Optimized reconfigurable fast convolution based transmultiplexers for flexible radio access

Optimized Reconfigurable Fast Convolution based Transmultiplexers for Flexible Radio Access

Juha Yli-Kaakinen and Markku Renfors

***Abstract*—Multirate fast-convolution (FC) processing can be used for realizing low-complexity filter banks (FBs) and transmultiplexers (TMUXs). The main advantage of the FC based realizations when compared with the polyphase FBs is the increased configurability, that is, the number of subchannels, their bandwidths, and the center frequencies can be adjusted independently. In general, FC based FBs are linear periodically shift-variant systems. In this paper, novel matrix representations for the FC synthesis and analysis FBs are first derived. These representations give all the shift-variant impulse responses of the FC based FBs. Then the TMUX optimization criteria are expressed using these representations. Two examples are included to demonstrate the performance of the optimized designs as well as to illustrate the flexibility of the resulting FC based TMUXs.**

***Keywords*—Multirate signal processing, filter banks, fast-convolution, waveforms, multicarrier, single-carrier, 5G**

I. Introduction

THE ever increasing growth in wireless communications necessitates improving the effectivity in radio spectrum utilization. Dynamic spectrum allocation is one of the key technologies where, for instance, the radio spectrum left unused by the primary (licensed) user is dynamically aggregated for high-speed communications [1]. In such cases, high flexibility (agility) of the signal processing techniques and the generated waveforms becomes crucial. In order to avoid undesired interferences between the primary users and the dynamic spectrum users, narrow transition bands and good frequency selectivity of the channelization filters are also essential requirements.

For uniform filter banks (FBs), cosine or exponential modulation is typically used for generating the FB responses based on a lowpass prototype filter [2], [3]. This approach has been extended also for realizing fixed non-uniform FBs with near perfect reconstruction (NPR) characteristics [4], [5]. Fast-convolution filter bank (FC-FB) is an efficient implementation for realizing highly tunable multirate FB configurations [6] with arbitrary bandwidths. FC-FBs are based on forward-inverse discrete Fourier transform (DFT) pair with overlapped block processing and their frequency-domain (FD) characteristics can be straightforwardly adjusted, even in real time. Each subband can be easily configured for unequal bandwidths, different center frequencies, and adjustable sampling rate conversion factors, including also full- or partial-band NPR systems. Such FB systems find various applications, for example, as flexible channelization filters for software defined radios. Simultaneous waveform processing for multiple single-carrier (SC) and/or multicarrier (MC) transmission channels with different bandwidths and subcarrier spacings is also possible by using FC-FBs.

The basic idea of FC-based filters and FBs is that the convolution can be implemented effectively through multiplication in frequency domain, after taking DFTs of the input sequence and the filter impulse response. The resulting time-domain output signal is obtained by inverse discrete Fourier transform (IDFT) [6]. In practice, computationally efficient implementation techniques, like fast Fourier transform (FFT) and inverse fast Fourier transform (IFFT), are used for realizing the transforms. In order to process long sequences, FC processing is carried out blockwise in conjunction with overlap-add or overlap-save processing [7]. In this contribution, the latter one is used due to fact that overlap-add processing may induce undesirable transients in finite wordlength implementations.

The idea of flexible FB-implementation of NPR FB systems, with applications in filter bank multicarrier waveforms, has been in introduced in [8] and the analysis and FC-FB optimization methods are developed in [6]. For the application of FC to multirate filters, see [9] and the references therein whereas the FC-based filtered waveform processing for 5th generation new radio (5G) is considered in [10]. This contribution develops a matrix model and novel tools for the effective transfer function analysis and optimization of FC based FBs and TMUXs, explores the variability of the FC processing, as well as reports the optimized designs versus direct square root raised cosine (RRC) based designs and original designs reported in [6]. The proposed model is shown to be a flexible tool for FB based SC and MC waveform optimization in terms of inter-symbol interference (ISI) and inter-carrier interference (ICI).

This paper is organized as follows. In Section II, TMUXs are briefly described, the linear periodically shift variant (LPSV) models of the FC based synthesis and analysis FBs are derived, and the optimization problems for designing optimized FC based TMUXs are stated. The examples demonstrating the performance of the optimized designs are included in Section III and Section IV summarizes this paper.

