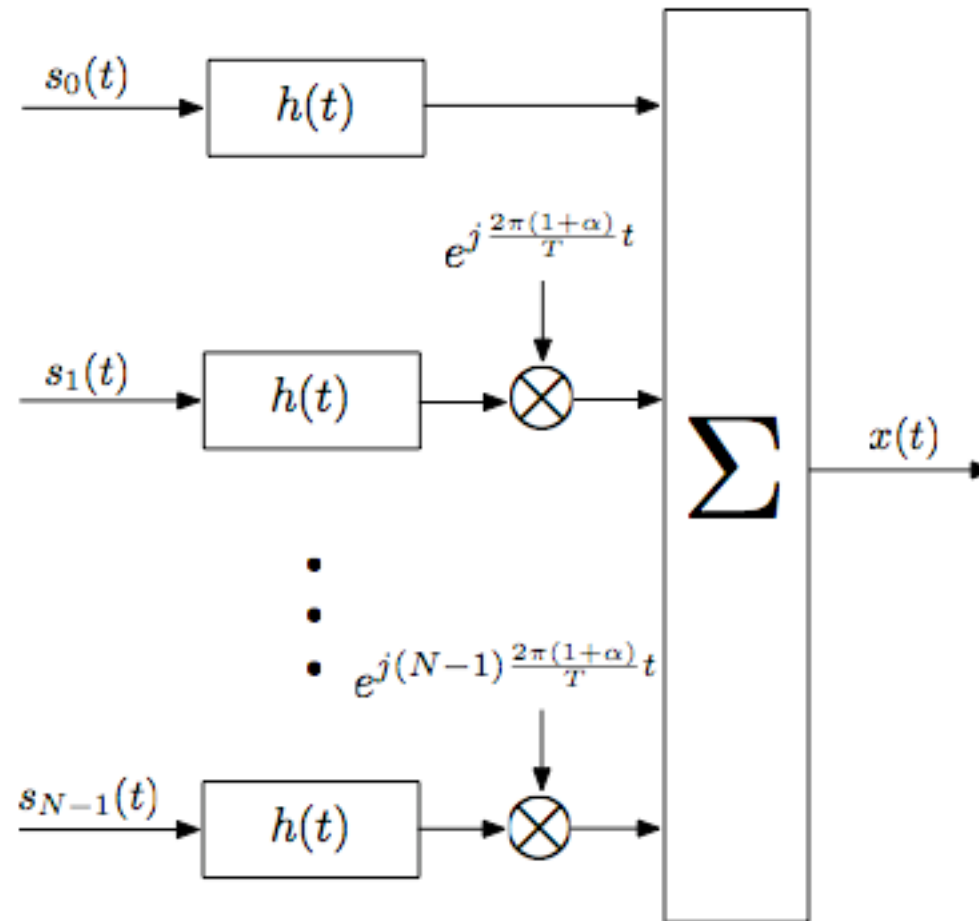


Filtered Multitone (Details)



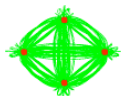
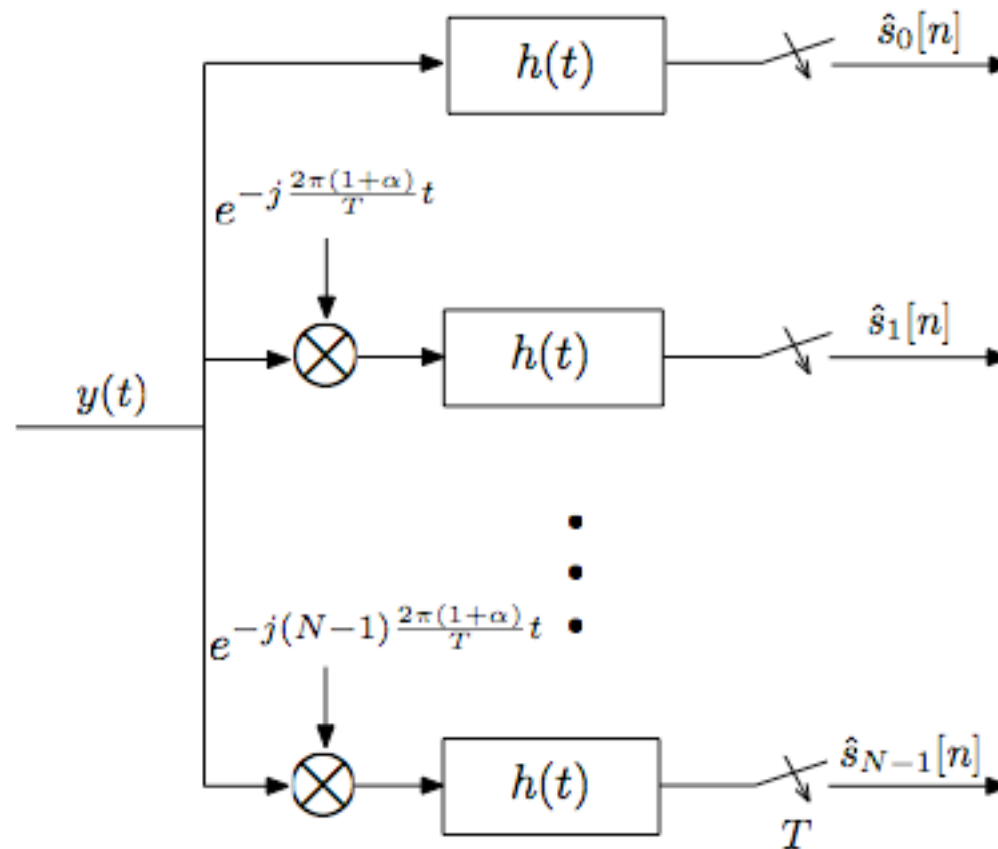
Filter Bank Multicarrier (FBMC) Methods: FMT (Synthesis/TX)

Transmitter:



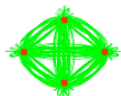
Filter Bank Multicarrier (FBMC) Methods: FMT (Analysis/RX)

Receiver:



Conclusions

- We showed that OFDM is not a good choice for cognitive radios.
 - This may also be extended to multiuser networks
- FBMC, on the other hand, is a perfect match for cognitive radios
 - Filter banks can be used for both spectrum sensing and data transmission.



Problems to solve

- Compared to OFDM, FBMC studies are very limited.
- A few suggestions:
 - More papers have to be written to convince the SP community that FBMC is a viable competitor to OFDM and it should be more seriously considered
 - Synchronization issues of FBMC should be further addressed
 - The case of multiuser network needs a lot of attention
 - MAC and upper layers should be addressed
 - MIMO FBMC systems should also be developed and evaluated
 - Testbeds and real world evaluation of FBMC system is a good way of convincing people

