

# OFDM and Its Wireless Applications: A Survey

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**Abstract**—Orthogonal frequency-division multiplexing (OFDM) effectively mitigates intersymbol interference (ISI) caused by the delay spread of wireless channels. Therefore, it has been used in many wireless systems and adopted by various standards. In this paper, we present a comprehensive survey on OFDM for wireless communications. We address basic OFDM and related modulations, as well as techniques to improve the performance of OFDM for wireless communications, including channel estimation and signal detection, time- and frequency-offset estimation and correction, peak-to-average power ratio reduction, and multiple-input-multiple-output (MIMO) techniques. We also describe the applications of OFDM in current systems and standards.

**Index Terms**—Channel estimation, frequency-offset estimation, intercarrier interference (ICI), multicarrier (MC), multiple-input-multiple-output (MIMO) orthogonal frequency-division multiplexing (OFDM), peak-to-average power reduction, time-offset estimation, wireless standards.

## I. INTRODUCTION

HIGH-DATA-RATE transmission over mobile or wireless channels is required by many applications. However, the symbol duration reduces with the increase of the data rate, and dispersive fading of the wireless channels will cause more severe intersymbol interference (ISI) if single-carrier modulation, such as in time-division multiple access (TDMA) or Global System for Mobile Communications (GSM), is still used. From [1], to reduce the effect of ISI, the symbol duration must be much larger than the delay spread of wireless channels. In orthogonal frequency-division multiplexing (OFDM) [2]–[4], the entire channel is divided into many narrow-band subchannels,<sup>1</sup> which are transmitted in parallel to maintain high-data-

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<sup>1</sup>It is also called the subcarrier or tone in the literature.

rate transmission and, at the same time, to increase the symbol duration to combat ISI.

OFDM is a special form of multicarrier (MC) that dates back to 1960s. The concept of MC transmission was first explicitly proposed by Chang [5] in 1966. A detailed description of MC can also be found in [6] and [7]. Before Chang, Doelz *et al.* [8] had implemented a special MC system for a single-sideband voice channel in 1957, and Holsinger [9] had implicitly introduced the MC system in his dissertation at the Massachusetts Institute of Technology in 1964. In 1971, Weinstein and Ebert [3] proposed time-limited MC transmission, which is what we call OFDM today. The implementation of MC systems with equalization was investigated by Hirosaki *et al.* [10] and [11] and Peled and Ruiz [12]. Zimmerman and Kirsch [13] published one of the earliest papers in the application of MC in HF radio in 1967. More materials on the HF application of MC can be found in [14] and the references therein. The capacity of OFDM was investigated in [15] and [16]. In 1985, Cimini first applied OFDM in mobile wireless communications [2]. In [17], Casas and Leung discussed the application of MC over mobile radio FM channels. Bingham [18] studied the performance and complexity of MC modulation and concluded that the time for MC has come. The application of original OFDM, clustered OFDM, and MC code-division multiple access (CDMA) in mobile wireless systems can be found in [19]–[26].

The flexibility of OFDM provides opportunities to use advanced techniques, such as adaptive loading, transmit diversity, and receiver diversity, to improve transmission efficiency. Shannon's classical paper in 1948 [27] suggested that the highest data rate can be achieved for frequency-selective channels by using an MC system with an *infinitely dense* set of subchannels and adapting transmission powers and data rates according to the signal-to-noise ratio (SNR) at different subchannels. Based on his theory, a *water-filling* principle has been derived [28]. Cioffi and his group have extensively investigated OFDM with performance optimization for asymmetric digital subscriber line, which they more often called discrete multiple tone (DMT). Some of their earlier inventions on practical loading algorithms for OFDM or DMT systems were in [29]. More results on this topic can be found in [30]–[32].

The capacity of a wireless system can significantly be improved if multiple transmit and receive antennas are used to form multiple-input-multiple-output (MIMO) channels [33]–[37]. It is proved in [33] that, compared with a single-input-single-output (SISO) system, a MIMO system can improve the capacity by a factor of the minimum number of

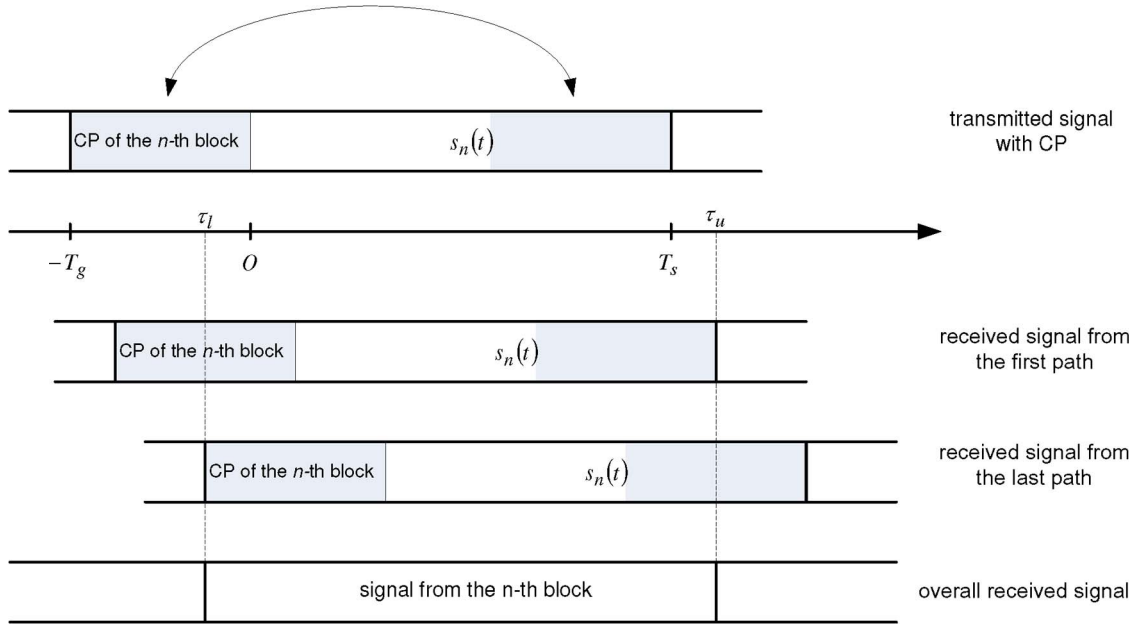


Fig. 1. Function of the CP.

transmit and receive antennas for flat fading or narrow-band channels. For wideband transmission, it is natural to combine OFDM with space-time coding (STC) or spatial-temporal processing to deal with frequency selectivity of wireless channels and to obtain diversity and/or capacity gains. Therefore, MIMO-OFDM has widely been used in various wireless systems and standards.

In this paper, we present a comprehensive survey of OFDM for wireless communications. We start with basic OFDM systems and techniques to improve the system performance in Section II. Then, we address various OFDM-related modulations in Section III. Next, we briefly introduce MIMO techniques for OFDM and its applications in current systems and standards in Sections IV and V, respectively. Finally, we present our concluding remarks in Section VI.

## II. BASIC OFDM

In this section, we will first describe basic OFDM and then introduce techniques to improve the performance of OFDM for wireless communications, including channel estimation, time- and frequency-varying impairment mitigation, and peak-to-average power ratio (PAPR) reduction techniques.

### A. OFDM

Let  $\{s_{n,k}\}_{k=0}^{N-1}$  with  $E|s_{n,k}|^2 = \sigma_s^2$  be the complex symbols to be transmitted at the  $n$ th OFDM block, then the OFDM-modulated signal can be represented by

$$s_n(t) = \sum_{k=0}^{N-1} s_{n,k} e^{j2\pi k \Delta f t}, \quad 0 \leq t \leq T_s \quad (1)$$

where  $T_s$ ,  $\Delta f$ , and  $N$  are the symbol duration, the subchannel space, and the number of subchannels of OFDM signals, respectively. For the receiver to demodulate the OFDM signal, the

symbol duration should be long enough such that  $T_s \Delta f = 1$ , which is also called the orthogonal condition since it makes  $e^{-j2\pi k \Delta f t}$  orthogonal to each other for different  $k$ . With the orthogonal condition, the transmitted symbols  $s_{n,k}$  can be detected at the receiver by

$$s_{n,k} = \frac{1}{T_s} \int_0^{T_s} s_n(t) e^{-j2\pi k \Delta f t} dt \quad (2)$$

if there is no channel distortion.

The sampled version of the baseband OFDM signal  $s(t)$  in (1) can be expressed as

$$s_n \left( m \frac{T_s}{N} \right) = \sum_{k=0}^{N-1} s_{n,k} e^{j2\pi k \Delta f m \frac{T_s}{N}} = \sum_{k=0}^{N-1} s_{n,k} e^{j \frac{2\pi m k}{N}} \quad (3)$$

which is actually the inverse discrete Fourier transform (IDFT) of the transmitted symbols  $\{s_{n,k}\}_{k=0}^{N-1}$  and can efficiently be calculated by fast Fourier transform (FFT). It can easily be seen that demodulation at the receiver can be performed using DFT instead of the integral in (2).

A cyclic prefix (CP) or guard interval is critical for OFDM to avoid interblock interference (IBI) caused by the delay spread of wireless channels. They are usually inserted between adjacent OFDM blocks. Fig. 1 shows the function of the CP. Without the CP, the length of the OFDM symbol is  $T_s$ , as shown in (1). With the CP, the transmitted signal is extended to  $T = T_g + T_s$  and can be expressed as

$$\tilde{s}_n(t) = \sum_{k=0}^{N-1} s_{n,k} e^{j2\pi k \Delta f t}, \quad -T_g \leq t \leq T_s.$$

It is obvious that  $\tilde{s}_n(t) = s_n(t + T_s)$  for  $-T_g \leq t \leq 0$ , which is why it is called the CP.

The impulse response of a wireless channel can be expressed by [38]

$$h(t) = \sum_i \gamma_i \delta(t - \tau_i) \quad (4)$$

where  $\tau_i$  and  $\gamma_i$  are the delay and the complex amplitude<sup>2</sup> of the  $i$ th path, respectively. Then, the received signal can be expressed as

$$x_n(t) = \sum_i \gamma_i \bar{s}_n(t - \tau_i) + n(t)$$

where  $n(t)$  represents the additive white Gaussian noise (AWGN) at the receiver. As demonstrated in Fig. 1,  $x_n(t)$  consists of only the signal component from the  $n$ th OFDM block when  $\tau_l \leq t \leq \tau_u$ , where  $\tau_l = -T_g + \tau_M$ ,  $\tau_u = T_s + \tau_m$ ,  $\tau_m = \min_i \{\tau_i\}$ , and  $\tau_M = \max_i \{\tau_i\}$ ; otherwise, the received signal consists of signals from different OFDM blocks.