II. Multirate Filters and Filter Banks

The synthesis-analysis FB pair, i.e., TMUX configuration as shown in Fig. 1 consist of synthesis FB, transmission channel, and analysis FB. In this system, each of the M incoming signals is first upsampled by a factor of I_m and filtered by synthesis filters $F_m(z)$ for $m = 0, 1, \ldots, M-1$. Then the outputs are added to form a single output signal. The resulting output signal is transmitted through the transmission channel $D(z)$ to the analysis part. Finally, the subband signals are reconstructed

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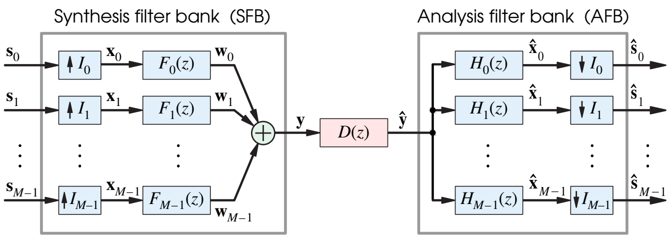

Fig. 1. General M-channel transmultiplexer with configurable sampling rate conversion factors.

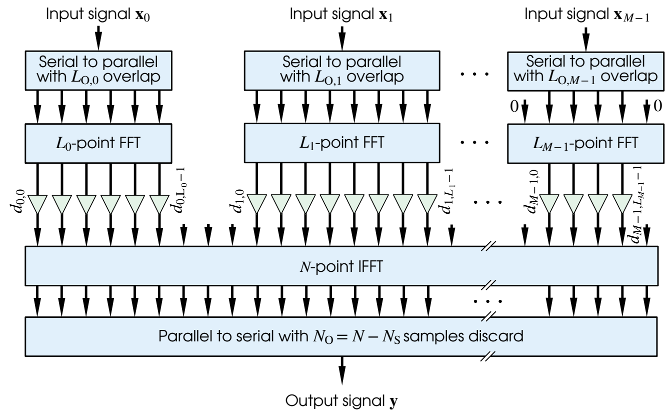

Fig. 2. FC based synthesis FB using overlap-save processing.

with the aid of analysis filters $H_m(z)$ for $m = 0, 1, \ldots, M-1$ followed by downsamplers.

In principle, the synthesis and analysis FBs can be designed such that the perfect reconstruction (PR) of the transmitted signals is guaranteed at the receiving end. In the case of ideal transmission channel, the recovered output signals are simply the delayed versions of those of the transmitted ones. Alternatively, the TMUX configuration may provide NPR characteristics. In this case, some of FB structure inflicted intrinsic interference is present even in the case of ideal channel. In practice, the level of this interference can be reduced by increasing the length of the prototype filter, and restricted to be insignificantly small when compared with the distortions induced by the transmission channel.

A. *Fast-Convolution Filter Banks*

The structure of the adjustable FC synthesis FB (SFB) is illustrated in Fig. 2. The general idea of this structure is a multirate version of FC [11], [12]. In the synthesis FB case, multiple low-rate, narrowband signals $\mathbf{x}_m$, for $m = 0, 1, \ldots, M-1$, are to be filtered with adjustable frequency responses and possibly adjustable sampling rates and then combined into single wideband signal $\mathbf{y}$, following the frequency-division multiplexing principle. Typically, these input signals are critically sampled or oversampled by a small factor. It is also possible that the subbands are overlapping. The dual structure of Fig. 2 can be used for splitting the incoming high-rate, wideband signal into several low-rate narrowband signals.

In this structure, each of the M incoming signals is first segmented into overlapping blocks of length L_m for $m = 0, 1, \ldots, M-1$. Then, each input block is transformed to FD using DFT of length L_m. The FD bin values of the converted signal are multiplied by the weight values corresponding to the DFT of the finite-length linear filter impulse response, $d_{m,\ell} = \sum_{n=0}^{L_m-1} h_m[n] \mathrm{e}^{-\mathrm{j}2\pi(\ell+L_m/2)n/L_m}$ for $m = 0, 1, \ldots, M-1$ and $\ell = 0, 1, \ldots, L_m - 1$. Here, m is the subband index and ℓ is the DFT bin index within the subband. For convenience of notation, we use the "FFT-shifted" indexing scheme in this context, i.e., index 0 corresponds to the lower edge of the subband. Finally, the weighted signals are combined and converted back to time-domain using IDFT of length N and the resulting time-domain output blocks are concatenated using the overlap-save principle. [13], [14].

