Chapter 2

OFDM Fundamentals

Despite being a nearly 50-year-old concept, it is only in the last decade that OFDM becomes the modem of choice in wireless applications. One of the biggest advantages of an OFDM modem is the ability to convert dispersive broadband channels into parallel narrowband subchannels, thus significantly simplifying equalization at the receiver end. Another intrinsic feature of OFDM is its flexibility in allocating power and rate optimally among narrowband sub-carriers. This ability is particularly important for broadband wireless where multipath channels are "frequency-selective" (due to cancellation of primary and echoed signals). From a theoretical standpoint, OFDM was known to closely approximate the "water-filling" solutions of information theory that are capacity achieving. Some early work of Weinstein and Ebert [7] and Hirosaki [8] based on an FFT implementation of OFDM achieved both complexity and decoded bit count that was comparable to single-carrier counterparts. OFDM potential came to fruition in the designs of discrete multi-tone systems (DMT) for xDSL/ADSL applications, IEEE 802.11.a wireless LAN, digital broadcasting systems DAB-T/DVB-T, the recent 802.16 broadband wireless access. A highlight of the wireless OFDM landscape is depicted in Figure 2.1.

This chapter describes the underlying principles of OFDM modem. To better appreciate the capability of OFDM in combating channel impairments, we will first devote the space to characterize the wireless fading phenomenon. By identifying the types of signal distortions a channel may cause, the choice of OFDM as a broadband modem solution will become evident.

2.1 Broadband radio channel characteristics

Two difficulties arise when a signal is transmitted over the wireless medium. The first is *envelope fading*, which attenuates the signal strength in an unpredictable way. The other is *dispersion*, which alters the original signal waveforms in both time and frequency domains.

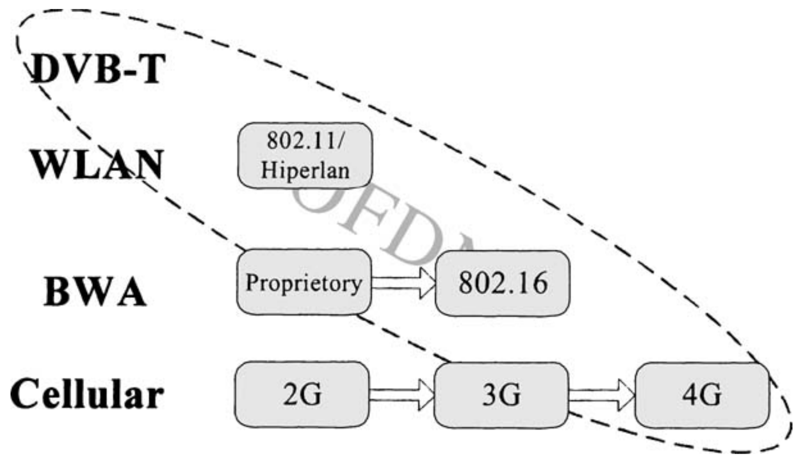

Figure 2.1: OFDM in broadband wireless networks

2.1.1 Envelope fading

The envelope fading manifests itself in the form of fluctuations in amplitude of received signals. The main causes are multipath reflections. Consider a scenario where the transmitted signal arrives at the receiver through two paths with negligible delay between them. The random scattering gives rise to different path attenuations in α_1, α_2.

$$x(t) = \alpha_1 s(t) + \alpha_2 s(t) = (\alpha_1 + \alpha_2)s(t).$$

In this case, the channel response can be modelled as a single delta function with a random envelope. Assuming α_1, α_2 are equal strength complex Gaussian, then the envelope of their sum, $r = |\alpha_1 + \alpha_2|$, obeys a *Rayleigh* distribution:

$$p(r) = \frac{r}{\sigma^2} e^{-\frac{r^2}{2\sigma^2}}$$

with the mean value and variance

$$E\{r\} = \sigma\sqrt{\tfrac{\pi}{2}}, \quad \sigma_r^2 = \sigma^2 \left(\tfrac{4-\pi}{2}\right)$$

In the case when the multipath components are not of the same strength (e.g., dominant line-of-sight scenarios), the envelope can be more accurately described by the *Rice* distribution [2].

In addition to the signal strength, the wireless medium may also affect the original signal through dispersion, which includes time dispersion (frequency selective) and frequency dispersion (time selective) fading.

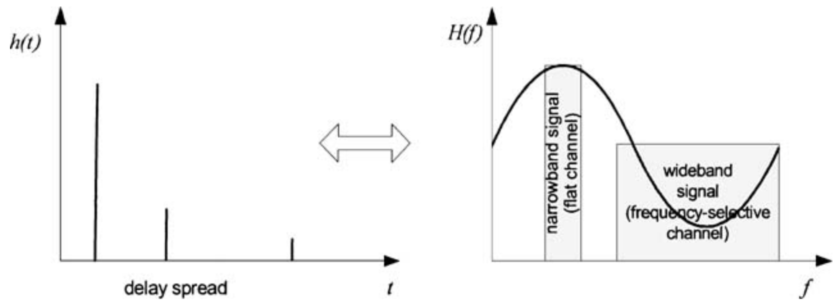

Figure 2.2: A time dispersive (frequency-selective) channel and its effect on narrow- and broad-band signals

2.1.2 Time dispersive channel

The arrival time of scattered multipath signals are inevitably distinct. Whether these delays smear the transmitted signal depends on the product of the signal bandwidth and the maximum differential delay spread. A pictorial view of the time dispersive channel is depicted in Figure 2.2.