If  $\tau_l \leq 0$  and  $\tau_u \geq T_s$ , then

$$\begin{aligned} x_{n,k} &= \frac{1}{T_s} \int_0^{T_s} x_n(t) e^{-j2\pi f_k t} dt \\ &= \frac{1}{T_s} \int_0^{T_s} \left\{ \sum_i \gamma_i \bar{s}_n(t - \tau_i) + n(t) \right\} e^{-j2\pi f_k t} dt \\ &= H_k s_{n,k} + n_k \end{aligned} \quad (5)$$

for  $0 \leq k \leq N - 1$  and all  $n$ , where  $H_k$  denotes the frequency response of the wireless channel at the  $k$ th subchannel and is defined as

$$H_k = \sum_i \gamma_i e^{-j2\pi k \Delta f \tau_i}$$

and  $n_k$  is the impact of AWGN and is defined as

$$n_k = \frac{1}{T_s} \int_0^{T_s} n(t) e^{-j2\pi f_k t} dt.$$

It can be proved that  $n_k$  are *independent identically distributed* complex circular Gaussian with zero mean and variance  $\sigma_n^2$ . With  $H_k$ , transmitted symbols can be estimated. For single-carrier systems, the received signal is the convolution of the transmitted sequences or symbols and the impulse response of wireless channels in addition to AWGN, whereas the impact of the channel is only a multiplicative distortion at each subchannel for OFDM systems, which makes signal detection in OFDM systems very simple and is also one of the reasons why OFDM is very popular nowadays.

As we have seen from the previous discussion, the CP or guard interval effectively avoids IBI. More comments on the CP and guard interval are listed here.

- 1) If a guard interval is used instead of the CP, that is, transmitting no signal in the place of the CP in Fig. 1, the

IBI and intercarrier interference (ICI) can also be avoided by the following simple processing:

$$\tilde{x}_n(t) = \begin{cases} x_n(t + T_s), & \text{if } 0 \leq t \leq \tau_m \\ x_n(t) + x_n(t + T_s), & \text{if } \tau_m < t < \tau_M \\ x_n(t), & \text{if } \tau_M \leq t \leq T_s. \end{cases}$$

It can easily be seen that if  $x_n(t)$  in (5) is substituted by  $\tilde{x}_n(t)$ , then the demodulated signal can also be expressed as

$$\tilde{x}_{n,k} = \frac{1}{T_s} \int_0^{T_s} \tilde{x}_n(t) e^{-j2\pi f_k t} dt = H_k s_{n,k} + \tilde{n}_k \quad (6)$$

if  $\tau_l \leq 0$  and  $\tau_u \geq T_s$ . However, the power of the demodulated noise  $\tilde{n}_k$  is a little bit higher than that in (5) since the noise power in  $\tilde{x}_n(t)$  for  $t \in (\tau_m, \tau_M)$  is doubled.

- 2)  $\tau_l \leq 0$  and  $\tau_u \geq T_s$  are not the necessary conditions for (5) or (6) to hold. By selecting the proper starting time of the integral, the condition can be relaxed into  $\tau_u - \tau_l \geq T_s$  or, equivalently,  $\tau_M - \tau_m \leq T_g$ . The time difference between the first and last paths of a wireless channel  $\tau_d = \tau_M - \tau_m$  is also called the delay span, which is different from the frequently used delay spread.<sup>3</sup>
- 3) As indicated in [4], if there is no CP or guard interval, or its length is not long enough, the delay spread of wireless channels will cause both IBI and ICI.

To further improve the performance of OFDM systems, link adaptation, including adapting transmission power or data rate of each subchannel, has extensively been investigated [29]–[32]. Link adaptation exploits the frequency-selectivity nature of wideband channels. It is performed in a subchannel-by-subchannel basis and may sometimes result in a signaling overhead. A historical review on the topic has been addressed in [39] and the references therein.

From the aforementioned discussion, channel state information (CSI) is required for coherent detection. Various channel estimation approaches have been developed for OFDM communications [40]. In Section II-B, we will discuss the different channel estimation approaches.

Similar to single-carrier modulation, time- and frequency-varying wireless channels affect the performance of OFDM systems. Time-varying impairments may come from the carrier frequency offset (CFO) caused by the mismatch of frequencies between the oscillators at the transmitter and the receiver, or from the Doppler spread due to the relative movement between the transmitter and the receiver. Frequency-varying impairments are from the timing offset or the delay spread of wireless channels. Since the duration of the OFDM symbol is longer and its bandwidth is narrower compared with single-carrier modulation, OFDM systems are more robust to frequency-selective fading but more sensitive to the time-varying impairment of

<sup>2</sup>When Doppler spread exists,  $\gamma_i$  changes with time and will be a narrow-band stochastic process.

<sup>3</sup>The delay spread of a wireless channel is defined as  $\tau_{\text{spread}} = \sqrt{\sum_i \sigma_i^2 (\tau_i - \bar{\tau})^2 / \sum_i \sigma_i^2}$ , where  $\sigma_i^2 = E|\gamma_i|^2$ , and  $\bar{\tau} = \sum_i \sigma_i^2 \tau_i / \sum_i \sigma_i^2$ .

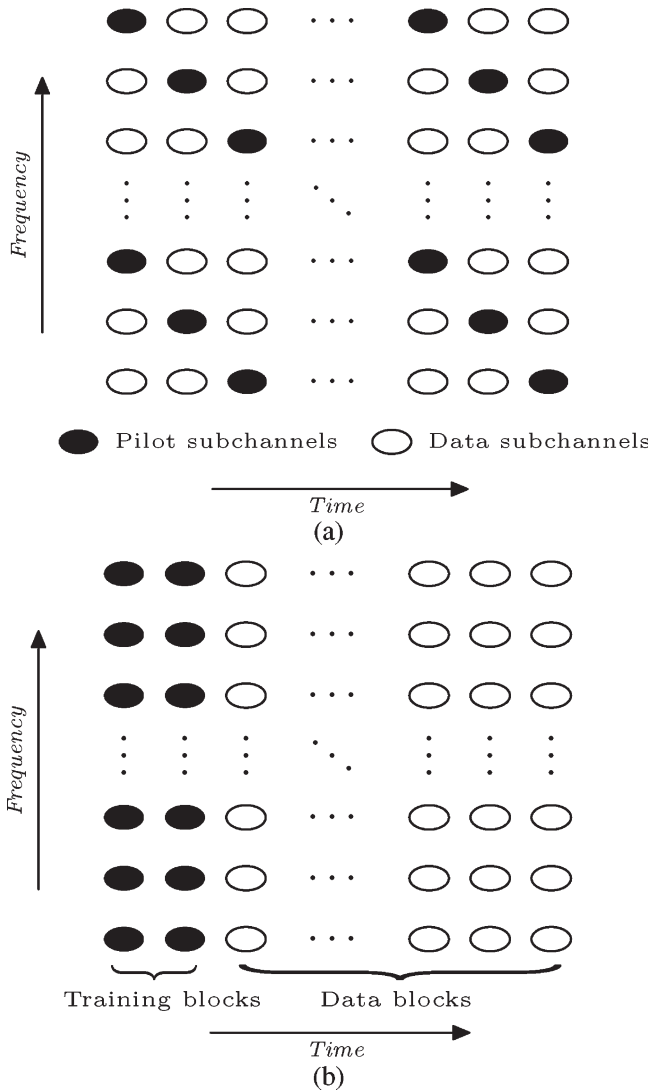


Fig. 2. Typical training blocks and comb pilots. (a) Comb pilots. (b) Preamble.

channels. In Section II-C, we will present approaches for time- and frequency-varying impairment mitigation.

From (1), the transmitted signal is a linear combination of the transmitted symbols. Therefore,  $s(t)$  is Gaussian alike from the central limit theorem in probability theory and has a very large PAPR compared with the signal from single-carrier modulation, which may cause nonlinear distortion of the amplifier at the transmitter. There have been many approaches to reduce the PAPR of OFDM signals. We will summarize them in Section II-D.

The multiple-receive antenna array, also called the adaptive antenna array (AAA), can be used in OFDM for interference suppression. AAA techniques have first been proposed for narrow-band TDMA in [41] to suppress cochannel interference. A comprehensive introduction of AAA techniques and beamforming can be found in [42]. The AAA techniques developed for flat fading channels can directly be applied to each subchannel in OFDM systems. The only challenge is how to estimate the spatial correlation of cochannel interference, which is required in finding the weights for interference suppression. By exploiting the correlation of cochannel interference at dif-

ferent subchannels in OFDM, instantaneous spatial correlation has successfully been obtained in [43]. MIMO techniques can also be used in OFDM to form MIMO-OFDM for performance and capacity improvement, which will be discussed in Section IV.

### B. Channel Estimation

In OFDM systems, CSI can be estimated using training symbols known at both the transmitter and the receiver. The training symbols may be inserted at different subchannels of different OFDM blocks, as shown in Fig. 2(a). These training symbols are more often called pilots. The CSI corresponding to the pilot subchannels is first estimated, and then, that corresponding to the data-bearing subchannels is obtained by interpolation. This is called pilot-aided channel estimation (PACE) [44]–[46]. In addition to interleaving the training symbols and the informative symbols by such frequency-division multiplexing, they may also be superimposed, which can be regarded as a special form of pilots [47]. This kind of training symbols are usually called superimposed pilots, which were first proposed to phase synchronization and originally called spread-spectrum pilots [48] and were later applied for channel estimation [49]–[53]. On the other hand, all training symbols may be arranged at the first (or couple of) OFDM blocks, as shown in Fig. 2(b). The training blocks in this case are sometimes called preamble. The CSI corresponding to the training blocks are first estimated, and that corresponding to the subsequent data blocks can be tracked and further improved with the help of the demodulated data. This is called decision-directed channel estimation (DDCE) [54], [55].

When channel statistics are unknown and CSI is treated as a deterministic parameter, maximum-likelihood (ML) channel estimation will be optimal and will approach the Cramer–Rao bound (CRB) [44]. For channels with AWGN, ML estimation of channel parameters is equivalent to finding channel parameters to minimize

$$\|\mathbf{x} - \mathbf{S}_p \mathbf{H}\|^2$$

where  $\mathbf{x}$  and  $\mathbf{H}$  are the received signal vector and channel frequency response vector, respectively, which are defined as<sup>4</sup>

$$\mathbf{x} = \begin{pmatrix} x_0 \\ \vdots \\ x_{N-1} \end{pmatrix} \quad \mathbf{H} = \begin{pmatrix} H_0 \\ \vdots \\ H_{N-1} \end{pmatrix}$$

and  $\mathbf{S}_p$  is the pilot symbol matrix, which is defined as

$$\mathbf{S}_p = \begin{pmatrix} s_0 & 0 & \cdots & 0 \\ 0 & s_1 & \ddots & 0 \\ \vdots & \ddots & \ddots & \vdots \\ 0 & \cdots & 0 & s_{N-1} \end{pmatrix}.$$

<sup>4</sup>The index that indicates the OFDM block is omitted in the subsequent discussion of this paper for simplicity.

It can easily be seen that ML estimation is the same as least-square (LS) estimation in this case. Direct calculation yields that LS estimation of the channel frequency response vector is

$$\hat{\mathbf{H}}_{\text{LS}} = \mathbf{S}_p^{-1} \mathbf{x}. \quad (7)$$

To reduce the computational complexity of ML or LS channel estimation,  $\mathbf{S}_p^{-1}$  in (7) can be calculated offline. Estimation in this case neither needs nor exploits the information of channel statistics. Consequently, estimation performance is usually not good enough; thus, it is often applied as initial estimation.