The multirate FC-processing of Fig. 2 increases the sampling rates of the subband signals by the factors of

$$I_m = N/L_m. \tag{1}$$

Given the IDFT length N, the sampling rate conversion factor is determined by the DFT length L_m, and it can be configured for each subband individually. Naturally, L_m determines the maximum number of non-zero frequency bins, i.e., the bandwidth of the subband.

The interpolation factor of the structure of Fig. 2 is given by (1) [11]. The FC output block length N_S is increased by the same factor and so is the overlapping part length N_O in overlap-save processing. The overlapping and non-overlapping input block lengths $L_{\mathrm{O},m}$ and $L_{\mathrm{S},m}$, respectively, and the corresponding output block lengths N_O and N_S, have to exactly match when taking into account the interpolation factor. Generally, $N = p\Gamma$ and $N_\mathrm{S} = q\Gamma$, where p and q are two relatively prime integers and $\Gamma = \gcd(N, N_\mathrm{S})$, where $\gcd(\cdot)$ is the greatest common divisor. Then for narrowest possible subband case satisfying the integer-length criterion, $L_m = p$ and $L_{\mathrm{S},m} = q$. Generally, L_m has to be a multiple of N/Γ, that is, the configurability of the output sampling rate depends greatly on the choice of N and N_S.

In [6], the performance of the FC-FB is analyzed using a periodically time-variant model and effective tools for frequency response analysis and FC-FB optimization are developed. It is also shown that the complexity of the FC-FB is considerably smaller when compared with the traditional polyphase implementations of FBs.

B. *Linear Periodically Shift-Variant Model of the FC-FB*

In the FC SFB case, the block processing of mth subcarrier signal $\mathbf{x}_m$ for the generation of high-rate subband waveform $\mathbf{w}_m$ can be represented as

$$\mathbf{w}_m = \mathbf{F}_m \mathbf{x}_m, \tag{2a}$$

where $\mathbf{F}_m$ is the block diagonal transform matrix of the form

$$\mathbf{F}_m = \mathrm{diag}(\mathbf{F}_{m,0}, \mathbf{F}_{m,1}, \ldots, \mathbf{F}_{m,R_m-1}) \tag{2b}$$

with R_m blocks. Here, the dimensions and locations of the $\mathbf{F}_{m,r}$'s are determined by the overlapping factor of the overlap-save processing, defined as

$$\lambda = 1 - L_{\mathrm{S},m}/L_m = 1 - N_\mathrm{S}/N, \tag{3}$$

where $L_{S,m}$ and N_S are the number of non-overlapping input and output samples, respectively.

The multirate version of the FC SFB can be represented using block processing by decomposing the $\mathbf{F}_{m,r}$'s as the following $N_S \times L_m$ matrix

$$\mathbf{F}_{m,r} = \mathbf{S}_N \mathbf{W}_N^{-1} \mathbf{M}_{m,r} \mathbf{D}_m \mathbf{P}_{L_m}^{(L_m/2)} \mathbf{W}_{L_m}. \quad (4)$$

Here, $\mathbf{W}_{L_m}$ and $\mathbf{W}_N^{-1}$ are the $L_m \times L_m$ DFT matrix (with $[\mathbf{W}_{L_m}]_{p,q} = \mathrm{e}^{-\mathrm{j}2\pi(p-1)(q-1)/L_m}$ for $p = 1, 2, \ldots, L_m$ and $q = 1, 2, \ldots, L_m$) and $N \times N$ inverse DFT matrix, respectively. The DFT shift matrix $\mathbf{P}_{L_m}^{(L_m/2)}$ is circulant permutation matrix obtained by cyclically left shifting the $L_m \times L_m$ identity matrix by $L_m/2$ positions. $\mathbf{D}_m$ is the $L_m \times L_m$ diagonal matrix with the FD weights of the subband m on its diagonal whereas $\mathbf{M}_{m,r}$ and $\mathbf{S}_N$ are the FD mapping and time-domain selection matrices, respectively. The $N \times L_m$ FD mapping matrix maps L_m FD bins of the input signal to FD bins $(c_m - \lceil L_m/2 \rceil + \ell)_N$ for $\ell = 0, 1, \ldots, L_m - 1$ of the output signal. Here c_m is the center bin of the subband m and $(\cdot)_N$ denotes the modulo-N operation. In addition, this matrix rotates the phases of the block by