The multipath channel can be represented as a linear transfer function $h(t)$. Because of the different propagation delays, the channel impulse response is superposition of delayed delta functions:

$$h(t) = \sum_{m=0}^{M-1} \alpha_i \delta(t - \tau_m)$$

In the case of Figure 2.2, $M = 2$.

Since the multipath delays, $\{\tau_m\}$, are distinct, the frequency response of $H(f) = \mathcal{F}\{h(t)\}$ will exhibit amplitude fluctuation. Such fluctuation in the frequency domain will distort the waveform of a broadband signal. More specifically in digital communication, a channel is considered *frequency-selective* if the multipath delays are distinguishable relative to the symbol period T_{symbol}:

$$\tau_{\max} - \tau_{\min} \approx T_{symbol} = \frac{1}{\text{BW of signal}} \Leftrightarrow (\tau_{\max} - \tau_{\min}) \times (\text{BW of signal}) \approx 1$$

On the other hand, if the signal bandwidth is sufficiently narrow, the channel frequency response within the signal bandwidth can be approximated as constant. A wireless channel is considered *flat* if the multipath delays are indistinguishable relative to the symbol period:

$$\tau_{\max} - \tau_{\min} << T_{symbol} = \frac{1}{\text{BW of signal}} \Leftrightarrow (\tau_{\max} - \tau_{\min}) \times (\text{BW of signal}) << 1$$

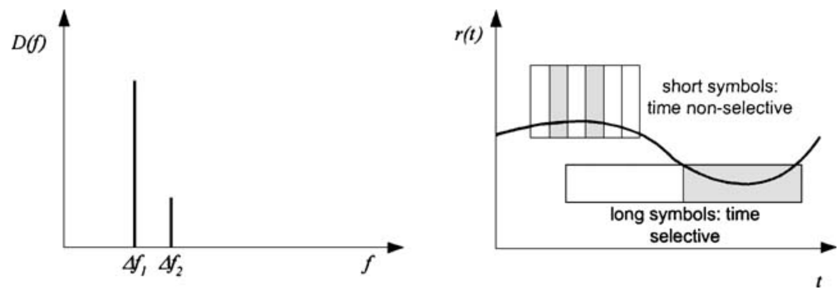

Figure 2.3: A frequency dispersive (time-selective) channel and its effect on short and long symbols

The often used parameters in characterizing a time dispersive channel include [2]:

- mean excess delay
- root-mean squared (rms) delay spread: τ_{rms}
- excess delay spread
- coherence bandwidth: B_c.

2.1.3 Frequency dispersive channel

The short-term fluctuation of the received signal in time domain can be best explained by the Doppler effects due to movement of the transmitter, the receiver, or the environment. The Doppler effect is multiplicative in time, rendering the channel impulse response linear, but time variant.

Consider Figure 2.3 which depicts the Doppler shifts associated with two multipaths in the frequency domain. For simplicity, let us assume the delay spread between the two multipath signals is negligible. At the baseband, the received signal is given by

$$\begin{aligned} x(t) &= s(t)e^{j2\pi\Delta f_1 t} + \alpha s(t)e^{j2\pi\Delta f_2 t} \\ &= (e^{j2\pi\Delta f_1 t} + \alpha e^{j2\pi\Delta f_2 t})s(t). \end{aligned}$$

The Dopplers introduce two types of distortion effects to the received signals: (i) signal variation over time, and (ii) broadened signal spectrum. Define channel coherence time as

$$T_c = 1/\Delta f_{\max}$$

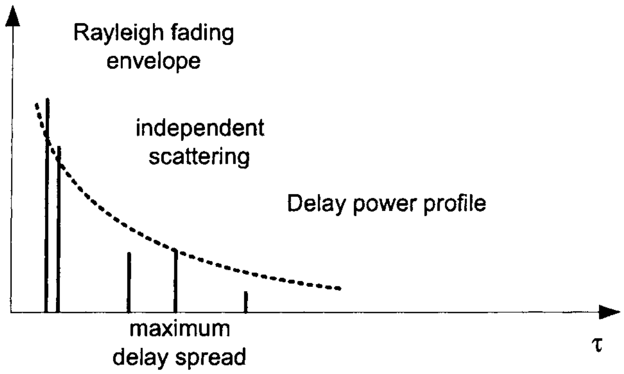

Figure 2.4: A time and frequency dispersive channel profile

where $\Delta f_{\max}$ is the maximum Doppler frequency. When the Doppler shift is comparable to the signal bandwidth (i.e., coherence time $T_c \sim$ the symbol period), the channel is termed *time selective* (fast fading) or *frequency dispersive.* On the other hand, if the Doppler shift is insignificant relative to the symbol rate (channel coherence time $\gg$ symbol period), the channel is termed time nonselective (slow fading).