By exploiting channel statistics, channel estimation can significantly be improved. With the correlation matrix of the channel frequency response vector  $\mathbf{R}_H = E\{\mathbf{H}\mathbf{H}^H\}$ , linear minimum mean square error (LMMSE) channel estimation can be obtained. For LMMSE estimation, channel frequency responses are regarded as random variables. Estimation of the channel frequency response vector is found to minimize the mean square error (MSE), i.e.,  $E\|\hat{\mathbf{H}} - \mathbf{H}\|^2$ . Consequently

$$\begin{aligned} \hat{\mathbf{H}}_{\text{LMMSE}} &= \mathbf{R}_H \left( \mathbf{R}_H + \sigma_n^2 (\mathbf{S}_p \mathbf{S}_p^H)^{-1} \right)^{-1} \mathbf{S}_p^{-1} \mathbf{x} \\ &= \mathbf{R}_H \left( \mathbf{R}_H + \sigma_n^2 (\mathbf{S}_p \mathbf{S}_p^H)^{-1} \right)^{-1} \hat{\mathbf{H}}_{\text{LS}}. \end{aligned}$$

Compared with ML or LS estimation, LMMSE estimation has much better performance; however, it requires channel statistics and has higher computational complexity.

With a minor modification, the principle of the aforementioned LMMSE estimation can also be used to estimate channel frequency responses other than pilot subchannels as in PACE, which is also called optimal interpolation in [45] and [46].

Training or pilot symbols can be designed to facilitate channel estimation. For example, if they are with constant modulus, that is,  $|s_k|^2 = \sigma_s^2$  for  $k = 0, \dots, N-1$ , then  $\mathbf{S}_p \mathbf{S}_p^H = \sigma_s^2 \mathbf{I}$ ,  $\sigma_n^2 (\mathbf{S}_p \mathbf{S}_p^H)^{-1} = (\sigma_n^2 / \sigma_s^2) \mathbf{I}$ , and no matrix inversion operation is required. Furthermore, if the training or pilot symbols are identical, that is,  $s_k = s$  for  $k = 0, \dots, N-1$ , then even  $\mathbf{S}_p = s \mathbf{I}$  and  $\mathbf{S}_p^{-1} = (1/s) \mathbf{I}$ . As indicated in [54], the coefficient matrix for LMMSE estimation can be computed with the help of transform domain processing, such as Fourier transform or singular value decomposition. The correlation matrix of the channel frequency response vector is determined by the power delay profile (PDP) of the wireless channel. Therefore, if the accurate PDP is known in advance, then the coefficient matrix for LMMSE estimation can be calculated offline. When the accurate PDP is unknown in advance, different solutions have been suggested. For example, channel estimation may set up coefficients according to a uniform or an exponential PDP [21], [46], [54], [56].

Estimation performance can further be improved by a 2-D LMMSE estimator exploiting the time-domain correlation of channels in addition to the frequency-domain correlation. To reduce the computational complexity of the 2-D LMMSE estimator, several methods have been proposed, and some of them bear the same spirit of the 1-D algorithms discussed before. By exploiting the separable feature of channel correlation at time and frequency domains [45], [56], 2-D LMMSE estimation is decoupled into two cascade 1-D LMMSE estimations. Another solution partitions the 2-D time–frequency region into several

small regions and performs LMMSE estimation by only considering the correlation among adjacent subchannels in each small region [54].

1) *Pilot-Aided Channel Estimation*: Using pilot tones to estimate channel coefficients was first proposed in [57]. The two major issues of pilot-aided channel estimation are pilot design and interpolation.

The optimal design for the pilot pattern, power allocation, and number of pilots has extensively been studied [57]–[63], which critically depends on a proper criterion and the channel model. The impact of pilots on system performance for time-varying channels has first been analyzed in [58]. The optimal pilot design for frequency-selective channels has been investigated in [59] and [60], whereas that for doubly selective channels has been investigated in [61]. The pilots have been designed to minimize the MSE of channel estimation [57] or CRB [62], maximizing the channel capacity [59]–[61] and minimizing the symbol error rate [63]. An extensive review on the topic has been addressed in [47].

As indicated before, LMMSE estimation can be applied for joint channel estimation and interpolation. However, it requires channel statistics and high computational complexity [45], [46]. This motivates us to develop low-complexity interpolation algorithms.

Two of the simplest ways are piecewise constant and linear interpolation [2], [64]. However, more pilots are required for them to achieve acceptable performance in frequency-selective channels. If the statistics of channel variation either in the frequency domain (PDP) or in the time domain (Doppler spectrum) are known *a priori*, a high-order polynomial can be applied to accurately fit wireless channels [64], [65].

In addition to the linear and high-order polynomial-based interpolation, pilot-aided channel estimation can also be based on FFT [46], [66]. FFT-based channel estimation essentially uses a low-pass filter as an interpolator, and the filter is implemented in a transform domain [55]. Instead of using a low-pass filter, we may catch significant taps in the transform domain and turn off those trivial taps, which can improve the performance of channel estimation, particularly when the SNR of the system is low. However, when turning off trivial taps or using a low-pass filter, a useful component may also be removed, which will cause a large estimation error for those subchannels on the edge [55], [56].

2) *DDCE*: For DDCE, CSI at the preamble block(s) is first estimated and then used to demodulate and detect the symbols at the next data block. CSI can be tracked by using detected symbols or data, either hard decision or soft decision, as shown in [67]–[69]. For systems with error-correction coding, redundancy in coding can be exploited by iteratively performing soft symbol decision and channel estimation [68], [70].

A major problem of DDCE is error propagation, which is particularly severe for a system with a large Doppler frequency. This can be solved by periodically inserting training blocks [71]. Comb pilots may also be applied instead of training blocks [46], and channel tracking or prediction can be used for further performance improvement [72].

For coded OFDM systems, expectation–maximization algorithms can improve estimation performance by exploiting the

error probability information available at the decoder [73], [74]. In [75], the theory of robust statistics is applied to alleviate the error propagation of DDCE in fast fading channels.

3) *Comparison*: DDCE methods fit in systems operating in static or quasi-static channels. It particularly fits in systems in a slot transmission mode, such as wireless cellular systems. Initial channel estimation is provided with the training blocks and is then followed by tracking or prediction. Their major advantage is that they are able to provide high spectrum efficiency by using detected data as pilots. However, error propagation will be induced in fast fading channels [75].

PACE methods can reliably estimate channels both in static and time-varying channels by appropriately designing pilot patterns. It is desired for systems in a continuous transmission mode, such as in digital TV. For the same percentage of overhead, the comb-pilot-aided estimation outperforms the training-block-aided methods, particularly for a system with high mobility [46]. When a superimposed pilot sequence is considered, bandwidth efficiency can be improved at the expense of an increase of transmit power. Although the superimposed training may not be the optimal approach for channel estimation, it will lead to a low-complexity estimator [49]–[53].

### C. Time- and Frequency-Varying Impairment Mitigation

In this section, we will address time- and frequency-varying impairment mitigation.

Frequency-varying impairments are caused by the timing offset between the transmitter and the receiver or the delay spread due to a multipath of wireless channels. As shown in (5), the impact of delay spread is a multiplicative channel distortion on the demodulated signal if the CP or guard interval is long enough, which can easily be mitigated once CSI is estimated. The impact of timing offset is much simpler than that of delay spread. If the timing offset  $\tau$  is less than the CP, then it will cause a phase rotation of  $2\pi k\Delta f\tau$  to the symbol at the  $k$ th subchannel [4]. If the timing offset exceeds the CP, then IBI will be generated, in addition to the phase rotation. The phase rotation due to the timing offset is different for different subchannels. This property can be exploited to estimate the timing offset. We will address how to estimate the timing offset and compensate for its impact later on in this section.

Time-varying impairments are caused by the frequency offset or Doppler shift and can be modeled by a multiplicative function of time on the transmitted signal  $s(t)$ . If only the time-varying impairment is considered, the received signal will be

$$x(t) = \gamma(t)s(t)$$

where  $\gamma(t)$  is a function of time and represents the time-varying impairment. If the time-varying impairment is the frequency offset  $\delta f$ , then  $\gamma(t) = e^{j2\pi\delta f t}$ . If it is a Doppler shift, then  $\gamma(t)$  can be modeled as an  $f_d$ -bandlimited wide-sense stationary random process, where  $f_d$  is the Doppler frequency of wireless channels and is related to the mobile speed  $v$ , the carrier frequency of signal  $f_c$ , and the speed of light  $c$  by

$$f_d = \frac{f_c v}{c}.$$

From [4], the demodulated signal can be expressed as

$$\begin{aligned} x_k &= \frac{1}{T_s} \int_0^{T_s} x(t) e^{-j2\pi k \Delta f t} dt \\ &= a_0 s_k + \underbrace{\sum_{m \neq k} a_{k-m} s_m}_{\text{ICI}} \end{aligned} \quad (8)$$

where

$$a_m = \frac{1}{T_s} \int_0^{T_s} \gamma(t) e^{-j2\pi m \Delta f t} dt. \quad (9)$$

$\sum_{m \neq k} a_{k-m} s_m$  in (8) is ICI as a result of the time-varying impairment. From (8), time-varying impairments will cause the OFDM signal to spread in the frequency domain and generate ICI, which is very similar to ISI (in the time domain) caused by the frequency-varying impairment, such as frequency-selective fading or timing offset in single-carrier modulation. Depending on the type of time-varying impairments, the power of ICI is usually different. It can be proved in [4] and [76] that the power of ICI is upper bounded by

$$P_{ICI} = E \left| \sum_{m \neq k} a_{k-m} s_m \right|^2 \leq \frac{\pi^2}{3} (f_b T_s)^2 \sigma_s^2$$

where  $f_b = \delta f$  or  $f_d$  depends on the type of impairments.

As discussed in [4] and the references therein, the frequency offset may cause a simple symbol or subchannel shift,<sup>5</sup> ICI, or both. However, once it is estimated, its impact can completely be canceled. The Doppler shift, however, is random and is hard to deal with. We will introduce techniques to mitigate the Doppler shift and to estimate and correct the frequency offset.

1) *Timing-Offset Estimation and Correction*: The timing offset can be estimated with pilot- and nonpilot-aided techniques. After the timing offset is estimated, its integer part, which is a multiple of the sampling interval, is used to adjust the starting position of the FFT window, and its (residual) fractional part will generate a phase offset and can be compensated at each subchannel when we cancel the impact of the delay spread of wireless channels [77].

One way to estimate the timing offset is by means of the slope of the curve of carrier phases in pilots versus the index of pilots [78], [79]. In [80], ML estimation is obtained using pilots in an OFDM block. Since only the phases at pilot subchannels in one OFDM block are measured, frequency-selective fading will have an adverse effect on the performance.

The methods in [81] and [82] estimate the timing offset with the help of two consecutive OFDM blocks. An unbiased weighted LS timing offset estimator in AWGN and frequency-selective fading channels is developed in [81]. The method in [77] uses a noncoherent delay-locked loop for joint ML estimation of the timing and carrier phase in AWGN channels. However, the estimator is biased in frequency-selective channels.

<sup>5</sup>It is also called the tone shift.

When training symbols are used, a systematic method is introduced in [83] to jointly estimate the timing and frequency offset. The CRB is derived, and an ML estimator is developed that can achieve the lower bound. To reduce the complexity of the ML estimator, a simple suboptimal estimator with a special training symbol structure similar to that in [82] is developed in [83].

The decision-directed method in [84] can obtain unbiased estimation in both AWGN and frequency-selective fading channels. The timing offset estimator in [85] is based on time-domain data and requires no pilots. It performs better than frequency-domain estimators in high-mobility environments.

2) *Frequency-Offset Estimation and Correction*: From the perspective of its impact and signal processing, the CFO can be divided into integer and fractional parts. The integer part of the CFO is a multiple of the subchannel space  $\Delta f$ , which will cause a symbol or subchannel shift, that is, the transmitted symbol in one subchannel is shifted to another at the receiver. The fractional part results in the loss of orthogonality among subchannels [86] and generates ICI. Once the CFO is estimated, its impact can completely be canceled in the time domain by multiplying the received signal  $x(t)$  by the frequency shift factor  $e^{-j2\pi\delta ft}$ .

