MIMO processing for highly efficient FBMC waveforms

Ana I. Pérez Neira

Universitat Politècnica de Catalunya (UPC)

Centre Tecnològic de Telecomunicacions de Catalunya (CTTC)
Traffic Forecasts

Compound annual growth rate by region

Exabytes per Month

- 61% CAGR 2013–2018

- Middle East and Africa (9.4%)
- North America (18.6%)
- Asia Pacific (42.4%)
- Central and Eastern Europe (10.3%)
- Western Europe (12.0%)
- Latin America (7.3%)

Figures in parentheses refer to regional share in 2018.
Source: Cisco VNI Mobile, 2014
Traffic Forecasts

Spectrum and traffic growth

Spectral efficiency and freq. reuse growth

Motivation: Where will an increase in wireless capacity come from?

- Strategies to keep pace with the demands of users:
  - Cell densification together with aggressive frequency reuse schemes
  - Utilize larger portions of radio spectrum, e.g. the millimeter wavelength regions and authorized spectrum sharing
  - New technologies

- Requirements imposed on the air-interface:
  - Allow a fine-grained control of the spectrum
  - Exhibit a reduced out-of-band radiation
  - Achieve robustness to synchronization errors
  - MIMO
Motivation
Motivation
Outline

1. Air-interface candidates
2. FBMC/OQAM overview
3. MIMO architectures with CSIT
   - MIMO processing in low frequency selective subchannels
   - MIMO processing in frequency selective subchannels
4. MIMO processing with no CSIT
5. MIMO processing in multi-user systems
6. Open topics
Air-interface candidates

OFDM vs. FBMC

- OFDM exhibits a sinc-shaped spectrum
- FBMC shapes subcarriers with localized pulses in the frequency domain
- FBMC transmission does not necessarily need a cyclic prefix
- FBMC modulation/demodulation can be efficiently realized with FFT and polyphase filtering
Experimental results: Spectral efficiency for downlink PUSC, Veh A 60 km/h, channel estimation

- **(5GNOW)** [http://www.5gnow.eu/](http://www.5gnow.eu/)
New waveform designs. Key ingredients:

- Cyclic prefix
- Circular or linear convolution
- Time windowing
- Orthogonality
- Pulse shaping

Efficiency, low-complexity and good localization in time and frequency domains
Air-interface candidates

Windowed OFDM (W-OFDM)

$n$th Interval

CP (n-1)th Symbol CS

+ CP n$\text{th}$ Symbol CS

PSD (dB)

Amplitude

(WSA 2015)
Filtered Multitone - FMT

\[(n-1)\text{th Symbol}\]

\[n\text{th Symbol}\]

\[(n+1)\text{th Symbol}\]

\[n\text{th Interval}\]

Generalized frequency division multiplexing - GFDM

Filter bank multicarrier/OQAM - FBMC/OQAM

Circular convolution in FBMC/OQAM

**Comparison**

- **Desired features:** Maximum bandwidth efficiency, simple equalization and good localization in the time domain

- These three metrics may tradeoff with each other, and thus there may not be a unified solution to solve all the issues and satisfy all the requirements

<table>
<thead>
<tr>
<th></th>
<th>CP</th>
<th>Orthogonal</th>
<th>Equalization</th>
<th>Block structure</th>
<th>Continuity on edges</th>
<th>$\eta = \frac{1}{T\Delta_f}$</th>
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<tr>
<td>W-OFDM</td>
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<td>Yes</td>
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<td>Yes</td>
<td>Yes</td>
<td>&lt;1</td>
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<tr>
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<td>FBMC/OQAM</td>
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<td>Yes</td>
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<tr>
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<td>Complex</td>
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- **FBMC/OQAM** is the most spectrally efficient contender
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Multicarrier communication systems based on QAM
Offset Quadrature Phase-Shift Keying (OQPSK)

In-phase

Quadrature
Multicarrier communication systems based on OQAM
The imaginary part is delayed half the symbol period on even subcarriers.
The real part is delayed half the symbol period on odd subcarriers.