2.1.4 Statistical characteristics of broadband channels

In reality, a wireless channel may be both time dispersive and frequency dispersive at the same time. Given its random nature, a system design must be based upon the statistical characteristics of wireless channels.

Mathematically, the time and frequency dispersive channel can be modeled as a linear time-variant (LTV) transfer function. A commonly used model for a broadband channel in Rayleigh fading assumes

- M significant solvable uncorrelated paths with normalized delays (by symbol duration T_{symbol}): $\tau_0, \tau_1, ..., \tau_{M-1}$ ($\tau_0 = 0$).
- the M path gains are complex Gaussian random variables $\{\alpha_0, \alpha_2, ..., \alpha_{M-1}\}$, having independent real and imaginary parts with zero mean and variance $\sigma_i^2/2$.

A profile of the LTV channel is depicted in Figure 2.4.

Assuming f_c is the carrier frequency, the time-variant channel impulse response is given by

$$h(\tau, t) = \sum_{m=0}^{M-1} \alpha_i e^{-j2\pi f_c \tau_m(t)} \delta(\tau - \tau_m(t))$$

Clearly, $h(\tau, t)$ is a complex-valued Gaussian random process in the t variable. Its envelope at any instant t is therefore Rayleigh-distributed. The time-varying $\tau_m(t)$ in $e^{-j2\pi f_c \tau_m(t)}$ captures the spectrum-smearing Doppler effect.

Applying the Fourier transform with respect to the delay variable τ, we obtain the *time-frequency channel response* of the time-variant channel:

$$h(f,t) = \sum_{m=0}^{M-1} \alpha_i e^{-j(\theta_m + 2\pi F_D t - 2\pi f \tau_m)} \tag{2.1}$$

Here, we re-express $2\pi f_c \tau_m(t)$ as $(2\pi F_D t + \theta_m)$ where F_D is the Doppler frequency and θ_m is a random phase attribute to the mth multipath.

In most analysis, the wide-sense stationary uncorrelated scattering (WSSUS) channel model is used [1]. The statistical characteristics are then described by its covariance matrix

$$R_h(\Delta f, \Delta t) = E\{h(f;t)h^*(f - \Delta f; t - \Delta t)\} \tag{2.2}$$

Several important profiles can be derived from the autocorrelation functions:

- *The frequency correlation function*: $p_h(\Delta f) = R_h(\Delta f, 0)$, quantifies the channel correlation in frequency domain. The nominal width of $p_h(\Delta f)$, termed *coherence bandwidth* B_C, is a statistical measure of the range of frequencies over which the channel can be considered flat.

- *The delay power profile*: $p_h(\tau) = \mathcal{F}^{-1}\{p_h(\Delta f)\}$, quantifies the time dispersive properties of the channel. The nominal width of $p_h(\tau)$ is known as the *multipath delay spread* $\tau_{\max}$. The rms delay spread τ_{rms} is defined as the square root of the second central moment of the delay power profile. In practice, we often use the following approximation

$$B_C = \frac{1}{5\tau_{rms}} \tag{2.3}$$

- *The time correlation function*: $p_h(\Delta t) = R_h(0, \Delta t)$ quantifies the time varying nature of the channel. Its Fourier transform is the *Doppler power spectrum* $\Phi_h(v) = \mathcal{F}\{p_h(\Delta t)\}$. The nominal width of $\Phi_h(v)$, termed the *Doppler spread* B_D, is defined as the range of frequencies over which the Doppler spectrum is essentially non-zero. The inverse of the Doppler spread,

$$T_C = 1/B_D \tag{2.4}$$

  is defined as the *channel coherence time,* which is a statistical measure of the time interval over which the channel response is essentially invariant.

To obtain Rayleigh fading with the Jakes' spectrum and an exponentially decaying power delay profile with RMS-value τ_{rms}, the following probability density functions are chosen for Eq (2.1):

$$p_\theta(\theta) = 1/2\pi$$

$$p_F(F_D) = \frac{1}{\pi F_{D,\max}\sqrt{1-(F_D/F_{D,\max})^2}}$$

$$p_\tau(\tau) = \frac{e^{-\tau/\tau_{rms}}}{\tau_{rms}\left(1-\tau_{\max}/\tau_{rms}\right)}$$

In summary,

- Envelope fading affects the signal strength and therefore fading margin in link budget calculation of the wireless system. Power control and spatial diversity techniques are among the most effective means to cope with envelope fading;
- Frequency-selective fading alters the signal waveform and therefore the detection performance. Traditionally, channel equalization is utilized to compensate the effect. Alternatively, one can overcome the frequency selectivity by transferring a broadband signal into parallel narrowband streams as shown in ensuing discussion;
- Time-selective fading smears the signal spectrum and introduces variation too fast for power control. Time interleaving and diversity techniques are most effective means of coping with time-selective fading.