CFO estimation can be catalogued as pilot-based and blind (or nonpilot-based) methods according to the required *a priori* information. Pilot methods may be based on the preamble blocks in the time domain or on the pilot subchannels inserted at different OFDM blocks. In addition to accuracy, a crucial issue in CFO estimation is to cope with the ambiguity or to provide consistent estimation, which can successfully be addressed by designing training blocks with a special structure or by separately estimating the integer and fractional parts.

ML CFO estimation is developed by measuring the phase rotation between two preamble/training blocks [86]. This method can be extended to systems with only one training block as long as the first half of the training block is identical to the second half [87], [88].

When pilots are inserted at subchannels of different OFDM blocks, ML estimation is developed in [89] by computing the phase difference between pilot subchannels in adjacent OFDM blocks. More references in this topic can be found in [90]–[92] and the references therein.

Channel statistics can be used with pilot methods to further improve CFO estimation performance. An ML CFO estimator based on pilot symbols has been developed in [93] for time-varying flat fading channels. This estimator is modified in [94] to extend the estimation range with the help of phase information in the second-order statistics of wireless channels. Based on the methods in [93] and [94], two simple but robust methods are developed in [95], which take channel model mismatch into consideration. In [96], the method for AWGN channels in [97] is extended to frequency-selective fading channels. The extended method is found to be biased in [98]. Furthermore, a joint CSI and CFO estimator is also developed in [98], which is unbiased and approaches the CRB.

Blind or nonpilot-based methods exploit the special structure in the OFDM signal, such as virtual subchannels

(VCs)<sup>6</sup> or CP, to estimate the CFO and do not need an extra overhead. However, their performance depends on channel features since dispersive fading of wireless channels may destroy the structure of the received signals.

In general, the VC-based CFO estimators exploit ICI information in the VC when the frequency offset is present. Subspace methods, which are analogous to the spatial spectrum estimators in array processing, have been proposed in [99] for systems with consecutive VCs, which have the same form as the ML CFO estimator in [100]. However, the aforementioned VC-based estimators exhibit ambiguity for frequency-selective fading channels. Different approaches have been suggested to address the issue, such as nonconsecutively allocating null subchannels [90] or separately estimating the fractional and integer parts of the CFO [101].

By exploiting the repetitive property of the CP, a joint ML time synchronization and frequency-offset estimation has been proposed in [102], and an optimal ML estimator and a minimum variance unbiased CFO estimator have also been developed in [103]. Most of these earlier methods can only be applied for flat fading channels and estimate the CFO within the range of  $\Delta f/2$ , that is, the fractional part of the CFO. By jointly exploiting the VC and CP, the ML method in [104] can be extended to estimate the CFO in frequency-selective fading channels.

3) *Mitigation of ICI Caused by the Doppler Spread*: ICI may be caused by the CFO, phase noise, timing offset, and Doppler spread [76], [86], [105], [106]. However, ICI induced by the first three impairments can completely be compensated or corrected. Since the Doppler spread or shift is random, we can only mitigate its impact. The existing ICI mitigation techniques include frequency equalization, ICI self-canceling, time-domain windowing, coding, etc.

As we can see from (8), ICI in the frequency domain in OFDM systems is similar to ISI in the time domain in single-carrier systems. Consequently, those approaches dealing with ISI in single-carrier systems can immediately be used here. It is well known that matrix inversion is required to calculate the coefficients of an equalizer. To reduce its high complexity, various methods have been developed. In [107], the channel matrix is partitioned into block diagonal matrices by exploiting the fact that the frequency response within a symbol duration will linearly vary with time when the duration of an OFDM symbol is much less than the channel coherence time. In [108], a time-domain successive interference cancellation (SIC) detector is presented to remove ICI, which is similar to SIC widely used in multiuser detection. To alleviate the impact of ICI for channels with a large delay spread and a high Doppler shift, two methods have been developed in [109], which use a piecewise linear model to approximate the time-varying features of channels and then use the CP or two adjacent symbols to estimate the channel time variation information required for equalization. Since the symbol transmitted through each subchannel only spreads to several adjacent subchannels at the receiver, a finite-tap ICI equalizer is proposed in [110], which can significantly

<sup>6</sup>VCs are the subchannels at the low- and high-frequency edges that are set to zero for mitigating out-of-band radiation and direct-current offset.

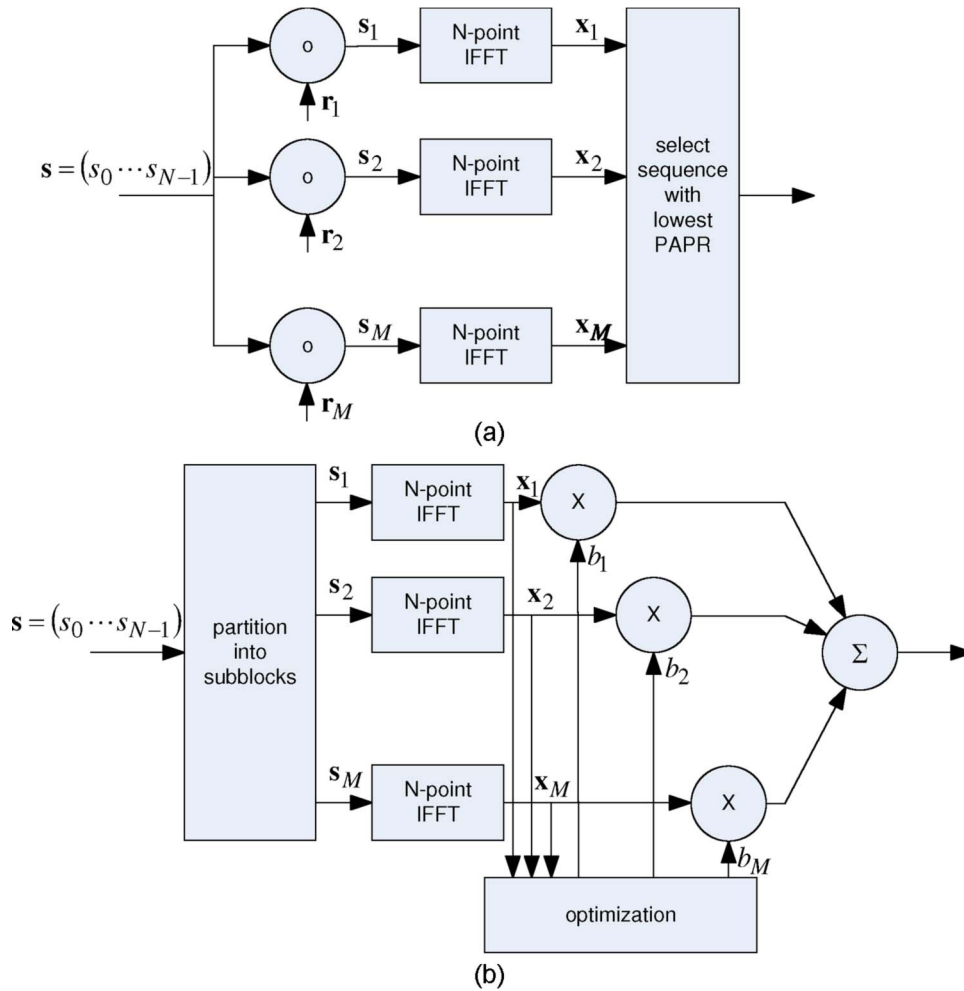


Fig. 3. SLM and PTS approaches. (a) Selective mapping approach. (b) Partial transmit sequence approach.

reduce the complexity at the expense of a minor performance degradation.

ICI self-cancellation is another simple but effective way. It is found in [111] that part of ICI can be canceled if the same symbol with different polarities is transmitted in two adjacent subchannels. The idea is extended to modulate the same data in several adjacent subchannels with optimal weights in [112]. A general ICI self-cancellation method is also proposed in [113]. Although these methods can effectively mitigate ICI, the data transmission rate is also reduced since it is, in fact, repetitive frequency coding.

By using time-domain windowing [114] and coding, the sensitivity of systems to ICI can be reduced. Borrowing the idea of partial response coding in single-carrier modulation to reduce the sensitivity to the timing offset, frequency-domain correlative coding is proposed in [115] to reduce ICI caused by the CFO without sacrificing the bandwidth efficiency. The method is extended to mitigate ICI induced by both the frequency offset and Doppler spread in [116].

By exploiting the unitary property of the ICI coefficient matrix, the Tomlinson–Harashima (TH) precoding method [117] and ICI suppression method [118] are also developed, which are applicable to open- or closed-loop, single- or multiple-antenna systems.

#### D. PAPR Reduction Techniques

As indicated before, the OFDM signal has a large PAPR. A traditional method dealing with the large PAPR is to back off the operating points of nonlinear power amplifiers; however, it severely reduces the efficiency of the power amplifiers. Therefore, by exploiting the special characteristics of the OFDM signal, various approaches have been proposed to cope with the issue. They include clipping and filtering [119], [120], selected mapping (SLM) [121]–[123], partial transmit sequence (PTS) [124], etc.

To reduce the PAPR of an OFDM signal, a clipper can directly be used. However, such nonlinear processing will cause in-band distortion and out-of-band radiation [125]. If the out-of-band interference is filtered out, then the PAPR of the clipped signal will regrow [119]. Therefore, if clipping and filtering are repeated several times, then both the PAPR and out-of-band radiation will be reduced, as proposed in [120]. However, the clipping and filtering techniques are unable to remove the in-band distortion. The technique is improved in [126] by limiting the distortion of each subchannel.

Fig. 3 demonstrates the basic principles of the SLM and PTS techniques. Both are distortionless compared with the clipping and filtering techniques. However, they are with heavy computational complexity and need to transmit an appropriate



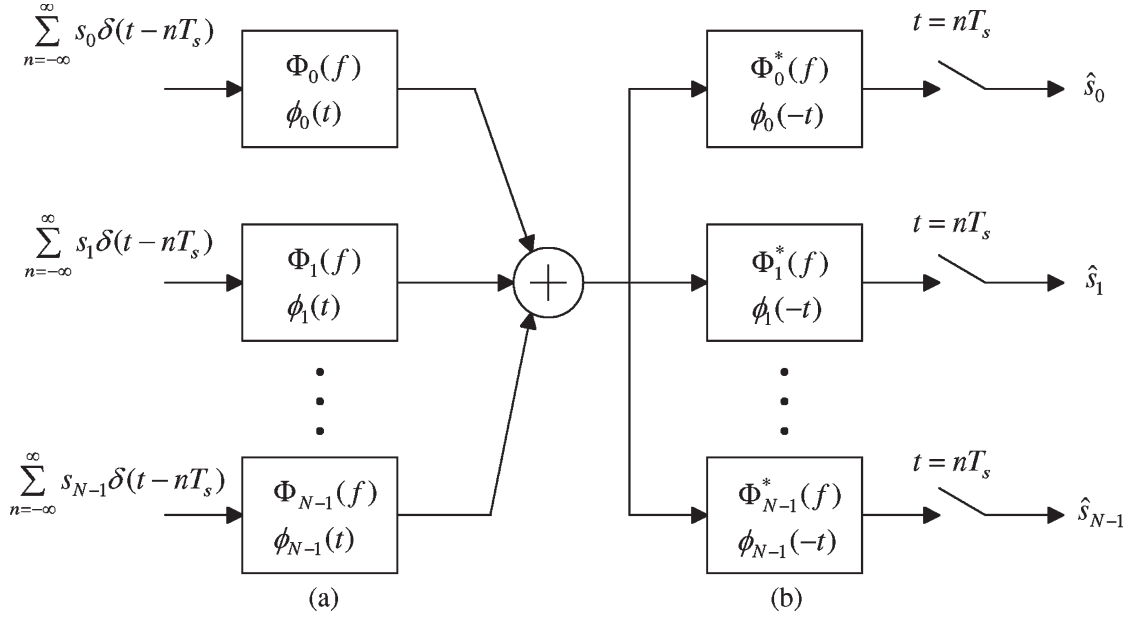


Fig. 4. Orthogonal approach for MC modulation.

side information indicating how the transmitter generates the signal.