QAM symbols $\rightarrow x_R,m[l] + jx_I,m[l]$
PAM symbols $\rightarrow d_m[k]$
Phase term $\rightarrow \theta_m[k] = \begin{cases} 1 & k + m \text{ even} \\ j & k + m \text{ odd} \end{cases}$
FBMC/OQAM overview

Transmitter

Transmitted signal

\[ s[n] = \sum_{k \in \mathbb{Z}} \sum_{m=0}^{M-1} d[m][k] \theta[m][k] f_m \left[ n - k \frac{M}{2} \right] \]

Subband pulses

\[ f_m[n] = p[n] e^{j \frac{2\pi}{M} m \left( n - \frac{L-1}{2} \right)} \]

PAM symbols

\[ d[m][k] \]

Phase term

\[ \theta[m][k] = \begin{cases} 1 & k + m \text{ even} \\ j & k + m \text{ odd} \end{cases} \]

Transceiver

FBMC/OQAM overview

Weak orthogonality
Synthesis filter bank: efficient implementation

Subband pulses
\[ f_m[n] = p[n] e^{j\frac{2\pi}{M} m(n - \frac{L-1}{2})} = p[n] e^{j\frac{2\pi}{M} mn} \beta_m \]

Prototype pulse
\[ p[n], \quad n \in \{0, \ldots, L-1\}, \quad L = KM \]

Polyphase decomposition
\[ P(z) = \sum_{n \in \mathbb{Z}} p[n] z^{-n} = \sum_{m=0}^{M-1} z^{-m} A_m \left( z^M \right) \]

- Complexity FBMC/OQAM: \(2M \log_2(M) + 4KM\)
- Complexity OFDM: \(M \log_2(M)\)
Analysis filter bank: efficient implementation

\[ r[n] \quad \rightarrow \quad y_q[k] = \left\langle r[n], f_q^* \left[ n - \frac{kM}{2} \right] \right\rangle \]

\[ y_q[k] = \sum_{m=q-1}^{q+1} g_{qm}[k] \ast (d_m[k] \theta_m[k]) + w_q[k] \]

Equivalent Channel \quad \rightarrow \quad g_{qm}[k] = (f_m[n] \ast h[n] \ast f_q^*[-n]) \downarrow \frac{M}{2}

Filtered Noise \quad \rightarrow \quad w_q[k] = (w[n] \ast f_q^*[-n]) \downarrow \frac{M}{2}
System model

\[ y_q[k] = g_{qq}[0]d_q[k]\theta_q[k] + \sum_{\tau \neq 0} g_{qq}[\tau]d_q[k - \tau]\theta_q[k - \tau] \]

Desired

\[ + \sum_{m \neq q} \sum_{\tau} g_{qm}[\tau]d_m[k - \tau]\theta_m[k - \tau] + w_q[k] \]

self ISI

ISI+ICI

Symbols \( \{d_q[k]\} \) are non-circular. Second order moments:

- Covariance: \( \mathbb{E} \{y_q[k]y_q^*[k]\} = C_{RR} + C_{II} + j(C_{RI} - C_{IR}) \)
- Pseudo-covariance: \( \mathbb{E} \{y_q[k]y_q[k]\} = C_{RR} - C_{II} + j(C_{RI} + C_{IR}) \)

Real and imaginary parts have to be independently processed to retrieve all the information. Particular case of Widely Linear:

- \( \mathbb{E} \{([\Re(y_q[k]) \ \Im(y_q[k])]^T [\Re(y_q[k]) \ \Im(y_q[k])]) \} \)
FBMC/OQAM overview

Channel model

Model 1  \( g_{qm}[k] = (f_m[n] \ast h[n] \ast f_q^*[-n]) \downarrow \frac{M}{2} \)

Model 2  \( g_{qm}[k] = H_m (f_m[n] \ast f_q^*[-n]) \downarrow \frac{M}{2} \)

Intrinsic interference

\[ \alpha_{qm}[k] = (f_m[n] \ast f_q^*[-n]) \downarrow \frac{M}{2} \]

<table>
<thead>
<tr>
<th>( k = -3 )</th>
<th>( k = -2 )</th>
<th>( k = -1 )</th>
<th>( K = 0 )</th>
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<th>( k = 2 )</th>
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<td>-0.0429j</td>
<td>0.1250</td>
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Channel frequency response

ITU Vehicular A
ITU Pedestrian A
ITU Vehicular B
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MIMO architectures with CSIT

General architecture

Input/Output relation \[ r(\omega) = A^H(\omega)H(\omega)B(\omega)s(\omega) + A^Hw(\omega) \]