2.2 Canonical form of broadband transmission

OFDM is a form of multicarrier modulation that transmits broadband data over parallel narrowband streams. The superiority of OFDM over single-carrier based modems can be better understood by answers to the following three fundamental questions:

1. *Q1: What kind of signal waveforms are immune to multipath effects?*
2. *Q2: How tight can we pack signals in a given bandwidth?*
3. *Q3: Does fast implementation exist for broadband modem?*

Q1: What kind of signal waveforms are immune to multipath?

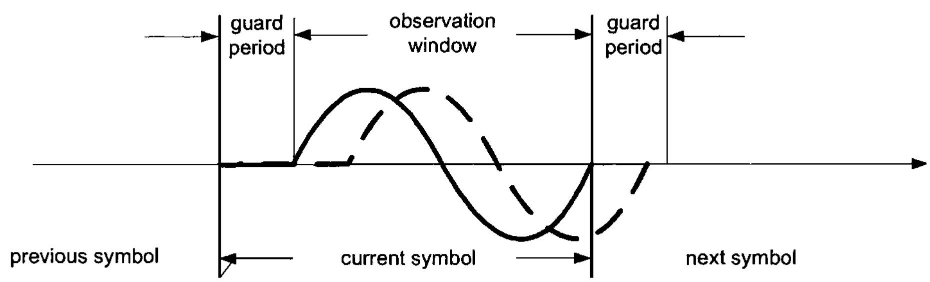

Figure 2.5: Zero-padding guard interval

Multipath induced intersymbol interference is traditionally handled with equalizers at the receiver side. On the other hand, if the signal waveform is designed to be immune (i.e., to a scalar ambiguity) to multipath distortions, the receipt can be dramatically simplified. Since the multipath channel can be perfectly modeled as a linear system, the right signal waveforms can be derived by examining the input-output relation of a linear operation:

$$y(t) = h(t) * x(t) = \int_{-\infty}^{\infty} h(\tau)x(t-\tau)d\tau$$

It is well known that with a complex exponential input, $x(t) = s \cdot w(t) = se^{j\omega t}$, the output signal of the linear channel is identical to the input (within a multiplicative constant):

$$\begin{aligned} y(t) &= h(t) * x(t) \\ &= s\int_{-\infty}^{\infty} h(\tau)e^{-j\omega(t-\tau)}d\tau \\ &= se^{-j\omega t}\int_{-\infty}^{\infty} h(\tau)e^{-j\omega\tau}d\tau = H(j\omega)se^{j\omega t} \end{aligned}$$

The above observation suggests that complex exponential signals can be used as waveforms for multipath channels. On the other hand, it is important to realize that the above relation is only valid for an infinitely long complex exponential. For practical digital communication, the signal waveforms must be confined to a symbol period.

Fortunately, if the channel response is FIR, i.e., $h(t) = 0, t \notin [\ 0 \quad \tau_{\max}\ ]$, the same input-output property holds within a finite observation window duration T. The idea is illustrated in Figures 2.5 and 2.6. In Figure 2.5, a zero-padded guard period is inserted between neighboring symbols to prevent intersymbol

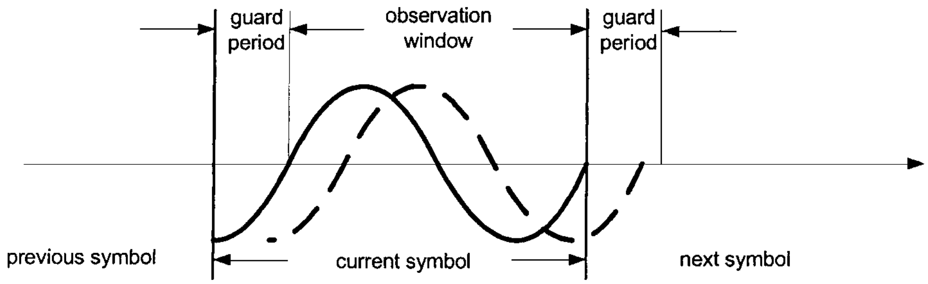

Figure 2.6: Cyclic-prefix guard interval

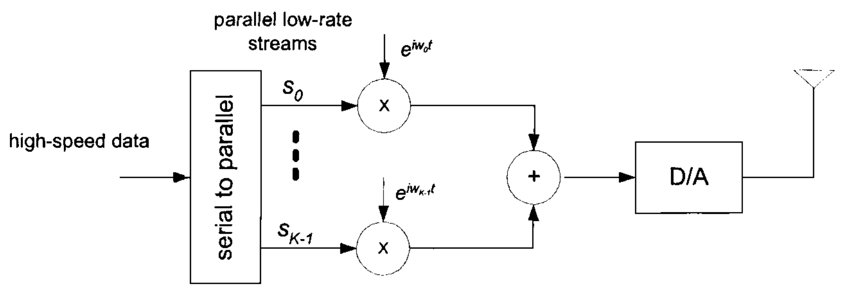

Figure 2.7: FDM scheme for high-speed transmission

interference (ISI). However the integrity of the waveform is not preserved[1]. Instead of using a silent period to guard against the ISI, a *cyclic-perfix (CP)* of duration at least $\tau_{\max}$ can be employed at the transmitter. As shown in Figure 2.6, any multipath components with delays less or equal to $\tau_{\max}$ will maintain their complex exponential waveforms within the observation window, leading to an intact signal waveform (within a scalar ambiguity) at the receiver side.

