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# Reconfigurable Multirate Systems in Cognitive Radios

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## 1. Introduction

One perspective in communication systems is to increase the spectrum utilization using cognitive radios. A cognitive radio is a network of intelligent co-existing radios which senses the environment to find the available frequency slots, white spaces, or the spectrum holes as noted in Akyildiz et al. (2008); Haykin (2005). Then, it modifies its transmission characteristics to use that particular frequency slot. Figure 1 illustrates the overlay spectrum sharing outlined in Cabric et al. (2006), or the opportunistic spectrum access discussed in Zhao & Sadler (2007); Zhao & Swami (2007), or the dynamic spectrum access which is considered in Sherman et al. (2008). Here, secondary users occupy the frequency slots which are not used<sup>1</sup> by the primary users. One of the main tasks in a cognitive radio is thus the spectrum mobility as in Akyildiz et al. (2006; 2008), or the dynamic frequency allocation as in Haykin (2005), or the dynamic spectrum allocation as in Leaves et al. (2004); Zhao & Swami (2007). This chapter uses the term dynamic frequency-band allocation (DFBA). Being dynamic means that the transmission parameters, e.g., bandwidth, center frequency, transmission power, and communication standard etc., may vary with time according to Akyildiz et al. (2006). One should at least be able to change the center frequency and bandwidth although other parameters may also change. This is also referred to as the reconfigurability according to Akyildiz et al. (2008); Haykin (2005); Jondral (2005); Leaves et al. (2004); Ramacher (2007); Sherman et al. (2008).

Another perspective in communication systems calls for satellites to play a complementary role in supporting various wideband services as proposed by Arbesser-Rastburg et al. (2002); Evans et al. (2005); Farserotu & Prasad (2000); Lucente et al. (2008); Nguyen et al. (2002); Re & Pierucci (2002); Wittig (2000). For this purpose, the European space agency has proposed three major network structures for broadband satellite-based communication systems as in Arbesser-Rastburg et al. (2002). This requires an efficient use of the limited available frequency spectrum by satellite on-board signal processing as discussed in Abdulazim & Göckler (2005a;b; 2006); Abdulazim et al. (2007); Arbesser-Rastburg et al. (2002); Eghbali et al. (2007a;b; 2009a); Eghbali, Johansson, Löwenborg & Göckler (2011); Evans et al. (2005); Farserotu & Prasad (2000); Göckler & Abdulazim (2005; 2007);

<sup>1</sup> Under certain conditions, the secondary users need not wait for a vacant channel. This allows a simultaneous transmission over the same time or frequency as noted in Devroye et al. (2006).

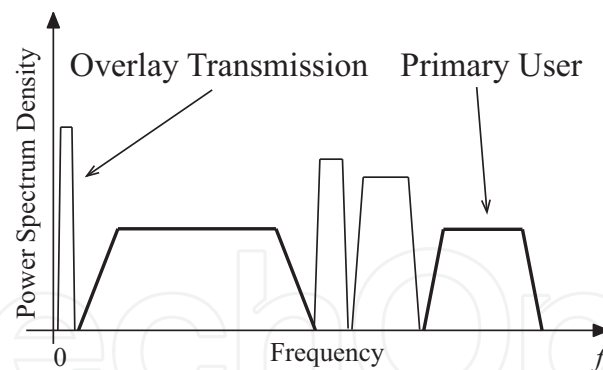


Fig. 1. Overlay approach for spectrum sharing.

Johansson & Löwenborg (2005; 2007); Lippolis et al. (2004); Lucente et al. (2008); Nguyen et al. (2002); Re & Pierucci (2002); Rosenbaum et al. (2006); Wittig (2000).

Like satellite-based communication systems which require both on-ground DFBA and on-board dynamic frequency-band reallocation (DFBR), the ad hoc- or infrastructure-based cognitive radios can also utilize DFBA and DFBR. In the ad hoc-based networks, individual users can utilize DFBA while DFBR can be performed by the base stations of infrastructure-based networks. The DFBA can also be deployed by the individual users of infrastructure-based networks. Both DFBA and DFBR need interpolation/decimation with variable parameters. For large sets of variable conversion factors, the implementation complexity increases. Complexity reduction can be achieved using reconfigurable structures which perform various tasks by simple modifications and without hardware changes. Also, the filter coefficients do not change thereby enabling us to solve the filter design problem only once and offline. Specifically, one must be able to reprogram the same hardware.

This chapter discusses the structure, reconfiguration, and the parameter selection when adopting the DFBA and DFBR for cognitive radios. Two approaches, i.e., Approach I and II, are discussed. They are appropriate based on the availability of (i) a composite signal comprising several user signals, or (ii) the individual user signals. Combinations of Approaches I and II provide increased freedom to allocate and reallocate the user signals.

## 2. Basics of multirate signal processing

This section treats some basics of sampling rate conversion (SRC), filter banks (FBs), perfect reconstruction (PR), and transmultiplexers (TMUXs).

### 2.1 Sampling rate conversion: conventional

Different parts of a multirate system operate at different sampling frequencies thereby necessitating interpolation (decimation) to increase (decrease) the sampling frequency of digital signals as outlined in Mitra (2006); Vaidyanathan (1993). Interpolation and decimation comprise lowpass filters as well as upsamplers and downsamplers whose block diagrams are shown in Fig. 2. In Fig. 2(a),

$$y(n) = x(nM) \iff Y(z) = \frac{1}{M} \sum_{k=0}^{M-1} X(z^{\frac{1}{M}} W_M^k), \quad W_M = e^{-j\frac{2\pi}{M}}. \quad (1)$$

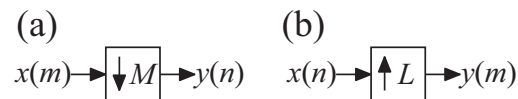


Fig. 2. (a)  $M$ -fold downsampler. (b)  $L$ -fold upsampler.

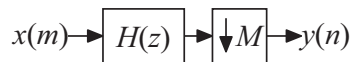


Fig. 3. Decimation by  $M$ .



Fig. 4. Interpolation by  $L$ .

Note that  $X(z^{\frac{1}{M}})$  is not periodic by  $2\pi$  but adding the shifted versions gives a signal  $Y(z)$  with a period of  $2\pi$  such that the Fourier transform can be defined. In Fig. 2(b),

$$y(n) = \begin{cases} x(\frac{n}{L}) & \text{if } n = 0, \pm L, \pm 2L, \dots \\ 0 & \text{otherwise} \end{cases} \iff Y(z) = X(z^L). \quad (2)$$

The upsampler and downsampler are linear time-varying systems. Unless  $x(n)$  is lowpass<sup>2</sup> and bandlimited, downsampling results in aliasing and decimation thus requires an additional filter as in Fig. 3. This anti-aliasing filter  $H(z)$  limits the bandwidth of  $x(n)$ . In Fig. 3,

$$y(n) = \sum_{k=-\infty}^{+\infty} x(k)h(nM - k). \quad (3)$$

As upsampling causes imaging, interpolation requires a filter as in Fig. 4. This lowpass anti-imaging filter  $H(z)$  removes the images and, as in Vaidyanathan (1993), we have

$$y(n) = \sum_{k=-\infty}^{+\infty} x(k)h(n - kL). \quad (4)$$

For SRC<sup>3</sup> by a rational ratio  $\frac{M}{L}$ , interpolation by  $L$  must be followed by decimation by  $M$ . Consequently, the cascade of the anti-imaging and anti-aliasing filters results in one filter, say  $G(z)$ , where the output, according to Vaidyanathan (1993), is

$$y(n) = \sum_{k=-\infty}^{+\infty} x(k)g(nM - kL). \quad (5)$$

Generally,  $G(z)$  is a lowpass filter with a stopband edge at, as in Mitra (2006); Vaidyanathan (1993),

$$\omega_s T = \min\left(\frac{\pi}{M}, \frac{\pi}{L}\right) = \frac{\pi}{\max(M, L)}. \quad (6)$$

In practice, there is a roll-off factor  $0 \leq \rho \leq 1$  so that  $\omega_s T = \frac{\pi(1+\rho)}{\max(M, L)}$ . If  $M$  and  $L$  are mutually coprime numbers, a decimator can be obtained by transposing the interpolator. The noble identities, defined as in Fig. 5, help move the filtering operations inside a multirate structure.