Design equations \[ A^H(\omega)H(\omega)B(\omega) = I_S \]
\[ B(\omega) \in \mathbb{C}^{NT \times S}, \ H(\omega) \in \mathbb{C}^{NR \times NT}, \ A(\omega) \in \mathbb{C}^{NR \times S} \]

- The complexity required to achieve a diagonal structure in the general architecture renders the solution impractical

Low-complexity architecture for flat fading subchannels

Response \( m \)th subcarrier \( \rightarrow F_m(\omega) = P \left( \omega - \frac{2\pi m}{M} \right) \)

SFB+precoder \( \rightarrow F_m(\omega)B(\omega) \approx F_m(\omega)B \left( \frac{2\pi m}{M} \right) = F_m(\omega)B_m \)

If the channel frequency response is flat at the subcarrier level, then \( B(\omega) \) does not present strong variations around \( \omega = \frac{2\pi m}{M} \).
Channel model

Model 1 → $g_{qim}^i[k] = (f_m[n] * h_{ji}[n] * f_q^*[-n]) \downarrow_{M/2}$, \quad $1 \leq j \leq N_R$, \quad $1 \leq i \leq N_T$

$G_{qm}[k] = \begin{bmatrix}
    g_{qm}^{11}[k] & \cdots & g_{qm}^{1N_T}[k] \\
    \vdots & \ddots & \vdots \\
    g_{qm}^{N_R1}[k] & \cdots & g_{qm}^{N_RN_T}[k]
\end{bmatrix} \in \mathbb{C}^{N_R \times N_T}$

Model 2 → $g_{qim}^i[k] = H_{ji}(m) (f_m[n] * f_q^*[-n]) \downarrow_{M/2}$, \quad $1 \leq j \leq N_R$, \quad $1 \leq i \leq N_T$

$G_{qm}[k] = \alpha_{qm}[k]H_m = \alpha_{qm}[k] \begin{bmatrix}
    H_{11}(m) & \cdots & H_{1N_T}(m) \\
    \vdots & \ddots & \vdots \\
    H_{N_R1}(m) & \cdots & H_{N_RN_T}(m)
\end{bmatrix} \in \mathbb{C}^{N_R \times N_T}$

$G_{qm}[k] \rightarrow$ Impulse response from subcarrier $m$ to subcarrier $q$
The optimal joint transceiver design is too complex. Proposed approach:

- Choose the least complex architecture depending on the selectivity of the channel frequency response

- Play with the available DoF: spatial and temporal diversity
Joint tx-rx beamforming for flat fading subchannels

- Input/output relation when $\theta_q[k] = 1$:

$$
\tilde{d}_q[k] = \Re (A_q^H H_q B_q) d_q[k] + \sum_{(m,\tau)\neq(q,0)} \Re (\alpha_{qm}[\tau] \theta_m[k-\tau] A_q^H H_m B_m) d_m[k-\tau] \\
+ \Re (A_q^H w_q[k])
$$

- Matrix pair $\{A_q, B_q\}$ in OFDM $\rightarrow A_q^H H_q B_q$ becomes diagonal

- MIMO techniques designed for OFDM yield ISI and ICI. Alternatives:
  
  ▶ Zero Forcing [1]:
  
  $\Re (A_q^T H_q B_q)$ becomes diagonal and $\Im (A_q^H H_q B_q) = A_q^T \Im (H_q B_q)$

  ▶ Coordinated beamforming [2]:
  
  $\Re (A_q^H H_q B_q)$ becomes diagonal and $0 = \Im (A_q^H H_q B_q) = A_q^T \Im (H_q B_q)$


Linear vs. Widely linear processing

- Linear processing (OFDM):
  - $A_q, B_q$ complex-valued.
  - The channel is diagonalized with these spatial channel gains:
    $$\{ \beta_1^q > \cdots > \beta_{NR}^q \}$$

- Widely linear processing (FBMC/OQAM ZF):
  - Non-circular nature of the symbols is taken into account
  - $A_q$ real-valued, $B_q$ complex-valued.
  - Dimensionality constraint imposed by the ZF condition: $2N_T > N_R$
  - The channel is diagonalized with these spatial channel gains:
    $$\{ \lambda_1^q > \cdots > \lambda_{NR}^q \}$$

- Gains are less spread out in FBMC/OQAM than in OFDM:
  - $\beta_1^q \geq \lambda_1^q$
  - $\beta_{NR}^q \leq \lambda_{NR}^q$