As shown in Fig. 3(a), the SLM approach first multiplies the transmitted symbols  $\mathbf{s} = (s_0, \dots, s_{N-1})$  by  $M$  predetermined sequences  $\mathbf{r}_m = (r_{0m}, \dots, r_{N-1m})$  for  $m = 1, \dots, M$ , respectively, to obtain  $\mathbf{s}_m = (s_0 r_{0m}, \dots, s_{N-1} r_{N-1m})$ , converts  $\mathbf{s}_m$  into OFDM signals by inverse FFTs (IFFTs), and then selects the one with a minimum PAPR to transmit. If the receiver knows the (predetermined) sequence that enables a minimum PAPR, the transmitted symbols  $s_k$  can be recovered. The *book* on predetermined sequences  $\mathbf{r}_m = (r_{0m}, \dots, r_{N-1m})$  for  $m = 1, \dots, M$  are known to the transmitter and the receiver. Therefore, only the index of the predetermined sequence needs to be sent to the receiver for each OFDM signal. To further improve transmission efficiency, transmission of side information can be omitted by embedding it into an OFDM signal, for example, through a label inserted scrambler and an encoder [127], or by blindly detecting the optimal phase sequence index in the receiver [128]. A modified SLM scheme is also proposed in [129] to reduce the complexity of the original SLM scheme.

The PTS approach, as shown in Fig. 3(b), partitions transmitted symbols for an OFDM block into  $M$  disjoint subblocks of size  $L = N/M$  and pads zeros for  $N$ -point IFFT for each subblock. As a result, the  $n$ th component of  $\mathbf{x}_m$  will be

$$x_{nm} = \sum_{k=(m-1)L}^{mL-1} s_k e^{j2\pi \frac{kn}{N}}.$$

The PTS approach finds  $b_m \in \{-1, 1\}$  such that the PAPR of  $x_n = \sum_{m=1}^M b_m x_{nm}$  for  $n = 0, \dots, N-1$  is minimized. Similar to the SLM approaches, the sequence  $\{b_m\}_{m=1}^M$  to optimize the PAPR of  $\{x_n\}_{n=0}^{N-1}$  needs to be sent to the receiver for the receiver to detect the transmitted symbols. At the cost of a minor performance degradation, a suboptimal iterative PTS scheme has been proposed in [130] to reduce the computational complexity of the PTS.

In addition the aforementioned PAPR reduction approaches, there are many other techniques that we are unable to introduce one by one. They include tone injection [131], tone reservation [131], [132], active set extension [133], [134], coding [135]–[138], and companding [139], [140] approaches.

### III. OFDM-RELATED MODULATION AND ACCESS TECHNIQUES

There are many other modulation or access techniques related to OFDM. MC modulation is a general category of modulation to which OFDM belongs. A single-carrier system with frequency-domain equalization (SC-FDE) and energy-spreading transform (EST)-based modulation are two block transmission schemes that exploit the CP to mitigate the delay spread of wireless channels, which share the same spirit as OFDM. Furthermore, based on OFDM, many access techniques have been developed. MC-CDMA and OFDM access (OFDMA) are two of the examples. In this section, we will briefly describe MC modulation, SC-FDE, EST-based modulation, MC-CDMA, and OFDMA.

#### A. MC Modulation

MC modulation divides the whole bandwidth into many subchannels and transmits multiple data streams in parallel over them. Depending on specific modulation, the ways of dividing the available bandwidth or forming subchannels are different.

The first kind of MC modulation is called the *orthogonal approach*, which is shown in Fig. 4. It is one of the earliest MC techniques proposed by Chang [5]. The transmitted symbol  $s_k$  at the  $k$ th subchannel passes through the  $k$ th band-limited filter with the frequency response  $\Phi_k(f)$ . The frequency response of the filter  $\Phi_k(f)$  overlaps with adjacent filters  $\Phi_{k-1}(f)$  and  $\Phi_{k+1}(f)$ . It is proved in [5] that the transmitted symbols can be recovered at the receiver without any ICI if the magnitudes and

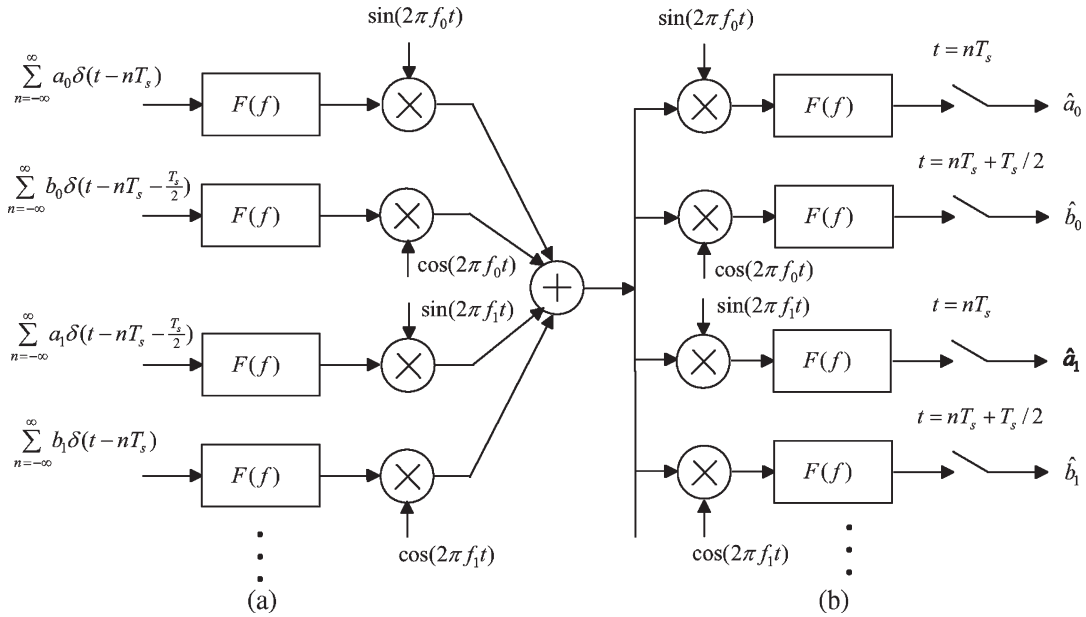


Fig. 5. Filter approach for MC modulation. (a) Transmitter. (b) Receiver.

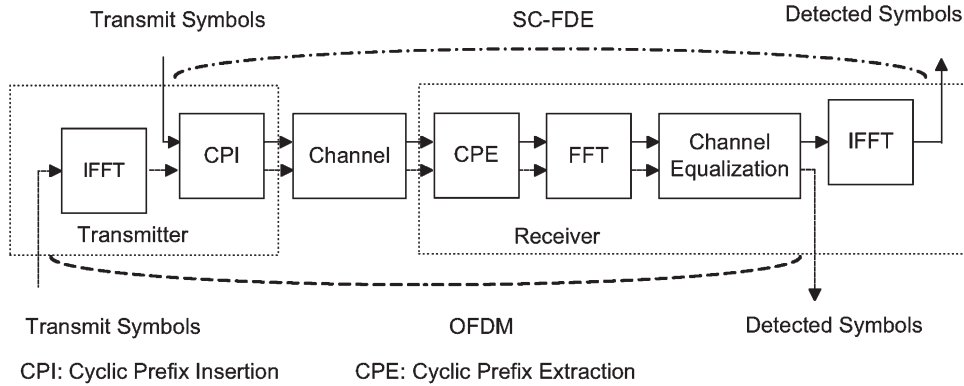


Fig. 6. SC-FDE versus OFDM.

phases of the filters satisfy certain conditions. An example of such filters has the following magnitude response:

$$|\Phi_k(f)| = \begin{cases} \cos\left(\pi \frac{f-f_k}{2\Delta f}\right), & f_k - \Delta f \leq f \leq f_k + \Delta f \\ 0, & \text{otherwise.} \end{cases}$$

Another kind of MC modulation is based on the filter bank and is called the *filter approach* [7]. As shown in Fig. 5, the frequency response of filters  $F(f)$  in the figure is identical and real. It has been proved [7] that there will be no ICI if  $F(f)$  in the figure satisfies the following two conditions:

$$\begin{aligned} F(f) &= 0, & |f| &\geq \Delta f \\ F^2(f) + F^2(\Delta f - f) &= 1, & 0 \leq f &\leq \Delta f. \end{aligned}$$

As before, the symbol duration and subchannel space satisfy  $T_s \Delta f = 1$ . As shown in Fig. 5, the in-phase and quadrature streams corresponding to the same subchannel are staggered by  $T_s/2$ . Adjacent channels are oppositely staggered so that the data streams that modulate the cosine carriers of the even-numbered channels are in-phase with the data streams that modulate the sine carriers of the odd-numbered channels, as well as conversely.

Although we still call  $T_s$  the symbol duration in the aforementioned two kinds of MC modulation, the length of the shaping pulse corresponding to each symbol is actually infinite since it is limited in the frequency domain. OFDM introduced in Section II is time limited, which is a critical difference from the aforementioned two kinds of MC modulation.

B. SC-FDE

OFDM transmits symbols in blocks with the CP or guard interval inserted between adjacent OFDM blocks to deal with ISI caused by the delay spread of wireless channels. SC-FDE in [141] also has a blockwise transmission format and incorporates the CP for the same purpose. Fig. 6 compares SC-FDE and OFDM.

Each symbol occupies the whole bandwidth in SC-FDE, whereas it only occupies a narrow subchannel in OFDM. Therefore, SC-FDE has frequency diversity even without employing channel coding. SC-FDE requires no FFT at the transmitter and employs two FFTs at the receiver for equalization, whereas OFDM uses one FFT at the transmitter and one FFT at the receiver to convert the OFDM signal to the time domain and

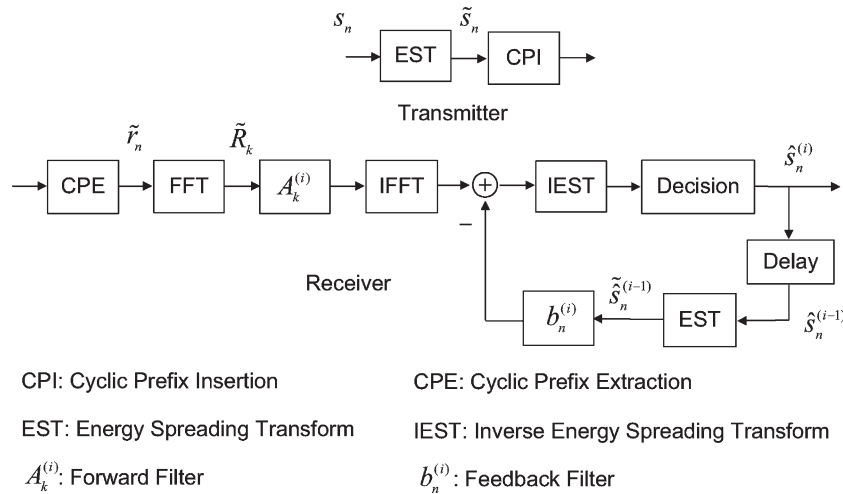


Fig. 7. EST-based modulation.

back to the frequency domain, respectively. As a consequence, SC-FDE has a less complex transmitter than OFDM. Another attractive feature of SC-FDE is that it has a lower PAPR than OFDM because of its inherent single-carrier structure. Therefore, SC-FDE is an ideal modulation for uplink of wireless communications, where the transmitter (handset) has a limited power amplifier and limited signal processing capability.