$$\theta_m(r) = \exp(\mathrm{j}2\pi r \theta_m) \quad \text{with} \quad \theta_m = c_m L_{S,m}/L_m \quad (5)$$

in order to maintain the phase continuity between the consecutive overlapping processing blocks [6], [11]. The $N_S \times N$ selection matrix $\mathbf{S}_N$ selects the desired N_S output samples from the inverse transformed signal corresponding to overlap-save processing.

In general, the bandwidth of each subband can be easily tuned by adjusting the L_m FD weights in $\mathbf{D}_m$. Here, the maximum available bandwidth is determined by L_m. On the other hand, the center frequency of the subband can be adjusted by only cyclically shifting the columns of the FD mapping matrix $\mathbf{M}_{m,r}$ to their desired locations (with the corresponding phase rotations) whereas the interpolation factor is determined simply by the ratio of N and L_m. Therefore, this FC SFB provides an easily configurable basis for flexible multichannel channelization FB for flexible radio access.

In the analysis filter bank (AFB) case, the corresponding analysis sub-block matrix of size $L_{S,m} \times N$ can be decomposed as

$$\mathbf{G}_{m,r} = \mathbf{S}_{L_m} \mathbf{W}_{L_m}^{-1} \mathbf{P}_N^{(N/2)} \mathbf{D}_m \mathbf{M}_{m,r}^T \mathbf{W}_N, \quad (6)$$

where $\mathbf{P}_N^{(N/2)}$ is the inverse Fourier-shift matrix and the $L_{S,m} \times L_m$ selection matrix $\mathbf{S}_{L_m}$ selects the desired $L_{S,m}$ output samples from the inverse transformed output signal.

In general, the above FC based synthesis and analysis filter banks are LPSV systems with period of $L_{S,m}$, that is, the systems have $L_{S,m}$ different impulse responses. In the SFB case, the impulse responses are given by the $L_{S,m}$ shift-variant columns of the $\mathbf{F}_m$ as illustrated Fig. 3.

C. Transmultiplexer Optimization

The $P \times L_{S,m}$ sub-matrix $\mathbf{F}_m^{(\mathrm{sub})}$ with $P = N_S B_F$ and $B_F = \lceil N/N_S \rceil$ containing all the $L_{S,m}$ shift-variant impulse responses of the SFB can be obtained by selecting the desired rows and columns from $\mathbf{F}_m$ as follows:

$$[\mathbf{F}_m^{(\mathrm{sub})}]_{p,q} = [\mathbf{F}_m]_{p,q+S_{F,m}} \quad (7)$$

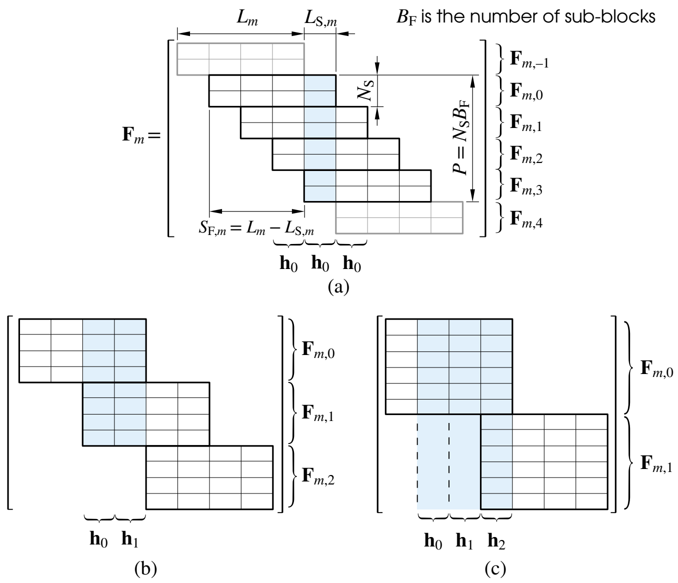

Fig. 3. Structure of block-diagonal synthesis matrix $\mathbf{F}_m$ for $L_m = 4$, $N = 8$, and $L_{S,m} = 1, 2, 3$. The colored columns illustrate the elements of the synthesis matrix that form the shift-variant impulse responses of the synthesis filter bank. (a) For $L_{S,m} = 1$, the system has only one impulse response of eight samples. (b) For $L_{S,m} = 2$, the system has two eight-samples long responses. (c) For $L_{S,m} = 3$, two among the three responses are six samples long and the remaining one is 12 samples long.