To carry more information in a given time window, the old-fashioned frequency-division multiplexing (FDM) can be utilized as shown in Figure 2.7. In particular, K information-bearing symbols, $s_0, \cdots, s_{K-1}$, can be modulated onto K different subchannels using different complex exponential $w_k(t) = e^{j2\pi kt/N}$as follows.

$$x(t) = \textstyle\sum_{k=0}^{K-1} s_k w_k(t), \quad t \in [\ -\tau_{\max},\ \ T\ ]\ .$$

[1]If the starting point of the observation windown is perfectly known, it is possible to restore the waveform by patching the distorted portion with samples within the guard period of the next symbol.

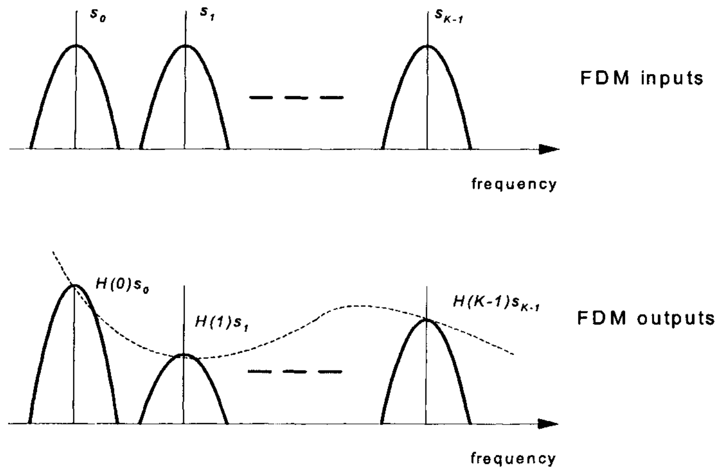

Figure 2.8: The input–output relation of FDM signals

With the CP, the output is only subject to a scalar multiplication on each of the information symbols within the observation window.

$$y(t) = \sum_{k=0}^{K-1} H(e^{j2\pi kt/N}) s_k w_k(t)$$

Graphically, the effect of the channel is a mere "scaling" on each subchannel as shown in Figure 2.8. Since the scalar ambiguity can be removed with channel estimation, it can be argued that the FDM is "immune" to the time-dispersion effect and thus has an advantage over single-carrier modulation with a linear equalizer.

Note however, such immunity is achieved at the expense of an unused CP (at the receiver). The ratio of $\tau_{\max}/T$ determines the signal overhead of the system. Since $\tau_{\max}$ is often fixed in a given application, there is an incentive to increase T in order to increase the system efficiency. The tradeoff though, is increased sensitivity to frequency offset - see Chapter 3.

Q2: How tight can we pack the signals in a given bandwidth?

While the CP in FDM reduces the multipath channel effect to a scalar on each subchannel, it is unclear whether the signals across different subchannels will interfere to each other. For two waveforms to be orthogonal within [0, T], they must fulfill the orthogonality constraint:

$$\int_0^T w_1(t) w_2^*(t) dt = 0$$

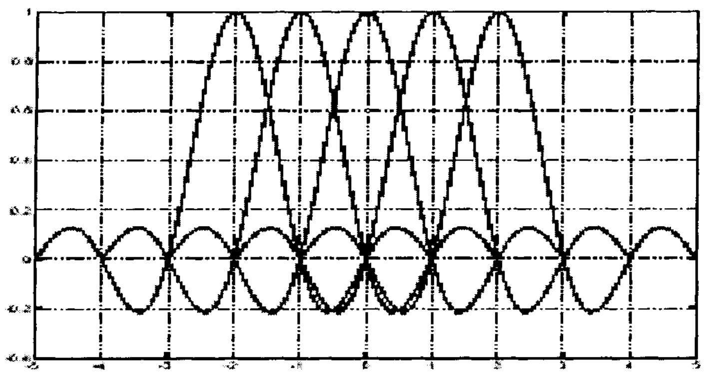

Figure 2.9: The spectra of OFDM signal

The minimum tone separation is thus $1/T$ hertz. Therefore, as long as $w_k(t)$ is placed $1/T$ hertz apart, there will be no inter-subchannel interference in FDM. The FDM that satisfies such frequency spacing requirement is therefore referred to as OFDM - orthogonal frequency division multiplexing. Each tone in OFDM is referred to as a *subcarrier*. The typical spectrum of OFDM signals is depicted in Figure 2.9. The overlapping *sinc* shaped spectra assure zero inter-subchannel interference at the right frequency sampling points.