<sup>2</sup> This is not necessary to avoid aliasing. For example, if  $X(e^{j\omega T})$  is nonzero only at  $\omega T \in [\omega_1 T, \omega_1 T + \frac{2\pi}{M}]$  for some  $\omega_1 T$ , there is no aliasing.

<sup>3</sup> If  $L > M$  ( $L < M$ ), we have interpolation (decimation) by a rational ratio  $\frac{L}{M} > 1$  ( $\frac{M}{L} > 1$ ). This chapter frequently refers to SRC by a rational ratio  $R_p > 1$ .

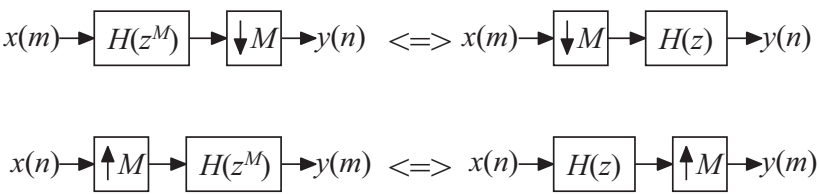


Fig. 5. Noble identities which allow us to move the arithmetic operations to the lower sampling frequency.

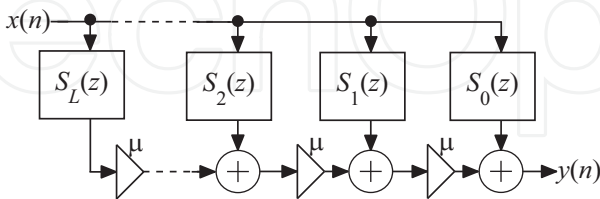


Fig. 6. Farrow structure with fixed subfilters  $S_k(z)$  and variable fractional delay (FD) of  $\mu$ .

2.2 Sampling rate conversion: Farrow structure

In a conventional SRC and if the SRC ratio changes, we need new filters thereby reducing the flexibility. The Farrow structure, introduced in Farrow (1988) and shown in Fig. 6, obtains flexibility in an elegant way. The Farrow structure is traditionally composed of linear-phase finite-length impulse response (FIR) subfilters  $S_k(z)$ ,  $k = 0, 1, \dots, L$ , with either a symmetric (for  $k$  even) or antisymmetric (for  $k$  odd) impulse response<sup>4</sup>.

When  $S_k(z)$  are linear-phase FIR filters, the Farrow structure is sometimes referred to as the modified Farrow structure, e.g., Vesma & Saramäki (1996), but we simply refer to it as the Farrow structure. The Farrow structure is efficient for interpolation whereas, for decimation, it is better to use the transposed Farrow structure, as discussed in Babic et al. (2002); Hentschel & Fettweis (2000).

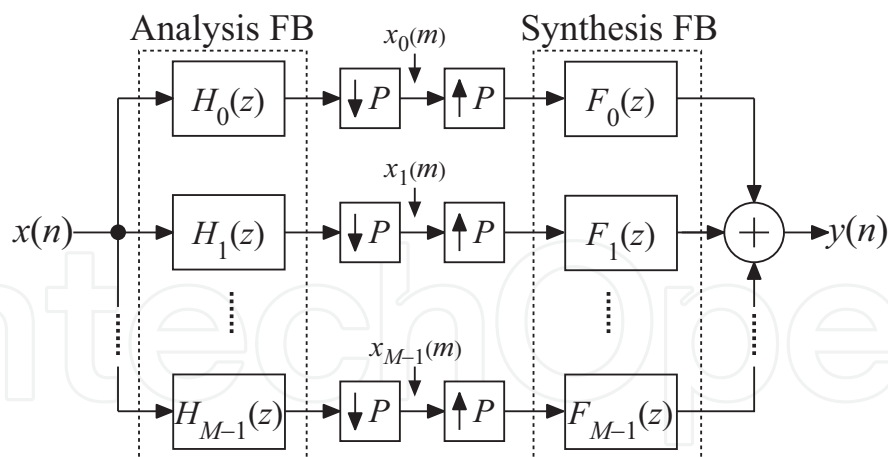
The transfer function of the Farrow structure is

$$H(z, \mu) = \sum_{k=0}^L S_k(z) \mu^k = \sum_{k=0}^L \sum_{n=0}^{N_k} s_k(n) z^{-n} \mu^k = \sum_{n=0}^N \sum_{k=0}^L s_k(n) \mu^k z^{-n} = \sum_{n=0}^N h(n, \mu) z^{-n}. \tag{7}$$

Here,  $|\mu| \leq 0.5$  and  $N$  is the order of the overall impulse response  $h(n, \mu)$ . Further,  $\mu$  is the FD value, i.e., the time difference between each input sample and its corresponding output sample. In the rest of the chapter, we will use  $h(n)$  and  $H(z)$  instead of  $h(n, \mu)$  and  $H(z, \mu)$ , respectively. If  $\mu$  is constant for all input samples, the Farrow structure delays a bandlimited signal by a fixed  $\mu$ . In general, SRC amounts to delaying every input sample with a different  $\mu$ . Thus, by controlling  $\mu$  for every input sample, the Farrow structure can perform SRC.

Generally,  $S_k(z)$  are designed so that  $H(z)$  approximates an allpass transfer function with FD of  $\mu$  over the frequency range of interest according to Babic et al. (2002); Johansson & Hermanowicz (2006); Johansson & Löwenborg (2003); Pun et al. (2003); Tseng (2002); Vesma & Saramäki (1996; 1997; 2000). The desired causal magnitude and unwrapped

<sup>4</sup> With infinite-length impulse response (IIR) filters, care must be taken to avoid transients as  $\mu$  changes for every sample.

Fig. 7. General  $M$ -channel FB.

phase responses are

$$H_{\text{des}}(e^{j\omega T}) = e^{-j(\Delta+\mu)\omega T}, \quad (8)$$

$$\Phi_{\text{des}}(\omega T) = -(\Delta + \mu)\omega T, \quad (9)$$

where

$$\Delta = \frac{\max_k(N_k)}{2}. \quad (10)$$

The main advantage of the Farrow structure is its ability to perform rational SRC using only one set of  $S_k(z)$  and by simple adjustments of  $\mu$ .