Single-carrier frequency-domain multiple access (SC-FDMA) [142] is a multiple-access scheme based on SC-FDE, whereas OFDMA is a multiple-access technique utilizing OFDM. Due to its inherent characteristics of a low PAPR, SC-FDMA is considered in the uplink transmission of the long-term evolution (LTE) cellular system, where power efficiency for mobile terminals is a premium. Similar to subchannel allocation for OFDMA, SC-FDMA has subband [142] and interleaved approaches [143] for allocating resources to users.

### C. EST-Based Modulation

Fig. 7 shows the block diagram of the EST-based modulation [144], [145]. A block of  $N$  symbols  $\{s_n\}_{n=0}^{N-1}$  are transformed into  $\{\tilde{s}_n\}_{n=0}^{N-1}$  by EST and appended with the CP for transmission. At the receiver, signal detection is iteratively performed by forward and (circular) feedback filters in the frequency and time domains, respectively. Symbol decisions  $\{\hat{s}_n^{(i)}\}_{n=0}^{N-1}$  are made after inverse EST. Without the feedback path in Fig. 7, the system is equivalent to OFDM if EST is chosen to be IDFT or IFFT, whereas it turns to SC-FDE if EST is an identity transform. Therefore, from that point of view, EST-based modulation is a general category of modulation that contains OFDM and SC-FDE as special cases.

To optimize the performance of EST-based modulation, EST should uniformly spread the energy of a symbol over both the frequency and time domains. Frequency-domain spreading obtains frequency diversity, whereas time-domain spreading increases reliability of the feedback signal. Therefore, neither OFDM nor SC-FDE is optimal in that sense since they have no spreading in the frequency and time domains, respectively. Furthermore, EST enables iterative detection at the receiver even without employing a channel code and a soft-output

channel decoder [146], which are the basic ingredients of turbo equalization [147], [148]. As shown in [144], EST can practically be implemented by concatenating a random permutation matrix and a unitary matrix, with each element having the same magnitude such as Fourier and Hadamard matrices. Therefore, the complexity of EST is essentially the same as that of FFT or fast Hadamard transform.

It has been proved in [144] and [145] that the postdetection error power diminishes as iteration proceeds when the SNR of the system is above a threshold, and after several iterations, the performance approaches the matched filter bound.

EST-based modulation is also related to linear constellation precoding (LPC) for OFDM [149]–[151]. LPC is similar to EST in the sense that it also spreads a symbol energy into multiple subchannels to obtain multipath diversity. Since LPC has been designed assuming the use of an ML detector at the receiver, subcarrier grouping [149] is a natural approach to reduce the size of LPC and, consequently, the complexity of ML detection. On the other hand, EST has been designed assuming the use of an iterative linear detector at the receiver. Therefore, block sizes such as  $N = 2048$  or  $N = 4086$  are usual for EST-based modulation. EST-based modulation can also be extended to MIMO systems to reach the performance of genie-aided signal detection [152].

### D. MC-CDMA

CDMA is a multiple-access technique that has widely been used before. For broadband transmission, MC-CDMA, which combines MC and CDMA, can deal with frequency-selective fading and provide multiple-access facility. In MC-CDMA, a symbol energy is spread into multiple chips by a spreading sequence. Each chip is then transmitted over a subchannel. The chips corresponding to the same symbols may be distributed along the time or frequency dimension. Multiple users (MUs) may transmit at the same time and frequency, and the transmitted symbols from each user are detected by using the difference of spreading sequences/codes for different users.

Depending on the way that the spreading chips are distributed, MC-CDMA can be classified into two different categories.

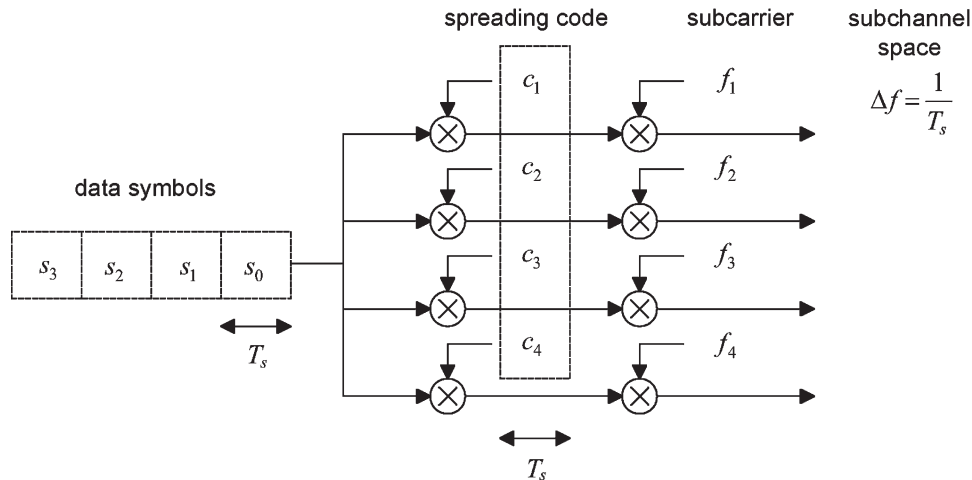


Fig. 8. MC-CDMA transmitter with four subcarriers.

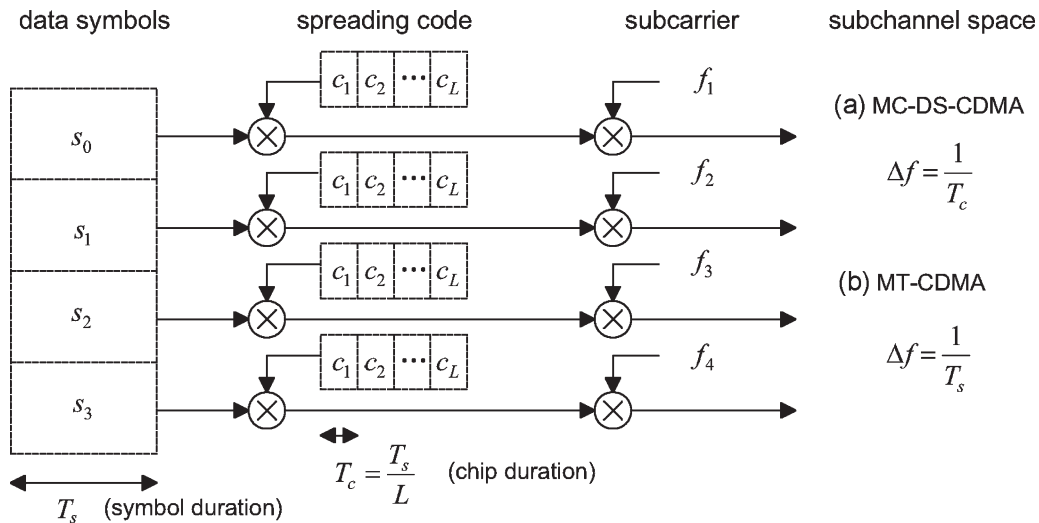


Fig. 9. Transmitter of MC-DS-CDMA and MT-CDMA with four subcarriers.

The MC-CDMA transmitter in the first category is shown in Fig. 8. In this scheme, the chips corresponding to the same symbols are spread over different subchannels of the same MC block. MC-CDMA proposed in [153]–[155] belongs to this category. For the other category, the chips corresponding to the same symbol are allocated to different MC blocks of the same subchannel, as shown in Fig. 9. Depending on the subchannel space, the schemes in the second category can be divided into MC direct-sequence CDMA (MC-DS-CDMA) [156] and multitone CDMA (MT-CDMA) [157]. For MC-DS-CDMA, the spectrum of each subchannel is shown in Fig. 10(a), and the orthogonal condition is satisfied. For MT-CDMA, the orthogonal condition is held *before* spreading. However, after spreading, the spectrum for each individual subchannel is extended, as shown in Fig. 10(b), and the orthogonal condition is not anymore held. A comprehensive description of MC-CDMA techniques can be found in [158] and [159].

E. OFDMA

OFDMA is a multiple-access scheme that has characteristics of OFDM and frequency-division multiple access. OFDM

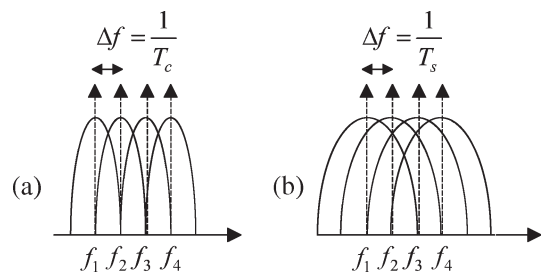


Fig. 10. Spectra of (a) MC-DS-CDMA and (b) MT-CDMA for four subcarriers.

transmits data from one user within a time slot, whereas OFDMA simultaneously transmits data for MUs. Inherited from OFDM, OFDMA is also immune against multipath and has other favorable characteristics. OFDMA was proposed for several broadband wireless systems such as the LTE downlink of cellular systems [160], IEEE802.16 standards for wireless metropolitan area network, and digital video broadcasting return channel terrestrial [161].

In OFDMA, the entire bandwidth is divided into a number of subchannels for parallel transmission of symbols from different

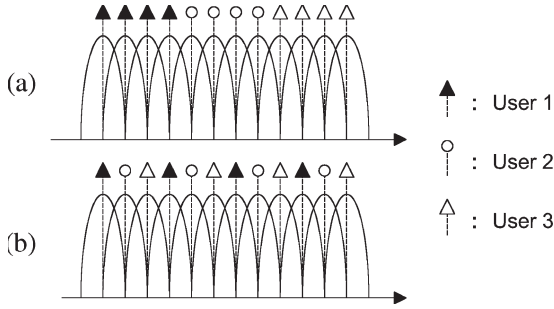


Fig. 11. Subcarrier allocation schemes for 12 subcarriers and three users. (a) Subband scheme. (b) Interleaved scheme.

users. How to assign subchannels to each user is an important issue and greatly influences the system performance. Fig. 11 shows two examples of the carrier assignment scheme: 1) the subband scheme and 2) the interleaved scheme. In the subband scheme, a group of adjacent subchannels is allocated to the same user, whereas in the interleaved scheme, uniformly spaced subchannels are allocated to the same user. The interleaved scheme can obtain frequency diversity, whereas channel estimation is easier for the subband scheme. Therefore, the interleaved scheme is preferable for fixed wireless communications, where channel slowly changes with time and channel estimation is easier than fast time-varying channels. The synchronization issues of OFDMA are summarized in [162]. For wireless links with channel feedback for adaptive subchannel/subcarrier allocation, the subband scheme requires less information for channel feedback [163], [164]. By adaptive subchannel allocation in OFDMA, the system capacity can significantly be increased, as shown in [165] and [166]. More information in this topic can also be found in [167], [168], and the references therein.

