Nadar Saraswathi College of Engineering & Technology

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DEPARTMENT OF ELECTRONICS AND COMMUNICATION ENGINEERING

III YEAR / VI SEMESTER

EC8652– WIRELESS COMMUNICATION

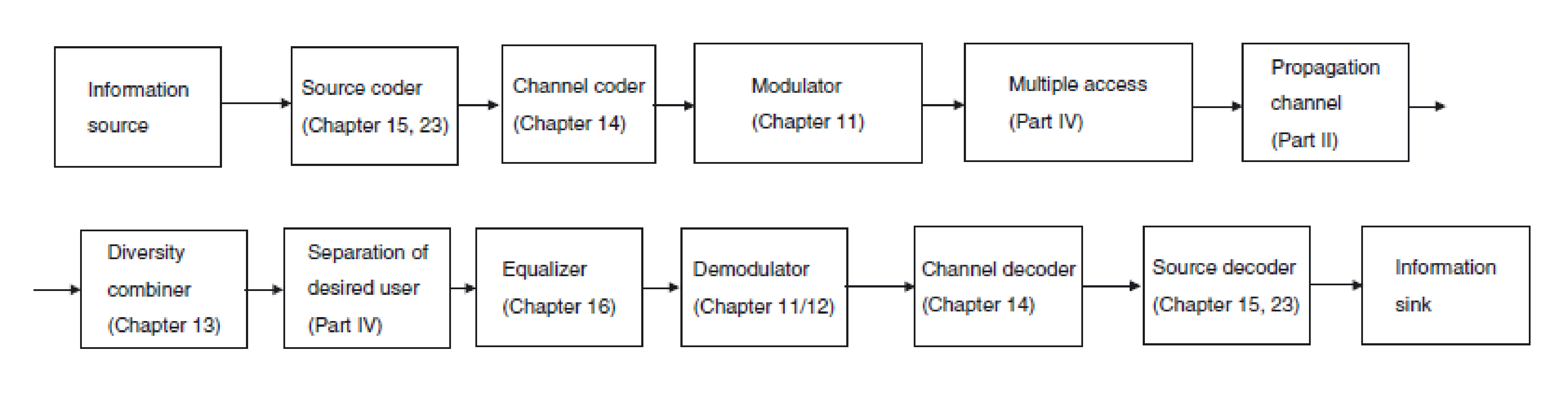
Dr.R.ATHILINGAM
Associate professor & Head (i/c)
Nadar Saraswathi College of & Technology,
Vadapudupatti, Annanji (po), Theni – 625531.

DIGITAL SIGNALING - 1

UNIT 03 – DIGITAL SIGNALING FOR FADING CHANNELS– LECTURE 01

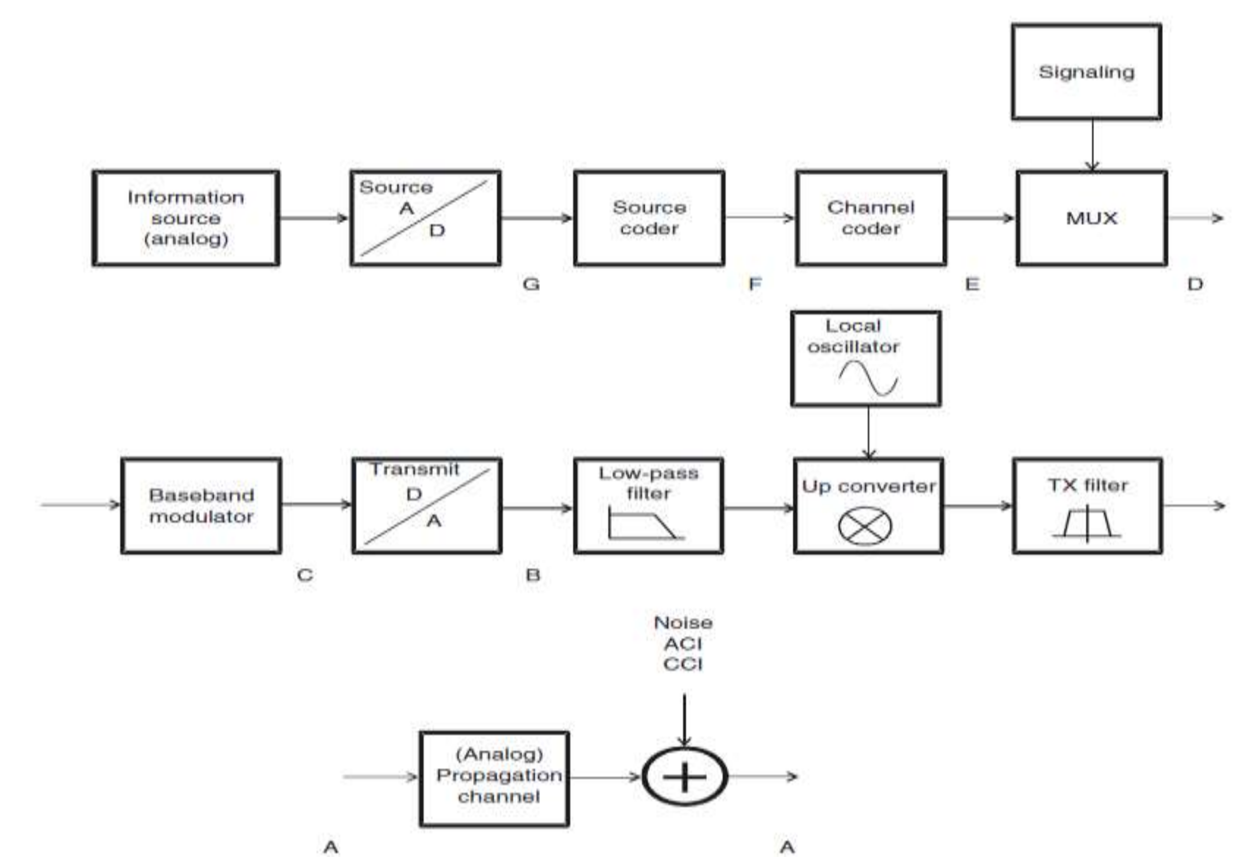
Structure of a Wireless Communication Link