### 2.3 General $M$ -channel FBs

An  $M$ -Channel filter bank (FB), shown in Fig. 7, splits  $X(z)$  into the  $M$  subbands  $X_m(z)$ ,  $m = 0, 1, \dots, M-1$ , using the analysis filter bank (AFB) filters  $H_m(z)$ . To reconstruct  $X(z)$ , we need the synthesis filter bank (SFB) filters  $F_m(z)$ . Furthermore, upsamplers and downsamplers by  $P$  are also required as in Fig. 7. The output of a general  $M$ -channel FB is

$$Y(z) = \frac{1}{P} \sum_{n=0}^{P-1} X(zW_P^n) \sum_{m=0}^{M-1} H_m(zW_P^n) F_m(z). \quad (11)$$

Ideally,  $Y(z)$  is a scaled (by  $\alpha$ ) and delayed (by  $\beta$ ) version of  $X(z)$ , i.e.,  $y(n) = \alpha x(n - \beta)$ . Such a system is referred to as PR. In near perfect reconstruction (NPR) FBs,  $\alpha$  is frequency dependent and the distortion transfer function is

$$V_0(z) = \frac{1}{P} \sum_{m=0}^{M-1} H_m(z) F_m(z), \quad (12)$$

whereas the aliasing transfer functions are

$$V_l(z) = \frac{1}{P} \sum_{m=0}^{M-1} H_m(zW_P^l) F_m(z), \quad l = 1, 2, \dots, P-1. \quad (13)$$



These FBs are generally linear periodic time-varying systems with a period  $M$ . Without aliasing, we have a linear time-invariant (LTI) system as defined in Vaidyanathan (1993). In a PR FB,

$$V_0(e^{j\omega T}) = c, \quad c > 0 \quad (14)$$

$$V_l(e^{j\omega T}) = 0, \quad l = 1, 2, \dots, P-1. \quad (15)$$

If  $P = M$ , the FB is maximally decimated but  $P < M$  leads to oversampled FBs as in Vaidyanathan (1993). If  $V_0(z)$  is allpass (has linear-phase), we have zero amplitude (phase) distortion.

## 2.4 Modulated FBs

To obtain the AFB and SFB filters, one can modulate a single  $N$ th-order linear-phase FIR prototype filter  $G(z) = \sum_{n=0}^N g(n)z^{-n}$ . With cosine modulation, which is outlined in Chen & Chiueh (2008); Ihalainen et al. (2007); Saramäki & Bregović (2001a), we have

$$h_m(n) = 2g(n) \cos[(m+0.5)\frac{\pi}{M}(N-n+\frac{M+1}{2})], \quad (16)$$

$$f_m(n) = 2g(n) \cos[(m+0.5)\frac{\pi}{M}(n+\frac{M+1}{2})] = h_m(N-n). \quad (17)$$

In a PR cosine modulated filter bank (CMFB),  $N = 2KM - 1$  where  $K$  is an integer overlapping factor as defined in Viholainen et al. (2006). For complex modulated FBs,

$$h_m(n) = g(n)W_M^{-mn}, \quad (18)$$

$$f_m(n) = h_m(n). \quad (19)$$

In the maximally decimated case, we can use modified discrete Fourier transform (MDFT) FBs, as in Bregović & Saramäki (2005); Karp & Fliege (1999). An  $M$ -channel MDFT FB can be realized as either of the following, according to Fliege (1995),

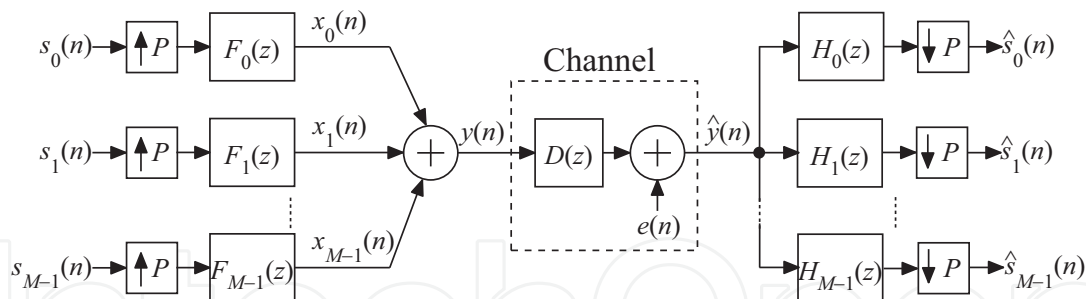
- Two SRC stages with ratios  $\frac{M}{2}$  and 2 while adding some phase offset between these stages.
- Two separate FBs where the phase offset is applied outside the AFB and SFB.

If an MDFT FB is PR,  $N$  is an integer as  $KM + s$  where  $0 \leq s < M$ . With the AFB and SFB filters, having uniform or nonuniform passbands, we have uniform or nonuniform FBs as in Vaidyanathan (1993). Nonuniform FBs can also be obtained by modulation, as discussed in Princen (1994), where

$$h_m(n) = a_m g_m(n) e^{-\frac{j\pi\alpha_m}{M_m}(n-\frac{L_m-1}{2})} + a_m^* g_m^*(n) e^{\frac{j\pi\alpha_m}{M_m}(n-\frac{L_m-1}{2})}, \quad (20)$$

$$f_m(n) = b_m g_m(n) e^{-\frac{j\pi\alpha_m}{M_m}(n-\frac{L_m-1}{2})} + a_m^* g_m^*(n) e^{\frac{j\pi\alpha_m}{M_m}(n-\frac{L_m-1}{2})}. \quad (21)$$

Here,  $\alpha_m = (K_m + 0.5)$  and  $g_m(n)$  is the (possibly complex) prototype filter of length  $L_m$  with  $M_m$  being the decimation factor in each branch. Each branch has a center frequency as  $\pm \frac{\pi\alpha_m}{M_m}$  with  $K_m$  being an integer where  $a_m$  and  $b_m$  define the modulation phase. The nonuniform FBs achieve a more general time and frequency tiling. The sine modulated filter bank (SMFB) is obtained similar to (16) and (17). The exponentially modulated filter bank (EMFB), with

Fig. 8. General  $M$ -channel TMUX.

complex filters, is a combination of SMFB and CMFB, as outlined in Chen & Chiueh (2008); Ihalainen et al. (2007).

#### 2.4.1 Filter design for modulated FBs

To design the prototype filter  $G(z)$ , we can use any standard filter design technique, e.g., Bregović & Saramäki (2005); Fliege (1995); Heller et al. (1999); Martin-Martin et al. (2008; 2005); Mirabbasi & Martin (2003); Mitra (2006); Saramäki & Bregović (2001a); Saramäki & Bregovic (2001b); Vaidyanathan (1993); Viholainen et al. (1999). The MDFT FB has a lowpass  $G(z)$  with a stopband edge at  $\omega_s T = \frac{2\pi}{M}$  according to Fliege (1995). The CMFB has a lowpass  $G(z)$  with a stopband edge at  $\omega_s T = \frac{\pi(1+\rho)}{2M}$  and a 3-dB cutoff frequency at  $\omega T = \frac{\pi}{2M}$  as discussed in Diniz et al. (2004); Martin-Martin et al. (2005). If  $0 < \rho \leq 1$ , only the adjacent branches overlap. With  $1 < \rho \leq 2$  (or  $\rho > 2$ ), two (or at least three) adjacent branches overlap. In both FBs,  $G(z)$  satisfies the power complementary property.