- FBMC/OQAM remains competitive with OFDM if $S = N_R \leq N_T$
Numerical results

- Carriers $M=1024$
- $N_T = 4$, $N_R = 2$, $S = 2$
- Subcarrier spacing $\Delta_f=15$ KHz
- Sampling frequency $f_s=15.36$ MHz
- Extended Vehicular A channel. Delay spread: $\sigma = 357$ ns
- Active carriers to comply with the spectral mask:
  - 600 (OFDM)
  - 650 (FBMC/OQAM)
Low-complexity architecture for frequency selective subchannels

- To combat the channel frequency selectivity:
  - Multi-tap equalization

System model:

\[
\tilde{d}_q[k] = \sum_{m=q-1}^{q+1} \Re \left( A_q^H[k] \ast G_{qm}[k] \ast (\theta_m[k] B_m d_m[k]) \right) + \Re \left( A_q^H[k] \ast w_q[k] \right)
\]
Compact formulation

\[
\tilde{d}_q[k] = \sum_{m=q-1}^{q+1} \bar{A}_q^T (\bar{E}_{qm}[k] \ast B_m d_m[k]) + \bar{A}_q^T \bar{w}_q[k]
\]

\[
= \bar{A}_q^T \bar{E}_{qq}[0] B_q d_q[k] + \sum_{(m,\tau) \neq (q,0)} \bar{A}_q^T \bar{E}_{qm}[\tau] B_m d_m[k - \tau] + \bar{A}_q^T \bar{w}_q[k]
\]
Zero forcing equalization

- Zero-interference: (2-dimensions, Interference from 3 subcarriers)
  - $\bar{A}_q^T (\bar{E}_{qm}[k]*B_m d_m[k]) = 0, \ m \neq q$
  - $\bar{A}_q^T (\bar{E}_{qq}[k]*B_q d_q[k]) = I\delta[k]$

DoF are increasing:
(S: streams, $N_T$: antennas tx, $N_R$: antennas rx)

- Processing on a per-subcarrier basis:

$$S \leq N_T, \ S < \frac{2}{3}N_R$$

$y_q[k] \xrightarrow{ZF} \tilde{d}_q[k]$
Zero forcing equalization

- Zero-interference: (2-dimensions, Interference from 3 subcarriers)
  - $\bar{A}_q^T (\bar{E}_{qm}[k] \star B_m d_m[k]) = 0, \ m \neq q$
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DoF are increasing:
(S: streams, $N_T$: antennas tx, $N_R$: antennas rx)

- Joint processing of $N$ multicarrier symbols on a per-subcarrier basis:

$$S \leq N_T, \ S \leq \frac{2}{3} N_R$$

Diagram:
- $y_q[k] \rightarrow \hat{d}_q[k]$
- $\vdots$
- $y_q[k-N+1] \rightarrow \hat{d}_q[k-N+1]$
Zero forcing equalization

- Zero-interference: (2-dimensions, Interference from 3 subcarriers)
  - \( \bar{A}_q^T (\bar{E}_qm[k] \ast B_m d_m[k]) = 0, m \neq q \)
  - \( \bar{A}_q^T (\bar{E}_{qq}[k] \ast B_q d_q[k]) = I \delta[k] \)

**DoF are increasing:**

(S: streams, \( N_T \): antennas tx, \( N_R \): antennas rx)

- Joint processing of \( N \) multicarrier symbols and two consecutive subcarriers:

  \[ S \leq N_T, \ S \leq N_R \]

\[ \begin{array}{c}
  y_{q-1}[k-1] \\
  \vdots \\
  y_q[k] \\
  \vdots \\
  y_{q}[k-N+1] \\
\end{array} \rightarrow \begin{array}{c}
  \tilde{d}_q[k] \\
  \vdots \\
  \vdots \\
  \vdots \\
  \tilde{d}_q[k-N+1] \\
\end{array} \]
Zero forcing equalization

- Zero-interference: (2-dimensions, Interference from 3 subcarriers)
  - $\bar{A}_q^T (\bar{E}_{qm}[k] \star B_m d_m[k]) = 0, m \neq q$
  - $\bar{A}_q^T (\bar{E}_{qq}[k] \star B_q d_q[k]) = I \delta[k]$

**DoF are increasing:**
(S: streams, $N_T$: antennas tx, $N_R$: antennas rx)

- Processing on a per-subcarrier basis:
  $$S \leq N_T, \quad S < \frac{2}{3}N_R$$