Q3: *Does fast implementation exist for broadband modem?*

We now consider the digital implementation and the complexity issue of the OFDM modem.

Assume there are K (usually power of 2) subcarriers in the system, and the time-domain sampling rate is $N/T = 1/T_s$, $N = K$ hertz. Further assume that the channel delay spread is L ($LT_s = \tau_{\max}$) samples. Let x_n, h_n and y_n be the sampled input, channel, and output, respectively; it is clear that they are related by a linear convolution:

$$y_n = x_n * h_n$$

Now let $Y(k), X(k)$,and $H(k)$ be the K-point Discrete Fourier transforms (DFTs) of the output, input, and channel, respectively, within the observation window. In ordinary cases,

$$Y(k) \neq X(k)H(k), \qquad k = 0, ..., K-1$$

since the y_n is related to x_n and h_n by a linear convolution not a *circular* convolution. However by appending an L point CP, $x_{N-L}, x_{N-L+1}, \cdots, x_{N-1}$, to x_n, the circular convolutional effect is created within the N point time window (please prove this as an exercise).

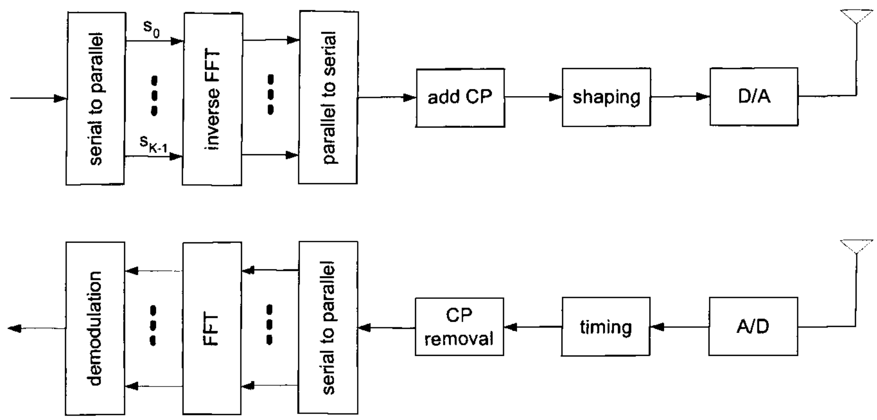

Figure 2.10: OFDM transceiver block diagram

In other words, in discrete frequency domain the channel effect is also a multiplicative constant:

$$Y(k) = X(k)H(k), \qquad k = 0, ..., K-1 \tag{2.5}$$

As a result, the OFDM modem can be elegently implemented in discrete-time using fast algorithms such as the Fast Fourier transform (FFT). Signals modulated on different subcarriers can be perfectly separated after the FFT operation at the receiver.

2.3 OFDM realization

Figure 2.10 depicts a typical transmitter and receiver chain of an OFDM modem. Unlike signal-carrier modulation, the OFDM modem is performed on a *block-by-block* basis. At the transmitter, a block of information-bearing symbols are first serial-to-parallel converted onto K *subcarriers*. The orthogonal waveform modulation is carried out using an inverse FFT and a parallel-to-serial converter. Following the converter, the last L points are appended to the beginning of the sequence as the cyclic prefix. The resulting samples are then shaped and transmitted. Each transmmited block is referred to as an *OFDM symbol.*

The receiver reverses the process using an FFT operation. In particular, the sampled signals are first processed to determine the starting point of a block and the proper demodulation window. By removing the CP (which now contains ISI), an N ($N = K$) point sequence is serial-to-parallel converted and fed to the FFT. The output of the FFT are the symbols modulated on the K subcarriers, each multiplied by a complex channel gain. Depending the availability of the

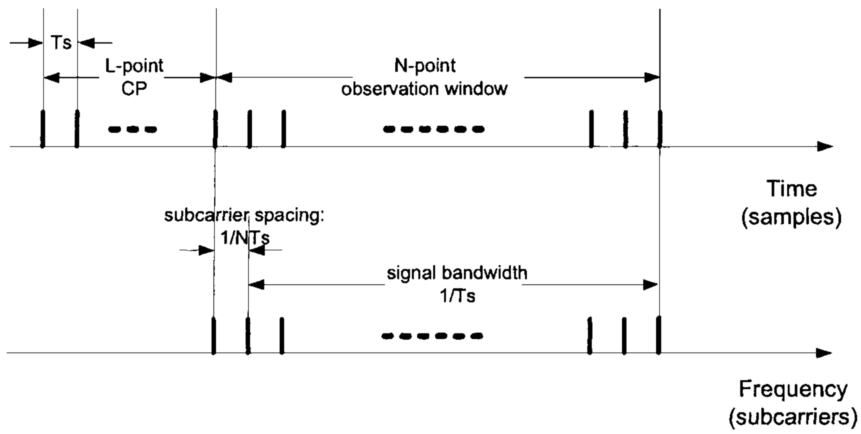

Figure 2.11: OFDM representation in time and frequency domains

channel information, different demodulation/decoding schemes are then used to recover the information bits.