#### 2.5 General $M$ -channel TMUXs

A transmultiplexer (TMUX) converts the time multiplexed components of a signal into a frequency multiplexed version and back so that several users transmit and receive over a common channel, as noted in Vaidyanathan & Vrcelj (2004). A TMUX is also referred to as a FB transceiver, e.g., Beaulieu & Champagne (2009); Bianchi & Argenti (2007); Borna & Davidson (2007); Chiang et al. (2007); Lin & Phoong (2001).

Assume that we want to transmit a series of symbol streams  $s_k(n)$ ,  $k = 0, 1, \dots, M-1$ , through a channel. As in Fig. 8, we can pass  $s_k(n)$  through the transmitter filters  $F_k(z)$ . Then, (4) gives

$$x_k(n) = \sum_{m=-\infty}^{\infty} s_k(m) f_k(n - mP). \quad (22)$$

Here, the channel is described by a possibly complex LTI filter  $D(z) = \sum_{n=0}^{L_D} d(n)z^{-n}$  followed by an additive noise source  $e(n)$ . The receiver filters  $H_k(z)$  separate the signals and only a downsampling by  $P$  is needed to retrieve the original symbol streams. Ignoring the channel and for  $i = 0, 1, \dots, M-1$ , we have

$$\hat{S}_i(z) = \sum_{k=0}^{M-1} S_k(z) T_{ki}(z^P), \quad T_{ki}(z^P) = \frac{1}{P} \sum_{l=0}^{P-1} F_k(zW_P^l) H_i(zW_P^l). \quad (23)$$



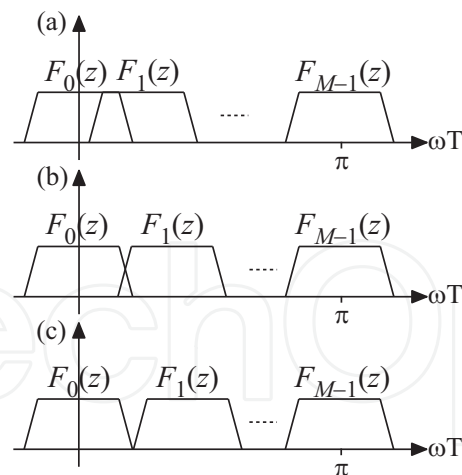


Fig. 9.  $M$ -channel TMUX filters. (a) Overlapping. (b) Marginally overlapping. (c) Non-overlapping.

Typical characteristics of  $F_k(z)$  and  $H_k(z)$  are shown in Fig. 9. Similar to FBs, TMUXs can be redundant ( $P > M$ ) or critically sampled ( $P = M$ ). To avoid inter-symbol interference (ISI), a level of redundancy may be needed such that  $P - M \geq L_D$ , according to Bianchi et al. (2005). The output of the TMUX in (23) is

$$\hat{S}_i(z) = T_{ii}(z)S_i(z) + \sum_{k=0, k \neq i}^{P-1} T_{ki}(z)S_k(z) \quad (24)$$

where  $T_{ii}(z)$  and  $T_{ki}(z)$  represent the ISI and the inter-carrier interference (ICI), respectively, as in Furtado et al. (2005). The ISI (ICI) is sometimes also referred to as interband (cross-band) ISI, e.g., Chiang et al. (2007).

If an LTI filter is placed between an upsampler and a downsampler of ratio  $P$ , the overall system is equivalent to the decimated (by  $P$ ) version of its impulse response, as mentioned in Vaidyanathan (1993). In this case, designing  $F_k(z)$  and  $H_k(z)$  so that the decimated (by  $P$ ) version of  $F_k(z)H_m(z)$  becomes a pure delay if  $k = m$  and zero otherwise, the TMUX becomes PR. In terms of (24), this means

$$T_{ii}(z) = \frac{1}{P} \sum_{l=0}^{P-1} F_i(z^{\frac{1}{P}} W_P^l) H_i(z^{\frac{1}{P}} W_P^l) = \alpha z^{-\beta}, \quad (25)$$

$$T_{ki}(z) = \frac{1}{P} \sum_{l=0}^{P-1} F_k(z^{\frac{1}{P}} W_P^l) H_i(z^{\frac{1}{P}} W_P^l) = 0. \quad (26)$$

In a PR system,  $\hat{s}_k(n) = \alpha s_k(n - \beta)$ . The PR properties can be satisfied for both critically sampled and redundant TMUXs. For the critically sampled case, there may not always exist FIR or stable IIR solutions. Therefore, some redundancy makes the solutions feasible, as in de Barcellos et al. (2006); Kovačević & Vetterli (1993); Li et al. (1997); Xie, Chan & Yuk (2005); Xie, Chen & Sho (2005), and it also simplifies the PR conditions. Duality of TMUXs and FBs allows one to obtain a TMUX from its corresponding FB, as noted in Vaidyanathan (1993).

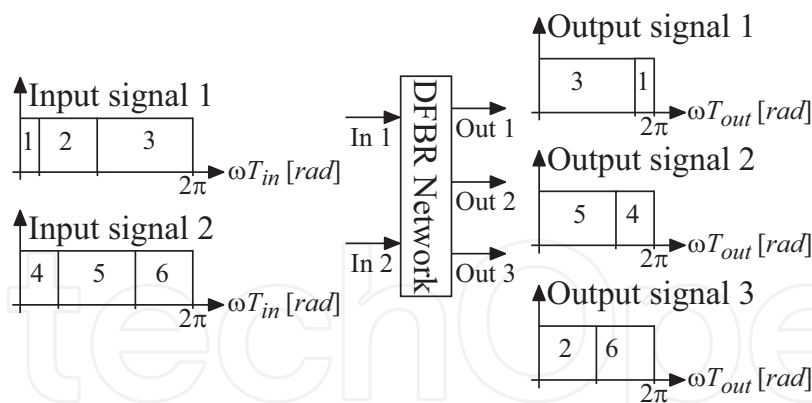


Fig. 10. Approach I: DFBR networks process composite signals to reallocate users from one composite input signal to another composite output signal.

This duality applies to both critically sampled and redundant systems. If a FB is free from aliasing, the corresponding TMUX is free from ICI, according to Fliege (1995).

### 3. Approach I: use of DFBR networks

For DFBR, we assume that signals from several users, e.g., mobile handsets in a cellular network or computers in a wireless local area network (WLAN), have been added into a composite signal at a main station, e.g., a base station in a cellular network or an access point in a WLAN. This main station then finds available frequency slots and reallocates each user to one of them. Such a main station is similar to a bentpipe satellite payload, as outlined in Nguyen et al. (2002), with its idea of operation shown in Fig. 10. The composite signals are processed by the DFBR network and the users are reallocated to new frequency slots. These slots could be different antenna beams of a satellite payload or different cells in a cellular network. Multiple antennas of a satellite payload perform signal filtering in spatial rather than frequency domain. This is similar to the techniques utilizing multiple antennas for cognitive radios which are discussed in Cabric & Brodersen (2005). The DFBR networks could also be useful for the centralized cooperative cognitive radios, as in Ganesan & Li (2005), and they can also be considered as secondary base stations in licensed band cognitive radios. In licensed band networks, the DFBR can coexist with the primary networks so as to opportunistically operate in an overlay transmission.

The DFBR network can be a multi-input multi-output system as it can have a number of composite input and output signals. The dynamic nature of the DFBR networks allows the users to occupy any suitable<sup>5</sup> frequency slot in a time-varying manner. Each user can be sent in contiguous or separate frequency bands leading to contiguous or fragmented DFBR which is outlined in Leaves et al. (2004). The separate frequency bands can be considered as a multi-spectrum transmission. Specifically, as white spaces are mostly fragmented, according to Yuan, Bahl, Chandra, Chou, Ferrell, Moscibroda, Narlanka & Wu (2007), the user signals can be transmitted in several non-contiguous frequency bands.