- Joint processing of $N$ multicarrier symbols on a per-subcarrier basis:
  $$S \leq N_T, \quad S \leq \frac{2}{3}N_R$$

- Joint processing of $N$ multicarrier symbols and two consecutive subcarriers:
  $$S \leq N_T, \quad S \leq N_R$$

A. I. Pérez-Neira and M. Caus "Dimensionality constraints imposed by highly frequency selective channels on MIMO-FBMC/OQAM systems," *in ITG Workshop on Smart Antennas (WSA), 3-5 March 2015*
For complexity reasons the zero-interference constraint is discarded.

It is possible to draw an analogy between the partial subcarrier overlapping and the inter-user interference in multi-user communication systems.
Iterative solutions

- Transmit-receive processing is governed by the MSE
  \[ \text{MSE}_q = \mathbb{E}\left\{ (\mathbf{d}_q[k] - \tilde{\mathbf{d}}_q[k]) (\mathbf{d}_q[k] - \tilde{\mathbf{d}}_q[k])^T \right\} \]

- MSE\(_q\) depends on \(\bar{\mathbf{A}}_q, \mathbf{B}_{q-1}, \mathbf{B}_q\) and \(\mathbf{B}_{q+1}\)

- Joint tx-rx design is challenging:
  - Subcarriers are coupled
  - MSE\(_q\) convex in \(\bar{\mathbf{A}}_q\) or \(\mathbf{B}_q\), but not jointly convex in \(\bar{\mathbf{A}}_q\) and \(\mathbf{B}_q\)

- Alternating optimization algorithms:
  - Optimize \(\{\bar{\mathbf{A}}_q\}\) fixing \(\{\mathbf{B}_q\}\)
  - Optimize \(\{\mathbf{B}_q\}\) fixing \(\{\bar{\mathbf{A}}_q\}\)

H. Shen, B. Li, M. Tao, and X. Wang "MSE-Based Transceiver Designs for the MIMO Interference Channel," *IEEE Transactions on Wireless Communications*, vol. 9, no.11, pp. 3480-3489, November 2010
Closed-form solutions

- ISI and ICI terms are upper bounded to get easy-to-handle expressions

\[
\| \bar{A}_q^T \bar{E}_{qm}[\tau] B_m \|_F^2 \leq \lambda_1 (B_m B_m^T) \| \bar{A}_q^T \bar{E}_{qm}[\tau] \|_F^2 \leq b_m \| \bar{A}_q^T \bar{E}_{qm}[\tau] \|_F^2
\]

\[
\text{tr} (BA) \leq \text{tr}(B) \lambda_1 (A), \quad \lambda_1 (B_m^T B_m) \leq b_m
\]

\[
\text{argmin}_{\{\bar{A}_q, B_q\}} f_0 (\{\text{UB}_q\})
\]

\[
\sum_{q=0}^{M-1} \text{tr} (B_q B_q^T) \leq P_T
\]

\[
\lambda_1 (B_m^T B_m) \leq b_m, \quad 0 \leq m \leq M - 1
\]

The bound is tight if

\[
1.25 \frac{P_T}{SM} \leq b_m \leq 2 \frac{P_T}{MS}
\]

\[
\text{UB}_q = \sum_{(m, \tau) \neq (q, 0)} b_m \bar{A}_q^T \bar{E}_{qm}[\tau] \left( \bar{A}_q^T \bar{E}_{qm}[\tau] \right)^T
\]

\[
+ \bar{A}_q^T R_{w_q} \bar{A}_q + I_S - 2 \bar{A}_q^T \bar{E}_{qq}[0] B_q + \bar{A}_q^T \bar{E}_{qq}[0] B_q \left( \bar{A}_q^T \bar{E}_{qq}[0] B_q \right)^T
\]
Numerical results

- Carriers $M=1024$
- $N_R > N_T = S = 2$
- Subcarrier spacing $\Delta_f=15$ KHz
- Sampling frequency $f_s=15.36$ MHz
- Extended Typical Urban channel. Delay spread: $\sigma = 991$ ns
- Active carriers to comply with the spectral mask:
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MIMO processing with no CSIT

No CSIT: Multiplexing

- Input/output relation when $\theta_q[k] = 1$:

$$y_q[k] = \sum_{(m,\tau)} \alpha_{qm}[\tau] \theta_m[k - \tau] H_q d_m[k - \tau] + w_q[k]$$

- 2-D Viterbi achieves the optimal performance: Alternatives:
  - MMSE equalizer: $\tilde{d}_q[k] = \Re(A_q^H y_q[k])$
  - Partial interference cancellation:

\[ r_q[k] = \sum_{\tau} \alpha_{qq}[\tau] \theta_q[k - \tau] H_q d_q[k - \tau] + w_q[k] \]

Alamouti scheme relies on a complex orthogonality whereas FBMC/OQAM has only a real orthogonality.