#### IV. MIMO TECHNIQUES IN OFDM

MIMO techniques or space–time processing can be used in wireless communications for diversity gain and capacity improvement [33]–[37]. Recent books [169]–[171] have given a comprehensive introduction of MIMO techniques. Here, we focus on special issues when MIMO techniques are used with OFDM.

Most of MIMO techniques are developed for flat fading channels. However, multipath will cause frequency selectivity of broadband wireless channels. Therefore, MIMO-OFDM, which has originally been proposed to exploit OFDM to mitigate ISI in MIMO systems, turns out to be a very promising choice for future high-data-rate transmission over broadband wireless channels. The earliest work in MIMO-OFDM can be found in [172] and [173]. Since that time, MIMO-OFDM has become a very popular area in wireless communications, particularly in the past several years [174]–[179]. In this section, we only very briefly provide an introduction of the topic.

##### A. Basic MIMO-OFDM

A MIMO-OFDM system with  $M_t$  transmit antennas and  $M_r$  receive antennas is shown in Fig. 12. Space–time processing

in the figure may exploit any space–time techniques developed for flat fading channels. Although it is still called space–time processing, it actually processes signal in the space and frequency domains in MIMO-OFDM.

As shown in the figure, space–time processing converts the transmitted symbol or data stream  $\{s_n\}$  into  $M_t$  substreams  $\{s_k^{(m)}\}$ , through simple multiplexing or STC, for OFDM modulation and transmission through different antennas. If only a multipath of wireless channels is considered and the CP is long enough, the demodulated signal at each receive antenna is a superposition of those from different transmit antennas and can be expressed as [4], [173]

$$x_k^{(i)} = \sum_{m=1}^{M_t} H_k^{(i,m)} s_k^{(m)} + n_k^{(i)} \quad (10)$$

for  $i = 1, \dots, M_r$ , where  $H_k^{(i,m)}$  denotes the frequency response at the  $k$ th subchannel corresponding to the  $m$ th transmit and the  $i$ th receive antenna, and  $n_k^{(i)}$  is the impact of channel noise at the  $k$ th subchannel of the  $i$ th receive antenna, which is usually independent for different  $k$  or  $i$ , Gaussian, and with zero mean. Equation (10) can also be written in matrix form as

$$\mathbf{x}_k = \mathbf{H}_k \mathbf{s}_k + \mathbf{n}_k \quad (11)$$

where  $\mathbf{x}_k$ ,  $\mathbf{s}_k$ ,  $\mathbf{n}_k$ , and  $\mathbf{H}_k$  are the received signal vector, transmitted signal vector, noise vector, and channel matrix at the  $k$ th subchannel, respectively, and are defined as

$$\begin{aligned} \mathbf{x}_k &= (x_k^{(1)} \ \dots \ x_k^{(M_r)})^T \\ \mathbf{s}_k &= (s_k^{(1)} \ \dots \ s_k^{(M_t)})^T \\ \mathbf{n}_k &= (n_k^{(1)} \ \dots \ n_k^{(M_r)})^T \\ \mathbf{H}_k &= \begin{pmatrix} H_k^{(1,1)} & \dots & H_k^{(1,M_t)} \\ \vdots & \ddots & \vdots \\ H_k^{(M_r,1)} & \dots & H_k^{(M_r,M_t)} \end{pmatrix}. \end{aligned}$$

The MIMO-OFDM system model can easily be generalized into an MU MIMO-OFDM system model by combining various multiple-access technologies, such as OFDMA, TDMA, CDMA, or space-division multiple access [179]. For MU MIMO-OFDM systems, MU interference (MUI) also degrades performance.

In the previous discussion, we have not taken into count the effect of ICI caused by the frequency offset or high Doppler spread. Similar to SISO-OFDM systems, MIMO-OFDM could be impaired by ICI. For MIMO-OFDM systems, ICI due to the Doppler shift is analyzed in [180], and a time-domain filtering-based ICI mitigation technique is also developed for fast fading wireless channels, where the channel cannot anymore be assumed to be constant within an OFDM symbol. ICI due to the frequency offset is investigated in [181]. Approaches to mitigate both MUI and ICI in MU MIMO-OFDM is developed in [182], which exploit TH precoding and iterative equalization.

CSI is required for signal detection of most of MIMO-OFDM systems, which can be obtained by channel estimation. Similar

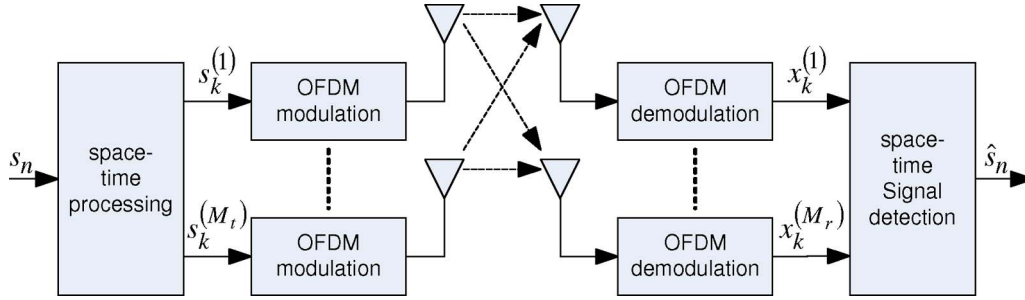


Fig. 12. MIMO-OFDM system.

to SISO-OFDM systems, the accuracy of channel estimation directly affects the performance of MIMO-OFDM systems. Therefore, channel estimation is an important topic to facilitate MIMO-OFDM for wireless communications. Channel estimation techniques for SISO-OFDM systems have been addressed in Section II-B. The challenge of channel estimation in MIMO-OFDM stems from that each receive signal corresponds to several channel parameters, which we also need to take into consideration when designing training or pilot sequences. By exploiting the correlation of channel parameters at different subchannels and OFDM blocks, channel estimation for MIMO-OFDM has been developed in [173]. The details on approach development, performance analysis, and enhanced approaches can be found in [4], [173], [176], and [183]. To optimize channel estimation performance, a pilot or training design is also very important. It is shown in [4] that the MSE of channel estimation can be minimized if the training sequence satisfies the following criterion:

$$\sum_{k=0}^{N-1} s_k^{(i)} \left( s_k^{(j)} \right)^* e^{j2\pi \frac{nk}{N}} = 0$$

for all  $1 \leq i \neq j \leq M_t$  and  $0 \leq n \leq N_o - 1$ . Furthermore, the training sequence that satisfies the aforementioned criterion also yields a low-complexity channel estimator [183]. A design of the aforementioned optimal training sequence has been proposed in [183]. The pilot sequence proposed in [184] also complies with the aforementioned optimal training criterion. More information on channel estimation of MIMO-OFDM can be found in [40], [185]–[188], and the references therein.

In the subsequent sections, we will focus on spatial multiplexing (SM) and STC for MIMO-OFDM systems.

### B. Performance Optimization

Bell Laboratories layered space–time architecture (BLAST) [33], including vertical BLAST (V-BLAST) and diagonal BLAST, can directly be used in OFDM for SM. Compared to SISO-OFDM systems, SM-based MIMO-OFDM approaches can achieve high data rates over wideband channels [175], [189], [190]. For SM-based MIMO-OFDM systems, performance and complexity tradeoff is one of the critical issues. In addition, link adaptation by utilizing CSI and/or channel quality information (CQI) at the transmitter and CSI or CQI feedback are also important to SM-based MIMO-OFDM systems. We will address these issues in the following sections.

1) *Performance and Complexity Tradeoff*: To support high-data-rate wireless access with required quality-of-service, turbo codes and turbo detection have been introduced in SM-based MIMO-OFDM systems. However, the complexity of soft demapping/decoding for turbo techniques exponentially increases with the number of transmit antennas and modulation order. To reduce the complexity, nonlinear interference cancellation with linear demapper and hard decision are used in [191]. In [192], a decision-feedback receiver, which takes decision errors into consideration for equalization formulation and soft-bit demapper, is developed to mitigate the effect of error propagation in V-BLAST OFDM systems.

By exploiting the correlation of frequency response at different subchannels, detection complexity in SM-based MIMO-OFDM systems can be reduced through subchannel grouping [193] or interpolation [194], where Gram–Schmidt operation or QR decomposition is used instead of direct matrix inversion to calculate the nulling vectors [189].

2) *Link Adaptation and CSI Feedback*: Link adaptation, including transmission power and data-rate allocation, has been studied for MIMO systems with flat fading channels or SISO-OFDM with frequency-selective channels. As in SISO-OFDM or MIMO systems, either transmission power adaptation or data-rate adaptation can improve the performance of MIMO-OFDM. However, performance improvement is limited if both are used [195]. Most of link adaptation approaches for MIMO or SISO-OFDM systems can directly be extended to MIMO-OFDM systems with frequency-selective channels. The only difference is that it is in the subchannel/frequency domain for SISO-OFDM, in the spatial domain for MIMO systems, and in both domains for MIMO-OFDM.

CSI or CQI is required for link adaptation at the transmitter, which is usually estimated at the receiver and fed back to the transmitter. The overhead of CSI or CQI feedback is sometimes huge, particularly for MIMO-OFDM systems. Therefore, CSI or CQI feedback reduction for MIMO-OFDM systems has become an important research topic in the past several years. Based on precoding approaches [196]–[198] for MIMO systems, various approaches [199]–[205] have been obtained for MIMO-OFDM systems. They can be divided into two categories.

1) *Clustering*: Instead of feeding quantized CSI for each subchannel back to the transmitter, the subchannels in an OFDM block are divided into several clusters, and adjacent subchannels in a cluster share the same CSI or CQI [199]. Furthermore, various interpolation methods

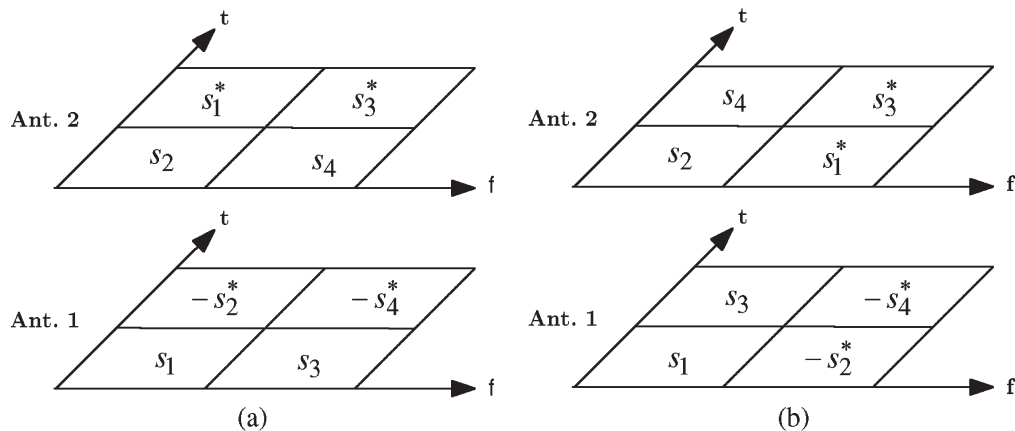


Fig. 13. Different combinations of Alamouti’s code and OFDM.

with different overheads for CSI feedback have been proposed [201]–[205].