<sup>5</sup> The frequency slot depends on spatial and temporal parameters, e.g., the number of available slots, user movement, and primary user activity, etc. Akyildiz et al. (2006) but the DFBR network is independent of these parameters.

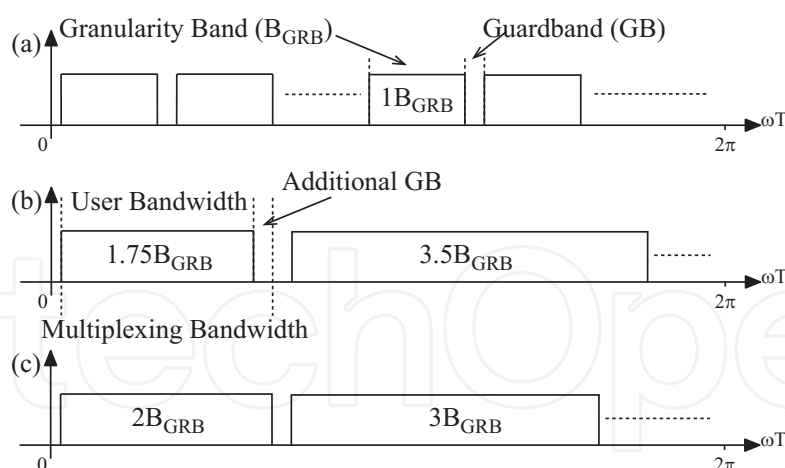


Fig. 11. User bandwidth versus multiplexing bandwidth. The multiplexing bandwidth is an integer multiple of the GRB and it contains a user bandwidth as well as some additional GB.

### 3.1 Structure of the DFBR network

This chapter uses the term DFBR which is essentially the same as the flexible frequency-band reallocation (FFBR) in, e.g., Abdulazim & Göckler (2006); Eghbali et al. (2009a); Johansson & Löwenborg (2007); Rosenbaum et al. (2006). For the illustrations, we will use the FFBR network, in Johansson & Löwenborg (2007), but one can use any other FFBR network as well.

### 3.2 User bandwidth versus multiplexing bandwidth

The DFBR networks divide the user signals into a number of granularity bands (GRBs) on which the frequency shifts are performed. As the DFBR networks utilize FBs, the multiplexing bandwidth must be an integer multiple of the granularity band (GRB). The DFBR networks perform frequency shifts on users whose bandwidths are, in general, rational multiples of the GRB. An important issue is to ensure that the users do not share a GRB. This can be achieved by allowing some additional guardband (GB). However, the additional GB affects the spectrum efficiency resulting in a trade-off. As in Fig. 11, a multiplexing bandwidth contains a user bandwidth and some additional GB.

### 3.3 Reconfigurability

A cognitive radio should adjust its operating parameters without hardware modifications as discussed in Jondral (2005). It is built on the platform for a software defined radio where the processing is mainly in the digital domain, according to Zhao & Sadler (2007). There are several reconfigurable parameters such as operating frequency, modulation method, transmission power, and communication standard etc. For adaptable operating frequency, or flexible frequency carrier tuning, as in Leaves et al. (2004), a cognitive radio should change its operating frequency without restricting the system throughput and hardware.

The DFBR networks can perform any frequency shift of any user having any bandwidth, using a channel switch. This switch seamlessly directs different FB channels to their desired outputs without any arithmetic complexity. In addition, the system parameters are determined and fixed only once, offline. The reconfigurable operation is then performed by reconfiguring

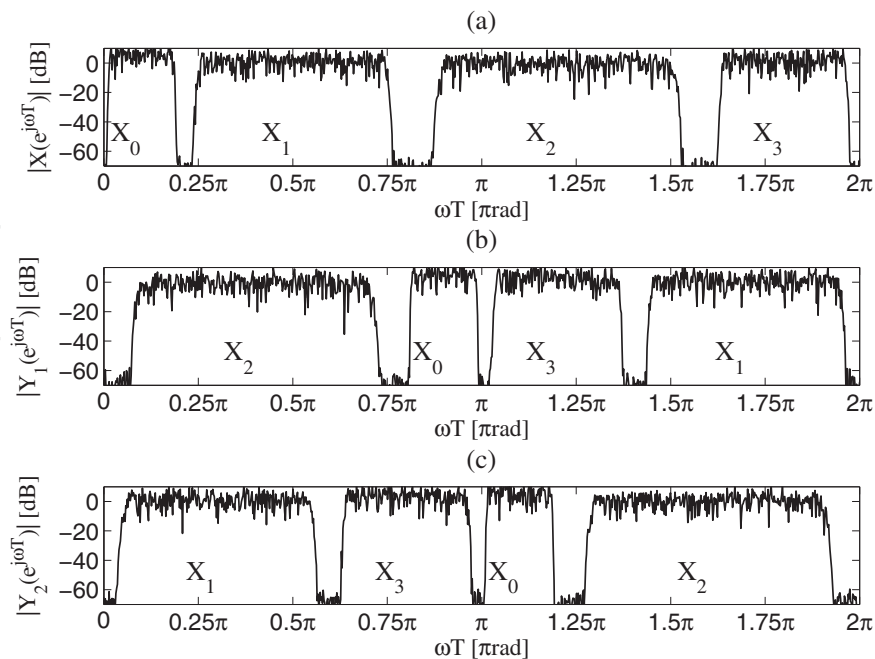


Fig. 12. Input and the reallocated outputs using the channel switch configurations in Figs. 14(a) and 14(b).

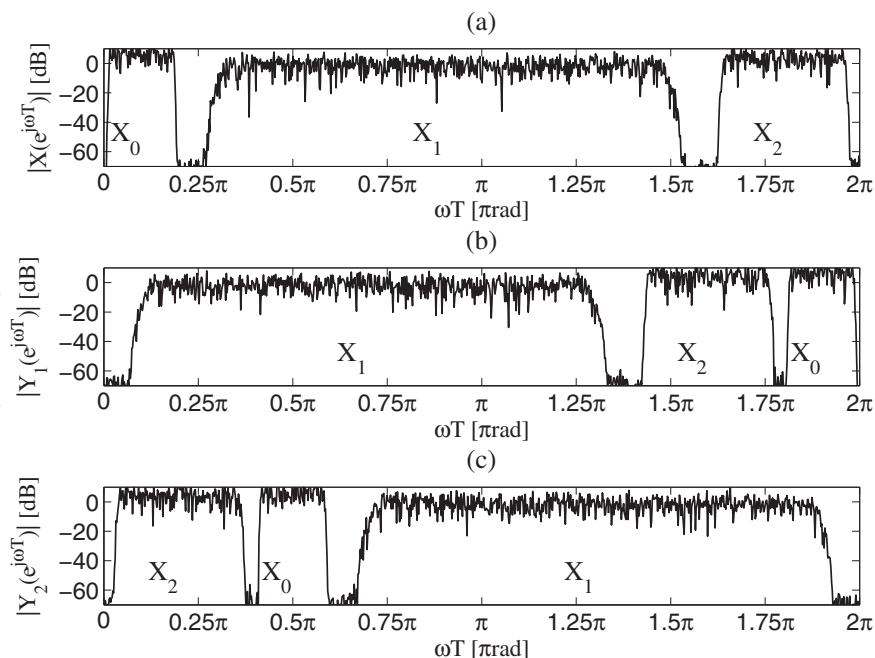


Fig. 13. Input and the reallocated outputs using the channel switch configurations in Figs. 14(c) and 14(d).