with $S_{F,m} = L_{O,m} = L_m - L_{S,m}$ for $p = 1, 2, \ldots, P$ and for $q = 1, 2, \ldots, L_{S,m}$. These sub-matrices are denoted in Fig. 3 by colored areas.

For the AFB case, the corresponding $Q_m \times P$ sub-matrix $\mathbf{G}_m^{(\mathrm{sub})}$ with $Q_m = L_{S,m} B_G$ and $B_G = \lfloor (P + N)/N_S - (B_F)_2 \rfloor$ containing all the possible impulse responses is given by

$$[\mathbf{G}_m^{(\mathrm{sub})}]_{q,p} = [\mathbf{G}_m]_{q,p+S_G} \quad (8)$$

with $S_G = \lfloor (N + [B_G - 1] N_S - P)/2 \rfloor$ for $q = 1, 2, \ldots, Q_m$ and for $p = 1, 2, \ldots, P$.

For TMUX, the transfer function from subchannel m to subchannel n is expressed as

$$\mathbf{T}_{m,n} = \mathbf{G}_n^{(\mathrm{sub})} \mathbf{F}_m^{(\mathrm{sub})}. \quad (9)$$

In the communications signal processing, the reconstruction error, i.e., the difference between the input signal and the (possibly delayed) output signal on subchannel m can be considered as ISI. In this case, the mean-squared ISI can be expressed as

$$\varepsilon_{\mathrm{ISI}} = \left\| \mathbf{I} - \mathbf{T}_{m,m} \right\|_F^2, \quad (10)$$

where $\mathbf{I} = \begin{bmatrix} \mathbf{0}_{U_m \times L_{S,m}} & \mathbf{E}_{L_{S,m}} & \mathbf{0}_{V_m \times L_{S,m}} \end{bmatrix}^{\mathsf{T}}$ with $U_m = \lfloor S_G L_m/N \rfloor$ and $V_m = P - U_m - L_{S,m}$. Here, $\mathbf{0}_{U_m \times L_{S,m}}$ and $\mathbf{0}_{V_m \times L_{S,m}}$ are the $U_m \times L_{S,m}$ and $V_m \times L_{S,m}$ zero matrices, respectively, $\mathbf{E}_{L_{S,m}}$ is the $L_{S,m} \times L_{S,m}$ identity matrix and $\|\cdot\|_F^2$ is the squared Frobenius norm. Correspondingly, the mean-squared ICI can be considered as a interference from input subchannel m to all the output subchannels n for $n \neq m$ as expressed as

$$\varepsilon_{\mathrm{ICI}} = \sum_{n=0, n\neq m}^{M-1} \left\|\mathbf{T}_{n,m}\right\|_{\mathrm{F}}^{2}. \tag{11}$$

The FC-FB TMUX design can be stated as a problem for finding the optimal values of the FD window (diagonal of $\mathbf{D}_m$ in (4) and (6)) to minimize

$$\varepsilon = \gamma\varepsilon_{\mathrm{ISI}} + \varepsilon_{\mathrm{ICI}}, \tag{12}$$

where γ is the desired weighting factor for trading between the ISI and ICI whereas $\varepsilon_{\mathrm{ISI}}$ and $\varepsilon_{\mathrm{ICI}}$ are given by (10) and (11), respectively. In the following, we have used $\gamma = 1$.

Here the FD weights in $\mathbf{D}_m$ have a RRC-type response and the number of free parameters in the optimization is $\lceil L_m/2 \rceil$ when taking into account the symmetrical transition bands. For certain bandwidth, the optimization is independent of the number of subchannels required since the same FD weights can be replicated for all the subchannels. Therefore, this problem can be straightforwardly solved using Levenberg-Maquardt type optimization algorithms [15] in less the ten seconds.[1]

In this approach, the channel model is assumed to be ideal in the optimization. In the real-time application, the FD weights are properly combined with the equalizer coefficient values for realizing frequency-sampled subcarrier-wise equalizer without the need of re-optimization of the coefficient values. For details, see e.g. [17].