- 2) *Tracking*: Since precoding matrices are viewed as points in the Grassmann manifold [196], the commonly used Euclidean distance does not apply. On the other hand, the frequency responses of wireless channels at adjacent subchannels are correlated, and therefore, the corresponding precoding matrices should be somehow *closed*. Based on this observation, tracking techniques have been developed [200], [202].

In addition to the aforementioned link adaptation approaches, power allocation or data-rate allocation without using CSI has also recently been investigated in [206] and [207].

### C. STFC

Most of space–time codes have originally been designed for flat fading channels by assuming no CSI or CQI at the transmitter. In general, STC can be divided into space–time block coding (STBC) [36], [37] and space–time trellis coding (STTC) [34]. STBC for OFDM is very popular due to its low-complexity ML decoding. STTC has also been considered for OFDM systems in [173] and [208]–[210] because of its better performance than STBC. At the very beginning, STC is directly combined with OFDM to mitigate ISI in broadband wireless channels. To exploit spatial and frequency diversities, the first STC, particularly for OFDM, has been constructed in [208]. Since that time, various STC design schemes have been proposed for OFDM.

There are two ways of combining STC with OFDM. One is to independently apply STC for each subchannel, and the other is to use STC in the space and frequency (subchannel) domains instead of in the space and time domains. Let us take Alamouti’s code [36], which is a special STBC, as an example for an OFDM system with two transmit antennas. Fig. 13(a) and (b) shows the two different combinations. If Alamouti’s code is used in each subchannel, it requires that the channel parameters are the same at two adjacent OFDM blocks; otherwise, there will be interantenna interference for low-complexity ML detection introduced in [36]. Therefore, this combination fits in OFDM systems with a lower Doppler frequency. On the other hand, if used in the frequency domain,

it requires that the frequency responses of the channel are the same at two adjacent subchannels of an OFDM block, which implies the very small delay spread of the channel. In reality, wireless channels are time varying and with delay spread. The impact of time variance or delay spread on Alamouti’s code for OFDM can be found in [211]. The channel estimation issue in this case has been investigated in [209].

Full diversity is not guaranteed if original STC for flat fading channels is directly used in OFDM. For example, we only obtain spatial diversity provided by multiple transmit antennas, and no frequency diversity is provided by broadband wireless channels in the aforementioned combinations of Alamouti’s code and OFDM.

To design optimal space, time, and frequency coding (STFC) with maximum diversity and coding gain, a full-diversity design criterion for STBC has been first proposed in [212]. Following the criterion, a variety of STBC schemes with full diversity has been developed in [213]–[215]. By analyzing the pairwise error probability, it has been proved in [209] that the achievable diversity order of a well-designed STBC-OFDM system should be the product of the numbers of transmitter antennas, receiver antennas, and multipaths of a wireless channel. The design criterion has then been further modified for LDPC-coded STC-OFDM systems in [216]. Different from the traditional orthogonal design of STBC, algebraic construction for binary phase-shift keying (PSK)/quaternary PSK (QPSK) space–frequency trellis codes is introduced in [217] based on the rank criteria proposed in [37]. More diversity gain can be obtained for channels with a slow time variance by generalizing codewords to simultaneously cross space, time, and frequency. In [218], *sum-of-ranks* and *product-of-determinants criteria* are proposed for a group of STFCs.

## V. APPLICATIONS

During the past decade, OFDM has been adopted in many wireless communication standards, including European digital audio broadcasting, terrestrial digital video broadcasting, and satellite–terrestrial interactive multiservice infrastructure in China. In addition, OFDM has been considered or approved by many IEEE standard working groups, such as IEEE 802.11a/g/n, IEEE 802.15.3a, and IEEE 802.16d/e.

The applications include wireless personal area networks, wireless local area networks, and wireless metropolitan networks. Currently, OFDMA is being investigated as one of the most promising radio transmission techniques for LTE of the *3rd Generation Partnership Project* (3GPP), International Mobile Telecommunications—Advanced Systems. Before introducing the major features of several OFDM applications, we briefly describe the design guideline of OFDM for wireless communications.

### A. Design Guideline

When OFDM is used for wireless communications, its parameters depend on the Doppler shift and delay spread of wireless channels, in addition to complexity consideration. From the discussion in Section II, to reduce the effect of Doppler shift, the symbol duration of OFDM should be chosen such that

$$T_s f_d \ll 1 \quad \text{or} \quad T_s \ll \frac{1}{f_d}$$

which implies that the symbol duration is constrained by the Doppler frequency of wireless channels, and symbol duration should be short to ensure small ICI. On the other hand, to deal with the delay spread of wireless channels, the length of the CP  $T_g$  must be larger than the delay span  $\tau_s$ . The modulation efficiency of OFDM depends on the ratio of the length of the CP and the symbol duration. Therefore, the symbol duration of OFDM should be as large as possible from the efficiency point of view. A state-of-the-art design should take both the Doppler shift and delay spread of wireless channels into consideration and balance their impacts.

Once the symbol duration  $T_s$  is determined, the subchannel space  $\Delta f$  will be determined using the orthogonal principle, i.e.,  $T_s \Delta f = 1$ . If the available bandwidth of an OFDM signal is  $B_w$ , the number of subchannels will be determined by

$$N = \left\lfloor \frac{B_w}{\Delta f} \right\rfloor$$

where  $\lfloor x \rfloor$  denotes the largest integer less than or equal to  $x$ . If FFT is used for OFDM modulation and demodulation, the length of FFT will usually be

$$M = 2^k \quad \text{and} \quad k = \lceil \log_2 N \rceil$$

where  $\lceil x \rceil$  denotes the smallest integer larger than or equal to  $x$ . From the length of the FFT and symbol duration

$$f_s = \frac{M}{T_s}$$

### B. Mobile and Fixed Wireless Systems

OFDM has been adopted in IEEE 802.16 standards [219] to support peak data rate up to 75 Mb/s at the frequency bands under 11 GHz. OFDMA in IEEE 802.16-2004 fixes the size of FFT to be 256 and varies the subchannel space according to the bandwidth of the system. Different from IEEE 802.16-2004, OFDMA in IEEE 802.16e-2005 maintains the same

TABLE I  
SOME PARAMETERS OF IEEE 802.16. (a) IEEE 802.16-2004.  
(b) IEEE 802.16e-2005

Bandwidth, $B_w$ , (MHz)	1.75	3.5	5.5	7
Subchannel space, $\Delta f$ , (kHz)	7.81	16.6	25.0	31.3
Symbol Duration, $T_s$ , ( $\mu$ sec.)	128	64	40	32
Sampling frequency, $f_s$ , (MHz)	2	4	6.32	8
FFT size, $M$	256	256	256	256

Bandwidth, $B_w$ , (MHz)	1.75	3.5	5.5	7
Subchannel space, $\Delta f$ , (kHz)	10.9	10.9	10.9	10.9
Symbol Duration, $T_s$ , ( $\mu$ sec.)	91.4	91.4	91.4	91.4
Sampling frequency, $f_s$ , (MHz)	1.40	5.60	11.2	22.4
FFT size, $M$	128	512	1024	2048

TABLE II  
PARAMETERS OF MB-OFDM-BASED UWB

Bandwidth, $B_w$	528 MHz
Subchannel No., $N$	128
Subchannel space, $\Delta f$	3.2 MHz
Symbol duration, $T_s$	312.5 nsec.
Length of CP, $T_g$	60.6 nsec.
Length of GI, $T_{GI}$	9.5 nsec.

subchannel space, i.e.,  $\Delta f = 10.94$  kHz, and changes the sizes of FFT according to the bandwidth of the system. Table I [219, Sec. 8.4.6, pp.883–955] compares the parameters for (a) IEEE 802.16-2004 and (b) IEEE 802.16-2005, respectively.

In both OFDM and OFDMA modes, the ratio of the length of the CP to the symbol duration may be 1/4, 1/8, 1/16, or 1/32, and the modulation scheme may be QPSK, 16 quadratic-amplitude modulation (16QAM), or 64QAM, depending on the channel environments and the targeted data rate. In addition, antenna arrays may be used for diversity and interference suppression. STC is also optional in IEEE 802.16 to increase the data rates and extend the coverage.

In the downlink of 3GPP LTE, OFDMA is a basic modulation scheme, which is with the length of the CP  $T_g = 4.7/16.74 \mu$ s (short/long CP) and the subchannel space  $\Delta f = 15$  kHz.

### C. MB-OFDM for UWB Systems

Multiband OFDM (MB-OFDM) was once a standard candidate for the IEEE 802.15.3a working group [220]–[229] for ultrawideband (UWB) systems. The basic idea of MB-OFDM is to divide the spectrum into several subbands, and a data stream is transmitted over each band by OFDM. The parameters of OFDM are listed in Table II. It should be noted that the actual bandwidth of the OFDM signal is 409.6 MHz, although the bandwidth of each subband is 528 MHz. Interleaving is used to exploit frequency diversity. The MB-OFDM-based UWB system achieves data rates ranging from 55 to 480 Mb/s over distances up to 10 m.

The combination of MIMO and MB-OFDM has also been investigated [227], [228] for high-data-rate transmission. However, there is some argument on whether we need multiple antennas in a UWB system because it should be with low complexity and low cost, whereas multiple antennas required by MIMO increase the cost of transceivers and obviously contradict them.



#### D. Cognitive Radio

Cognitive radio [230] has emerged as a promising technology to solve the current spectrum scarcity problem. Dynamic spectrum management and access is one of the key functions of cognitive radio [230]. OFDM can be used to construct the transceiver of cognitive radio by virtue of its flexibility for subchannel assignment and power allocation [231]–[233].

Secondary (unlicensed) users in cognitive radio exploit *spectrum holes*, which are bands that are not used by primary (licensed) users, and should not interfere with the operation of the primary users. Therefore, the available spectrum for the secondary users is usually disjoint bands. Furthermore, the available bands change with the activities of the primary users, which require the secondary users to flexibly adjust the frequency bands of their modulated signals. Moreover, geographic separation introduces discrepancies of the available spectrum between the transmitter and the receiver, particularly at the initial link establishment stage. Motivated by these demands, interleaved OFDM-based transform domain communication system (TDCS) has been proposed for cognitive radio [231]–[235]. In [234] and [235], the performance of OFDM-based TDCS has been investigated, and bit-error-rate formulas have been obtained. However, there are many unsolved issues for OFDM-based cognitive radio networks, such as spectrum sensing [233], interference identification [236], and transceiver design [234]–[239]. Due to the attractive advantages of MIMO-OFDM, a generalized design of MIMO-OFDM-based cognitive radio has been proposed in [240].

#### VI. CONCLUDING REMARKS

In this paper, we have briefly described OFDM for wireless communications. We start with the basic principle of OFDM and techniques to deal with impairments in wireless systems, including channel estimation, timing- and frequency-offset estimation, ICI mitigation, and PAPR reduction. Then, we introduced related modulation and access schemes, such as OFDM, SC-FDE, EST-based modulation, MC-CDMA, and OFDMA. We have also summarized the MIMO techniques for OFDM and the wireless applications of OFDM.

The OFDM-related technique has been invented over 40 years ago. OFDM for wireless communications has intensively been an active research area in the past 10 years. It is not our intention and is impossible either to provide an exhaustive literature search in the area through this paper. Due to page limit, we do not include performance optimization in OFDM systems, techniques on joint channel, time- and frequency-offset estimation, or applications other than wireless.

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