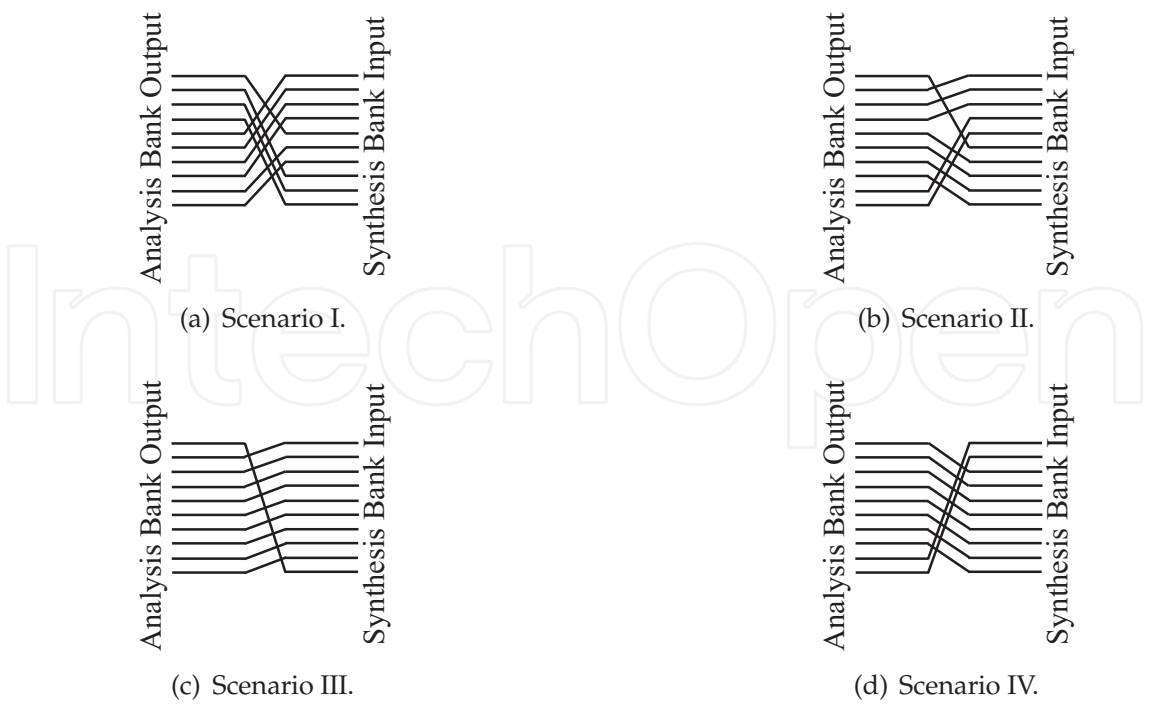


Fig. 14

the channel switch, online. Here, the user bandwidths are predetermined but they can be arbitrary. The DFBR network makes a hand off by changing the operating frequency.

Figures 12 and 13 show two cases where, respectively, four and three users have occupied the whole frequency band. To generate these user signals, the multimode TMUX of Eghbali et al. (2008b) has been used. In Fig. 12(a), the user signals  $\{X_0, X_1, X_2, X_3\}$  occupy, respectively, user bandwidths of  $\{1, 2.9, 3.6, 1.9\}$  GRBs. Each GRB has a spectral width of  $\frac{2\pi}{Q} - \frac{2\epsilon\pi}{Q}$  with  $Q = 10$  and  $\epsilon = 0.125$ . In Fig. 13(a), the user signals  $\{X_0, X_1, X_2\}$  occupy  $\{1, 6.9, 1.9\}$  GRBs, respectively. To ensure that the users do not share a GRB, one can add some additional GB. This difference in the amount of the GB, between different users, can be recognized from Figs. 12 and 13.

These examples assume the DFBR network to operate on the same antenna beam. By having several DFBR networks, the users can be reallocated between different antenna beams according to Johansson & Löwenborg (2007). This requires a duplication of DFBR networks and a channel switch which directs the user signals between different DFBR networks. Each branch of the channel switches, in Figs. 14(a)–14(d), represents the operation of two FB channels as each GRB contains two FB channels. Specifically, the values of  $N$ ,  $M$ , and  $L$ , in Johansson & Löwenborg (2007), are 20, 10, and 2, respectively, for the examples above.

3.4 Modifications

The use of DFBR networks in cognitive radios needs some modifications, mainly in the system parameters. For different system parameters, the implementation complexity may be different but once the parameters are chosen, the implementation complexity remains constant and the system can be easily reconfigured on the same hardware platform. For the DFBR networks, the width of a GRB must be proportional to that of the spectrum holes. Thus, one requires to

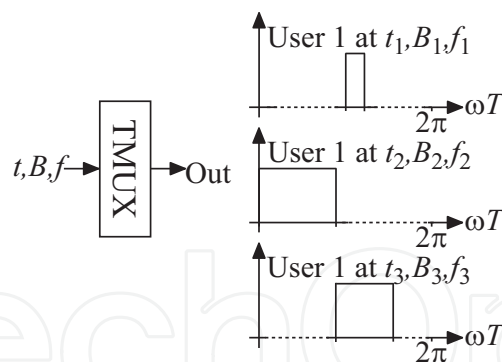


Fig. 15. Approach II: TMUXs to perform DFBA. At any time  $t_k$ , each user decides its bandwidth  $B_k$  and operating frequency  $f_k$ .

choose a value for the  $B_{GRB} = \frac{2\pi(1-\epsilon)}{Q}$ , in Fig. 11, so as to represent any spectrum hole as a rational multiple of  $B_{GRB}$ .

#### 4. Approach II: use of TMUXs

Using TMUXs, each user terminal can adjust its operating frequency and bandwidth. The basic idea is depicted in Fig. 15 where different bandwidths and center frequencies can be generated using multirate signal processing techniques. These TMUXs can also be regarded as the time-spectrum blocks, discussed in Yuan, Bahl, Chandra, Moscibroda, Narlanka & Wu (2007), which can transmit any amount of data at any time interval and on any portion of the frequency spectrum. This applies if the licensed users choose frequency division multiple access and/or time division multiple access as their spectrum access mode. Then, the spectrum holes are identified in the time/frequency plane, as outlined in Jondral (2007). As shown in Fig. 16, the interpolation part represents the transmitter with a variable filter placing the desired user signal at the desired center frequency. The receiver, i.e., the decimation part, retrieves the input signal.

Similar to straightforward DFBR solutions, one can use conventional nonuniform TMUXs to place users, with different bandwidths, at different center frequencies. This becomes inefficient when simultaneously considering the increased number of communication scenarios and the desire to support dynamic communications. In this context, TMUX structures of the general form shown in Fig. 17 can be used. In the SFB, the system  $C_p$  performs interpolation by a rational ratio  $R_p$  whereas the system  $\hat{C}_p$  in the AFB performs decimation by a rational ratio  $R_p$ .