Considering an OFDM system with N subcarriers and a time-domain sampling rate $1/T_s$, a time and frequency representation of an OFDM symbol is depicted in Figure 2.11.

Now let us examine the modem operations mathematically. Denote

$$\mathbf{s}(n) = [s_0(n)\ s_1(n) \cdots s_{P-1}(n)]^T$$

as the nth block of data to be transmitted. The number of subcarriers used, P, may be less or equal to the number of total available subcarriers, K: $P \leq K$. The OFDM modulation is implemented by applying an IDFT operator to the data stream $\mathbf{s}(n)$. Using matrix representation, the resulting N-point time domain signal is given by

$$\mathbf{x}(n) = [x_0(n)\ x_1(n) \cdots x_{N-1}(n)]^T = \mathbf{W}_P \mathbf{s}(n),$$

where $\mathbf{W}_P$ is an $N \times P$ submatrix of the IDFT matrix $\mathbf{W}$. The columns of $\mathbf{W}_P$ correspond to the subcarriers that are modulated with data.

For DFT-based OFDM, a cyclic prefix is added to the multiplexed output of the IDFT to form an OFDM symbol, before it is transmitted through a fading channel. Because of (2.5), the receiver output for the nth block within the demodulation window is given by

$$
\begin{aligned}
\mathbf{y}(n) &= [y_0(n)\ y_1(n) \cdots y_{N-1}(n)]^T && (2.6)\\
&= \mathbf{W}_P\mathbf{H}\ \mathrm{s}(n) && (2.7)\\
&= \mathbf{W}_P \begin{bmatrix} H(1) & & 0 \\ & \ddots & \\ 0 & & H(P) \end{bmatrix} \mathrm{s}(n), && (2.8)
\end{aligned}
$$

where $H(i),\ i = 1 \cdots, N$ is DFT of the channel response. In other words, each subchannel, with a scalar ambiguity, can be recovered by applying a DFT to $\mathbf{y}(k)$:

$$\mathbf{W}_P^H \mathbf{y}(n) = \mathbf{H}\ [s_1(n) \cdots s_P(n)]^T \tag{2.9}$$

In practice, several additional operations are often needed at the transmitter and the receivers:

- *Cyclic prefix and postfix:* the cyclic prefix provides a guard interval for all multipath following the first arrival signal. As a result the timing requirement of the observation window is quite relaxed (up to $\tau_{\max}$ ambiguity). On the other hand, timing estimation often hinges on the multipath signal with the highest strength, which in some cases may not be the first arrival signal. To increase the robustness of the receiver, the guard interval is often split into cyclic pre-fix and post-fix as in Figure 2.12 to guard against early and late (relative to the strongest path) multipath signals.

- *Pulse shaping:* since the time-limited signal waveforms have strong raised-cosine sidelobes in the frequency-domain, OFDM has been shown to be sensitive to frequency offset which leads to inter-carrier interference. The rectangular time window also leads to high out-of-band emission (Figure 2.9) which is undesirable in radio communications. An effective way to reduce the ICI sensitivity is to pulse shape (time domain multiplication) the OFDM symbol with a pulse-shaping window. The tradeoff is a reduced guard-inteval and increased complexity. Alternatively, one can apply filters to limit the spectrum of the OFDM signals. Notice that filtering introduces the same convolutional effects as the multipath channel, therefore reduces system tolerance to delay spread given a CP.

- *Virtual carriers:* In order to guard against neightboring band interference and the out-of-band emission, a portion of the subcarriers at the two edges of the band may not be modulated. These unused subcarriers are termed as the *virtual carriers.* The concept is illustrated in Figure 2.13. As a result, the number of subchannels that carry the information is generally smaller than the size of the DFT block, *i.e.*, $P < N = K$. The virtual carriers provide a guard band for neighboring channels. In the absence of