In the case of FBMC/offset quadrature amplitude modulation (OQAM), the $\mathbf{T}_{m,n}$'s have to be downsampled by two following the OQAM post-processing principle [18].

III. Numerical Examples

A. Example 1

The goal of the Example 1 is to illustrate the improvement in performance when comparing the analytical unoptimized RRC-type prototype filter with the optimized one as well as to compare the original design scheme in [6] with the proposed one. Here, we focus on a 5 MHz long-term evolution (LTE) like MC system utilizing the FBMC/OQAM waveform and using FC-FB based TMUX. In this case, the number of subcarriers is $M = 512$ out of which $M_{\mathrm{act}} = 300$ active subcarriers are used. These active subcarriers are scheduled in resource blocks (RBs) of 12 subcarriers, i.e., there are 25 RBs. In time domain, the resources are scheduled to different users in 1 ms subframes of 2 RBs.

Here the approach for selecting the main parameters of the FC-FB scheme is to use the same sampling rate, the same subcarrier spacing, and the same parameters for the RBs as in the 5 MHz LTE system. For clarity of the discussion, we focus here on the SFB of the transmitter. The subcarrier spacing is $L_m/2$ IFFT bins and, consequently, the IFFT length can be determined as $N = ML_m/2$. The choice of the IFFT length, together with the overlap factor, defines the time interval of each processing block and has a significant effect on the performance of the scheme with fast-fading channels. However, this depends greatly on the used channel equalization structure [17]. On the other hand, increasing L_m gives better changes to shape the spectrum to improve spectral containment. In any case, we assume that the N should be shorter than the RB length and we are looking for a good trade-off between the spectrum control and fading sensitivity. A third element in the trade-off is implementation complexity. Increasing the overlap factor improves the spectral containment for a given L_m but it also increases the computational complexity as the ratio of the useful part of IFFT to the IFFT length is reduced. The implementation complexity evaluations are available in [6].

For the two-times oversampled case, the suboptimal initial design can be obtained by simply using the RRC-type filter with roll-off of one. Fig. 4 compares the performance of the optimized RRC-type designs with the initial designs in terms of simulated MSE between the transmitted and received signals. The well-established PHYDYAS polyphase FBMC design [18] is here shown for reference. As can be seen from this figure, the length of the overlapping part $L_{\mathrm{O},m} = L_m - L_{\mathrm{S},m}$ determines the minimum achievable MSE and the performance is similar for the different values of L_m. In addition, the unoptimized RRC design is nearly optimal for the values of the overlapping part smaller than or equal to four and the overlap has to be at least six samples for FC-FB design to perform better than PHYDYAS FBMC with $K = 3$ or nine samples for $K = 4$. Here, K is the overlapping factor of the polyphase filter bank.

Table I gives the average improvement in MSE when comparing the proposed designs with the original ones [6]. These values are evaluated by averaging the difference in MSE for all possible values of overlap factor. However, it should

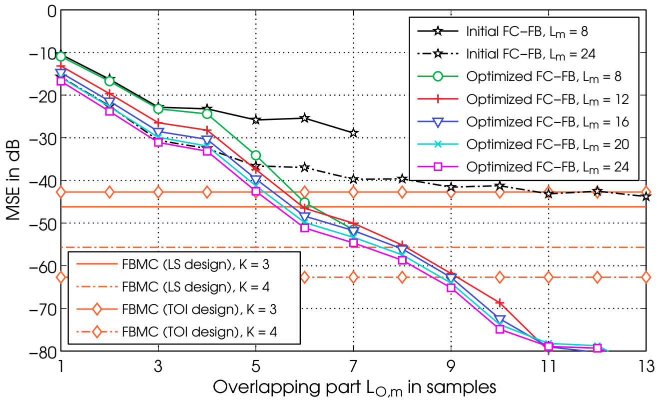

Fig. 4. Simulated MSE as a function of the overlapping part $L_{\mathrm{O},m} = L_m - L_{\mathrm{S},m}$ of the input samples for FC-FB with $L_m = 8, 12, 16, 20, 24$. For comparison the MSE of the initial design for $L_m = 8$ and $L_m = 24$ as well as the PHYDYAS polyphase FBMC [18] designs with $K = 3$ and $K = 4$ are also shown. Here, "LS design" and "TOI design" refer to designs for which the stopband energy and total FB structure based interference, respectively, are minimized [18].