##### 4.1 Structure of the TMUX

Any of the TMUXs, in the references Eghbali et al. (2007a; 2008a;b; 2009b; 2010); Eghbali, Johansson & Löwenborg (2011a;b), can be used here. The TMUX, in Eghbali et al. (2010); Eghbali, Johansson & Löwenborg (2011b), has a rather different structure. Instead of variable lowpass filters and frequency shifters as, in Eghbali et al. (2007a; 2008a;b; 2009b); Eghbali, Johansson & Löwenborg (2011a), it performs bandpass rational SRC using flexible commutators and fixed bandpass filters. However, one can generally describe it in terms of Fig. 17.



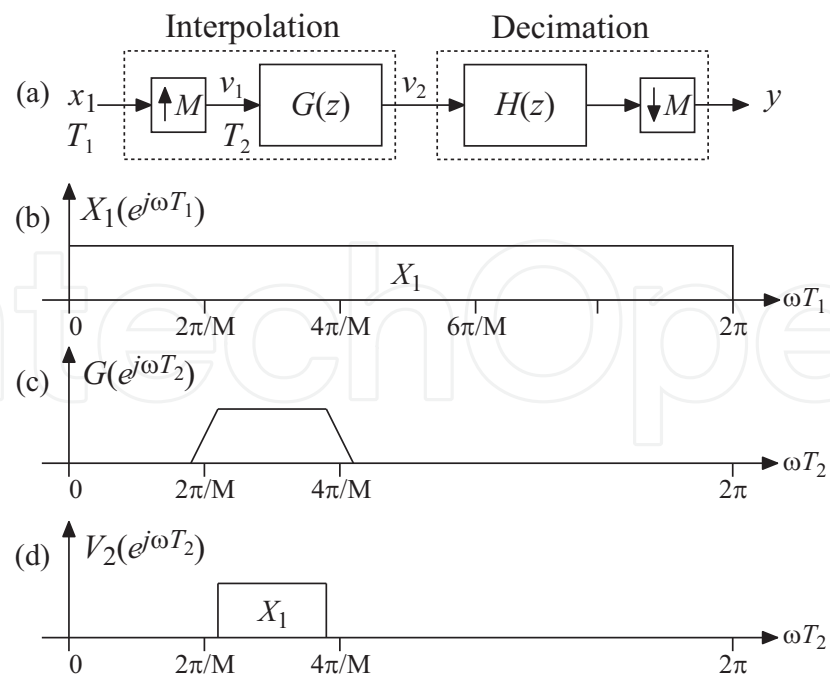


Fig. 16. Principle of TMUXs using multirate building blocks. The interpolation (decimation) part represents the transmitter (receiver). Variable filters place the desired user signal at the desired center frequency.

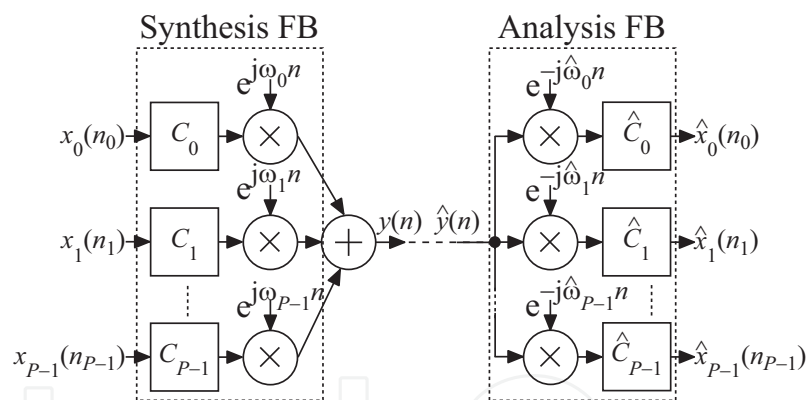


Fig. 17. General structure of a multimode TMUX where systems  $C_p$  and  $\hat{C}_p$  perform rational SRC.

4.2 Reconfigurability

A cognitive radio must adjust its operating frequency and bandwidth without hardware modifications. The DFBR networks partially provide this capability but they have no control over the user bandwidth. In contrast, the TMUX-based approaches add reconfigurability to the user bandwidth as well. Furthermore, they bring flexible receiver signal filtering, outlined in Leaves et al. (2004), by changing the transmitter and receiver filters. As can be seen from Figs. 12 and 13, the TMUX allows different numbers of users, e.g., four and three, with different user bandwidths to occupy the whole frequency band. These TMUXs provide this full reconfigurability without any hardware changes.

### 4.3 Modifications

Similar to the DFBR networks, we should have certain system parameters to eliminate the need for hardware changes while having simple reconfigurability. Regarding DFBA, there are different ways to perform SRC which could be useful in different scenarios. The TMUX, in Eghbali et al. (2007a; 2008b), generates a GRB through integer interpolation by, e.g.,  $W$ , thereby resulting in  $B_{GRB} = \frac{2\pi(1+\rho)}{W}$  where  $\rho$  is the roll-off. Then, rational  $R_p$  multiples of  $B_{GRB}$  can be created using the Farrow structure.

The TMUX, in Eghbali et al. (2008a; 2009b); Eghbali, Johansson & Löwenborg (2011a), assumes no GRBs and it allows the users to occupy any portion of the spectrum. It utilizes the Farrow structure to perform general rational SRC by, e.g.,  $R_p = \frac{A_p}{B_p}$ . Here, one can also assume a GRB of size  $B_{GRB} = \frac{2\pi(1+\rho)}{A_p}$ . Then, users can have bandwidths which are integer  $B_p$  multiples of  $B_{GRB}$ .

Although references Eghbali et al. (2010); Eghbali, Johansson & Löwenborg (2011b) propose a slightly different TMUX, one can also assume  $B_{GRB} = \frac{2\pi(1+\rho)}{M}$ . Then, users have bandwidths which are integer  $M_p$  multiples of  $B_{GRB}$ . This applies to the case with MDFT FBs but for a CMFB, similar formulae can be derived.

### 5. Choice of frequency shifters

To perform a hand off without information loss, the DFBR network requires the users not to share a GRB. Consequently, a lossless reallocation requires to (i) generate appropriate frequency division multiplexed (FDM) input patterns, and (ii) determine proper parameters for the DFBR networks. To generate the input patterns, the reconfigurability of the TMUXs in Fig. 17 can be used. After generating the user signals with desired bandwidths, the frequency shifters  $\omega_p$ ,  $p = 0, 1, \dots, P-1$ , can be computed to allow some additional GB. Here, an example using the TMUX in Eghbali et al. (2008b) is provided. Assuming some bandwidths which are rational, e.g.,  $R_p$ , multiples of  $B_{GRB}$ , the subcarrier  $\omega_p$  for user  $p$  is

$$\omega_p = \begin{cases} \frac{F_0}{2} & \text{if } p = 0 \\ \sum_{k=0}^{p-1} F_k + \frac{F_p}{2} & \text{if } p \neq 0. \end{cases} \quad (27)$$

where  $F_p = \lceil R_p \rceil \frac{2\pi}{Q}$ ,  $p = 0, 1, \dots, k$ , with  $\lceil \cdot \rceil$  being the ceiling operation. Here,  $F_p$  is the multiplexing bandwidth and the ceiling operation ensures that the users do not share a GRB. This formulation applies to the case where DFBA and DFBR are simultaneously used. Otherwise, similar formulae can be used but one may anyhow require some additional GB due to the design margins.