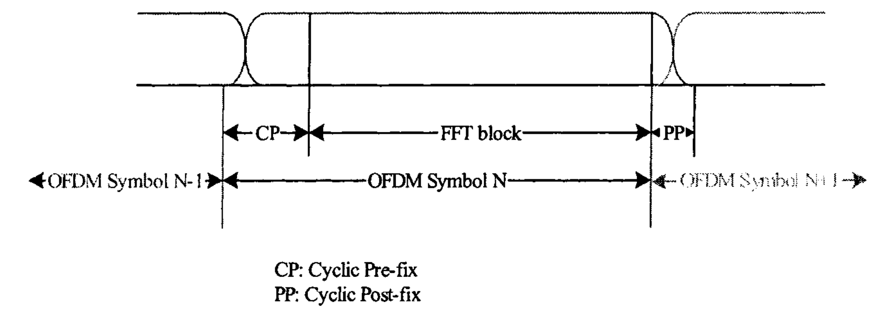

Figure 2.12: OFDM symbols with cyclic pre-fix and post-fix

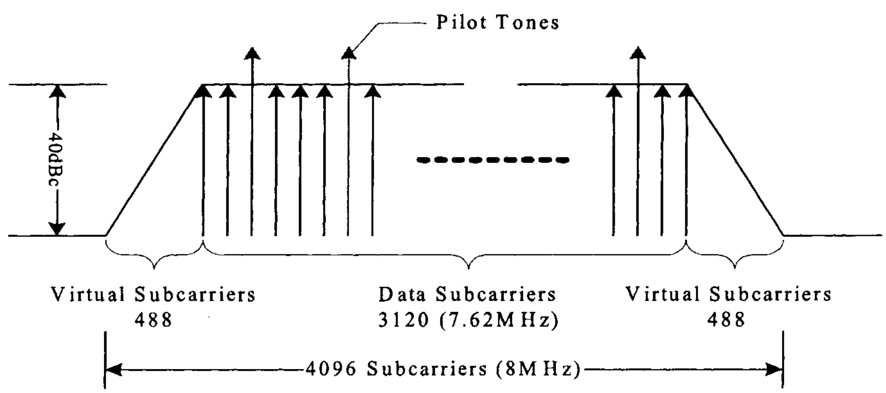

Figure 2.13: OFDM with virtual carriers

adjacent-channel-interference (ACI), the outputs from these virtual carriers are zero. For simplicity, we shall use the logic indices 1 to P to denote data subcarriers and $P+1$ to N to denote virtual carriers.

Example 1 *OFDM parameter selection in IEEE 802.11a: RF measurements have shown that in a typical LAN environment, the rms delay spread at 5GHz is on the order of 50ns, with maximum delay up to 500ns.*

The cyclic guard interval in 802.11a is conservatively set at 800 ns (out of each 4μs OFDM symbol interval). From (2.3), the coherence bandwith is on the order of 4MHz. The subcarrier spacing in 802.11a is chosen to be 0.3125 Mhz, which is far less than the coherence bandwidth. This setting guarantees flat fading on individual subcarriers.

Out of a size 64-FFT, 12 are virtual carriers while the rest 52 subcarriers are dedicated to data (48) and pilots (4).

Example 2 *The DVB-T parameters: Depending on the region, the DVB-T bandwidth can be 6, 7, or 8Mhz. The length of the cyclic guard interval must be selected to accommodate "single-frequency-network" (SFN), where neighboring TV towers deliver the same OFDM modulated signals. The SNF creates artificial multipath with delay spread dictated by the maximum permissible distance between neighboring transmitters (e.g., 60 km). Accordingly, the DVB-T supports two modes of operations: the 2K-FFT and the 8K-FFT, with six possible values for guard interval from 224 us to 7 us. SNF with spacing up to 67 km can be realized with the longest CP (*224 *us* $\times 300,000$ *km/s* ≈ 67 *km).*

The 8K-FFT mode has subcarrier spacing 1.25 KHz, and a significantly longer symbol duration (e.g., 1 ms) than that in the 2K-FFT mode. In general, the 8K-FFT mode can only support fixed applications whereas the 2K-FFT mode is more suitable for portable and possibly mobile services.

2.4 Summary

The salient features of OFDM are highlighted as follows:

- Advantages
  - High spectral efficiency: OFDM is a highly efficienct modulation scheme which has been shown to approach the information threotical capacity with water-filling across its subcarriers. Although subcarrier-based power loading is less feasible in practice, adaptive coded modulation on OFDM subchannels (each subchannel comprises of a group of subcarriers) has already been adopted into IEEE standards.
  - Simple implementation: the use of FFT and IFFT in OFDM reduces the modem complexity, especially at the receiver. With the FFT, the number of operations in each OFDM symbol is on the order of $N \log N$. The implementation complexity of single carrier system with an equalizer is at least NL_e, where L_e is the number of taps in the equalizer.
  - Resistance to fading and interference: OFDM is robust against frequency selective fading and interference. With channel state information, maximum likelyhood detection can be effectively implemented for any given channel profile or interference pattern.
- Disadvantages:
  - Sensitivity to frequency offset: OFDM has strong tolerance against timing offset because of the cyclic guard interval. On the other hand, its tightly packed subcarriers give rise to increased sensitivity with respect to carrier frequency errors (oscillator impairments etc.) and *interchannel interference* (ICI). The extend of performance degradation due to carrier frequency offset is a function of the subcarrier spacing, the size of the FFT, and the modulation schemes used.

- High peak-to-average power ratio: since the OFDM signal is the superposition of low rate streams modulated at different frequencies, its time-domain dynamic range increases with the number of subcarriers. The high peak-to-average power ratio (PAPR) imposes stringent requirements on the A/Ds and D/As, and more importantly, on the linearity of the power amplifier (PA).

In this chapter, we describe the challenges in broadband communications by characterizing the wireless fading channels. The statistical model of time- and frequency selective fading channels is presented. Based on a multi-carrier communication framework, we introduce the concept of OFDM and explain how channel fading can be mitigated with OFDM modulation. The canonical block diagram of an OFDM modem is described, along with implementation details, such as cyclic pre- and postfix, virtual carriers, and pulse shaping essential to a practical OFDM realization.