TABLE I. Average improvement in MSE when comparing the proposed designs with the original ones [6]

$L_m = 8$	$L_m = 12$	$L_m = 16$	$L_m = 20$	$L_m = 24$
2.61 dB	3.16 dB	5.88 dB	9.58 dB	14.85 dB

[1] The so-called frequency-spreading filter bank multicarrier (FBMC) [16] can be seen as a special case of the FC-FB when $L_{\mathrm{S},m} = 1$ ($L_{\mathrm{O},m} = L_m - 1$) and, therefore, also these filters can be efficiently optimized using this approach.

be pointed out that the major contribution of the difference comes for values of overlapping part $L_{O,m}$ greater than ten. The optimization time for the original algorithm is typically less than five minutes.

B. Example 2

This example illustrates the configurability of the FC processing and the quality of the optimized designs by realizing an hypothetical TMUX system such that adjacent subbands may have different modulations, roll-off factors, subchannel bandwidths, and subcarrier spacings. Assume that it is desired to use the guard subcarriers of the 5 MHz LTE system of Example 1 for carrying SC waveforms such that the spectral leakage effects between the SC and LTE systems are minimized. Here, the parameters of the LTE system are $N = 8192$, $L_m = 16$, and $L_{O,m} = 8$ for $m = 0, 1, \ldots, M_{\text{act}} - 1$ with $M_{\text{act}} = 300$. The lower guard subcarriers are occupied by eight SC waveforms with bandwidth of 96 IFFT bins and subcarrier spacing of 105 IFFT bins as well as the roll-off of $\alpha = 1/12$. Here, 96 IFFT bins with $L_m/2$ subcarrier spacing corresponds to the 12 subcarriers (1 RB). On the upper guard subcarriers there are four SC signals with bandwidth and subcarrier spacing of 192 (2 RBs) and 201 IFFT bins, respectively, and $\alpha = 1/32$. The FFT lengths used for generating the lower and upper SC signals are $L_m = 192$ and $L_m = 384$, respectively.

Fig. 5 shows the analytical magnitude-squared response of the analysis or the synthesis FB and the simulated MSE on active subcarriers. The simulated MSE between the transmitted and received quadrature phase-shift keying (QPSK) signals on LTE subcarriers is practically the same as in Example 1, whereas the corresponding values for the lower and upper SC signals are −69.84 dB and −72.95 dB, respectively. This means that the spectral containment of the optimized TMUXs with the given parametrization is sufficient to totally avoid the spectral leakage between the SC transmissions and LTE subcarriers.

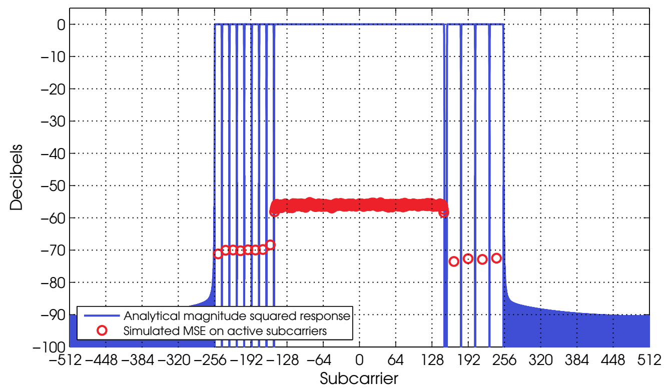

Fig. 5. Analytical magnitude-squared response of the synthesis/analysis FB as well as the simulated MSE on the active subcarriers.

IV. Summary

This paper proposes an effective matrix model for the analysis and optimization of the fast-convolution based transmultiplexers. It is shown through simulations that the optimized designs result in improved performance when compared with the original designs. The developed matrix formalism can used as a basis for a generic FC-FB optimization tool taking into account different performance metrics: in-band distortion (ISI), ICI in synchronous multicarrier operation, as well as out-of-band emission/power leakage effects between adjacent asynchronous spectral components.

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