The Alamouti coding is performed in a block-wise manner inserting gaps in order to isolate the blocks.

This solution is feasible when the FBMC/OQAM impulse response is conjugate symmetric, i.e. $\alpha_{qm}[\tau] = \alpha_{qm}^*[−\tau]$.
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A centralized transmitter serves $N_U$ decentralized receivers.

The block diagonalization concept is used to eliminate inter-user interference. Remaining ICI and ISI can be dealt with the ZF [1].

Inter-user interference, ISI and ICI are mitigated using an iterative algorithm [2].


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Open problems:

- Optimal joint tx-rx design that takes into account:
  - ISI, ICI and the non-circular nature of the data
- DoF study in multi-user communication systems
- IA techniques
- MISO/SIMO/MIMO MSE-duality for FBMC/OQAM systems
- Take advantage of FDSS [1]
- Low-complexity design that perform close to the ML detector when CSIT is not available

Conclusion:

- The application of MIMO to FBMC/OQAM is proven to be spectrally efficient
- Signal processing is crucial in FBMC/OQAM to manage interference

References


[12] A. I. Pérez-Neira and M. Caus "Dimensionality constraints imposed by highly frequency selective channels on MIMO-FBMC/OQAM systems," in ITG Workshop on Smart Antennas (WSA), 3-5 March 2015


THANK YOU FOR YOUR ATTENTION

ana.isabel.perez@upc.edu
http://www.cttc.es/people/aperez/
Enhanced OFDM

- Side lobe suppression techniques [1]:
  - **Time domain windowing**: redundancy increased wrt OFDM
  - **Cancellation carriers**: nonlinear optimization with quadratic inequality
  - **Polynomial cancellation coding**: spectral efficiency is reduced by the factor of 2
  - **Subcarrier weighting**: nonlinear optimization with quadratic equality and linear inequality

- Data symbol density: \( \frac{1}{T \Delta f} = \frac{T}{T + T_{CP}} \)

Filtered Multitone - FMT

- Originally conceived for DSL [1]

Pros:
  - pulses satisfy the Nyquist constraint for transmission without ISI
  - pulses are not allowed to overlap in the frequency domain

Cons:
  - Channel destroys orthogonality
  - Overhead in burst transmission
  - Maximum bandwidth efficiency is not achieved

Data symbol density: \[ \frac{1}{T \Delta f} = \frac{T}{(1+\alpha)T} \]

Generalized frequency division multiplexing - GFDM

- Transmission in blocks with one CP shared by $L$ symbols [1]

- Pros:
  - Low overhead in burst transmission by replacing linear filtering with circular filtering

- Cons:
  - GFDM is not an orthogonal system
  - Maximum bandwidth efficiency is not achieved

- Data symbol density:
  \[
  \frac{1}{T \Delta f} = \frac{T}{T + \frac{T_{CP}}{L}}
  \]

Circular filtering concept is further adopted for the FMT [1]

Pros:
- The kernel remains FMT, so orthogonality can be achieved
- CP can be optionally used to simplify equalization
- Low overhead in burst transmission

Cons:
- Maximum bandwidth efficiency is not achieved

Data symbol density:
\[
\frac{1}{T \Delta_f} = \frac{T}{(1+\alpha) \left( T + \frac{T_{CP}}{L} \right)}
\]

Filter bank multicarrier/OQAM - FBMC/OQAM

- Real and imaginary parts are delayed half the symbol period [1]

- Pros:
  - Maximum bandwidth efficiency is achieved
  - Orthogonality is satisfied in the real field

- Cons:
  - Channel destroys the orthogonality
  - Overhead in burst transmission

- Circular convolution can be adopted to remove edge transitions [2]

- Data symbol density: \( \frac{1}{T \Delta f} = 1 \)