In Figs. 12 and 13, the users occupy  $R_p = \{2.9, 3.6, 1.9, 6.9\}$  GRBs. This necessitates an additional GB which is  $E_p = \{0.1, 0.4, 0.1, 0.1\}$  multiples of  $\frac{2\pi}{Q}$ . Therefore, the spectrum efficiency decreases. For a set of values  $R_p$ ,  $p = 0, 1, \dots, P-1$ , about

$$\eta_{dec} = \frac{\frac{2\pi}{Q} \sum_{p=0}^{P-1} (\lceil R_p \rceil - R_p)}{2\pi} = \frac{\sum_{p=0}^{P-1} \lceil R_p \rceil - R_p}{Q} \quad (28)$$

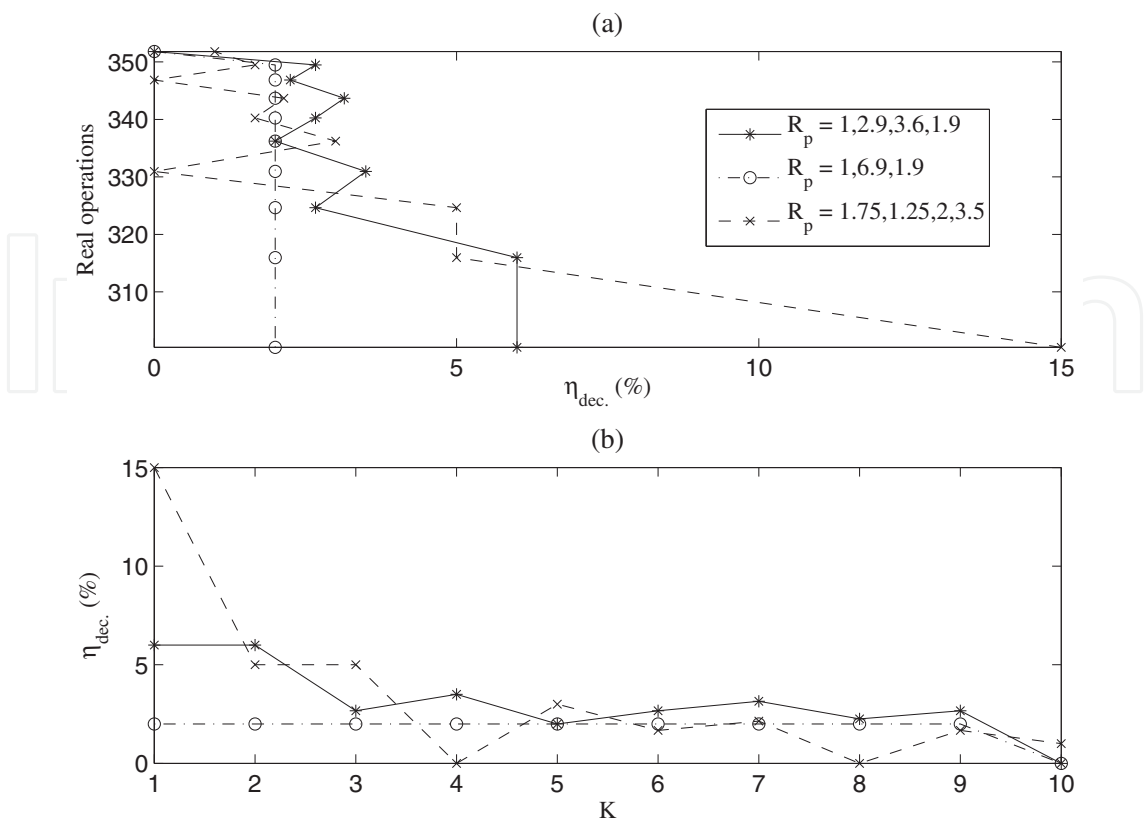


Fig. 18. Trade-off between spectrum efficiency and arithmetic complexity. (a) Decrease in spectrum efficiency versus per-sample arithmetic complexity. (b) Trend of spectrum efficiency versus different  $K$  in (29).

percent of the spectrum in  $[0, 2\pi]$  is not used. In the examples of Figs. 12 and 13, about 6% and 2% of the total spectrum is not used due to the additional GB. To decrease  $\eta_{dec}$ , one can increase  $Q$  by, e.g.,  $K$  times, so that (28) becomes

$$\eta_{dec} = \frac{\sum_{p=0}^{P-1} \lceil KR_p \rceil - KR_p}{KQ}. \tag{29}$$

However, increasing  $Q$  would increase the order of the prototype filter of the corresponding FB. For each  $K$ , the prototype filter of the DFBR network would have a transition band of  $\frac{2\pi\epsilon}{KQ}$  according to Eghbali et al. (2009a); Johansson & Löwenborg (2007). As the order of a linear-phase FIR filter is inversely proportional to the width of its transition band, there is a trade-off between the spectrum efficiency and the arithmetic complexity.

With a  $K$ -fold increase in  $Q$ , the length of the prototype filter and the number of FB channels increase proportional to  $K$ . Figure 18 shows the trend in spectrum efficiency with respect to the per-sample arithmetic complexity of the DFBR network in Johansson & Löwenborg (2007). Here, the examples of Figs. 12 and 13 as well as that of Eghbali et al. (2008b) with  $R_p = \{1.75, 1.25, 2, 3.5\}$  are considered. As can be seen, a larger  $K$  increases the per-sample arithmetic complexity but it decreases  $\eta_{dec}$ . The values of  $R_p$  mainly determine the maximum and minimum amounts of  $\eta_{dec}$ . Hence, for every set of  $R_p$ , one can determine a  $K$  such that  $\eta_{dec}$  and the per-sample arithmetic complexity are within the acceptable ranges.

## 6. Conclusion

This chapter discussed two approaches for the baseband processing in cognitive radios based on DFBR and DFBA. They can support different bandwidths and center frequencies for a large set of users while being easily reconfigurable.

In DFBR networks, composite FDM signals are processed and the users are reallocated to new center frequencies. They are applicable to cognitive radios with multiple antennas, centralized cooperative cognitive radios, and secondary base stations in licensed band cognitive radios. In DFBA networks, each user controls its operating frequency and bandwidth. These networks can be regarded as the time-spectrum blocks.

The reconfigurability of DFBA and DFBR is performed either by a channel switch, in DFBR, or by variable multipliers/commutators, in DFBA. The examples in Figs. 12 and 13 show the increased flexibility to allocate and reallocate any user to any center frequency by simultaneous utilization of DFBA and DFBR. In this case, the individual users can occupy any available frequency slot and be reallocated by the base station.

Basically, utilizing any of Approaches I and II only requires modifications imposed by the special choice of the system parameters. After choosing these parameters once, we must design the filters to satisfy any desired level of error. Then, the same hardware can be reconfigured in a simple manner.

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The fast user growth in wireless communications has created significant demands for new wireless services in both the licensed and unlicensed frequency spectra. Since many spectra are not fully utilized most of the time, cognitive radio, as a form of spectrum reuse, can be an effective means to significantly boost communications resources. Since its introduction in late last century, cognitive radio has attracted wide attention from academics to industry. Despite the efforts from the research community, there are still many issues of applying it in practice. This book is an attempt to cover some of the open issues across the area and introduce some insight to many of the problems. It contains thirteen chapters written by experts across the globe covering topics including spectrum sensing fundamental, cooperative sensing, spectrum management, and interaction among users.

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