- The goal of a wireless link is the transmission of information from an analog information source (microphone, video camera) via an analog wireless propagation channel to an analog information sink (loudspeaker, TV screen); the digitizing of information is done only in order to increase the reliability of the link.
- In order to make it more resistant to errors introduced by the channel (note that such encoding is done for most, but not all, wireless systems). The encoded data are then used as input to a modulator, which maps the data to output waveforms that can be transmitted.
- By transmitting these symbols on specific frequencies or at specific times, different users can be distinguished.
- The signal is then sent through the propagation channel, which attenuates and distorts it, and adds noise.
- The different users are separated.If the channel is delay dispersive, then an equalizer can be used to reverse that dispersion, and eliminate intersymbol interference.
- Afterwards, the signal is demodulated, and a channel decoder eliminates (most of) the errors that are present in the resulting bit stream.


Information source
Source coder (Chapter 15, 23)
Channel coder (Chapter 14)
Modulator (Chapter 11)
Multiple access (Part IV)
Propagation channel (Part II)
Diversity combiner (Chapter 13)
Separation of desired user (Part IV)
Equalizer (Chapter 16)
Demodulator (Chapter 11/12)
Channel decoder (Chapter 14)
Source decoder (Chapter 15, 23)
Information sink

- The *information source provides an analog source signal and feeds it into the source ADC* (Analog to Digital Converter).
- This ADC first band limits the signal from the analog information source (if necessary), and then converts the signal into a stream of digital data at a certain sampling rate and resolution (number of bits per sample).
- The *source coder uses a priori information on the properties of the source data in order to* reduce redundancy in the source signal. This reduces the amount of source data to be transmitted, and thus the required transmission time and/or bandwidth.
- The *channel coder adds redundancy in order to protect data against transmission errors. This* increases the data rate that has to be transmitted at interface. Data can be sorted according to importance; more important bits then get stronger protection.
- Furthermore, it is possible to use interleaving to break up error bursts; note that interleaving is mainly effective if it is combined with channel coding.
- *Signaling adds control information for the establishing and ending of connections, for associating* information with the correct users, synchronization, etc. Signaling information is usually strongly protected by error correction codes.
- The *multiplexer combines user data and signaling information, and combines the data from multiple* users.

- The *baseband modulator assigns the gross data bits (user data and signaling at interface D) to* complex transmit symbols in the baseband.
- Oversampling and quantization determine the aliasing and quantization noise. Therefore, high resolution is desirable, and the data rate at the output of the baseband modulator should be much higher than at the input.
- The *analog low-pass filter in the TX eliminates the (inevitable) spectral components outside the* desired transmission bandwidth.
- The *TX Local Oscillator (LO) provides an unmodulated sinusoidal signal, corresponding to* one of the admissible center frequencies of the considered system.
- The *upconverter converts the analog, filtered baseband signal to a passband signal by mixing* it with the LO signal.
- The *RF TX filter eliminates out-of-band emissions in the RF domain. Even if the low-pass* filter succeeded in eliminating all out-of-band emissions, upconversion can lead to the creation of additional out-of-band components.



Signaling
Information source (analog)
Source A D
Source coder
Channel coder
MUX
G
F
E
D
Local oscillator
Baseband modulator
Transmit D A
Low-pass filter
Up converter
TX filter
C
B
Noise
ACI
CCI
(Analog) Propagation channel
A
A

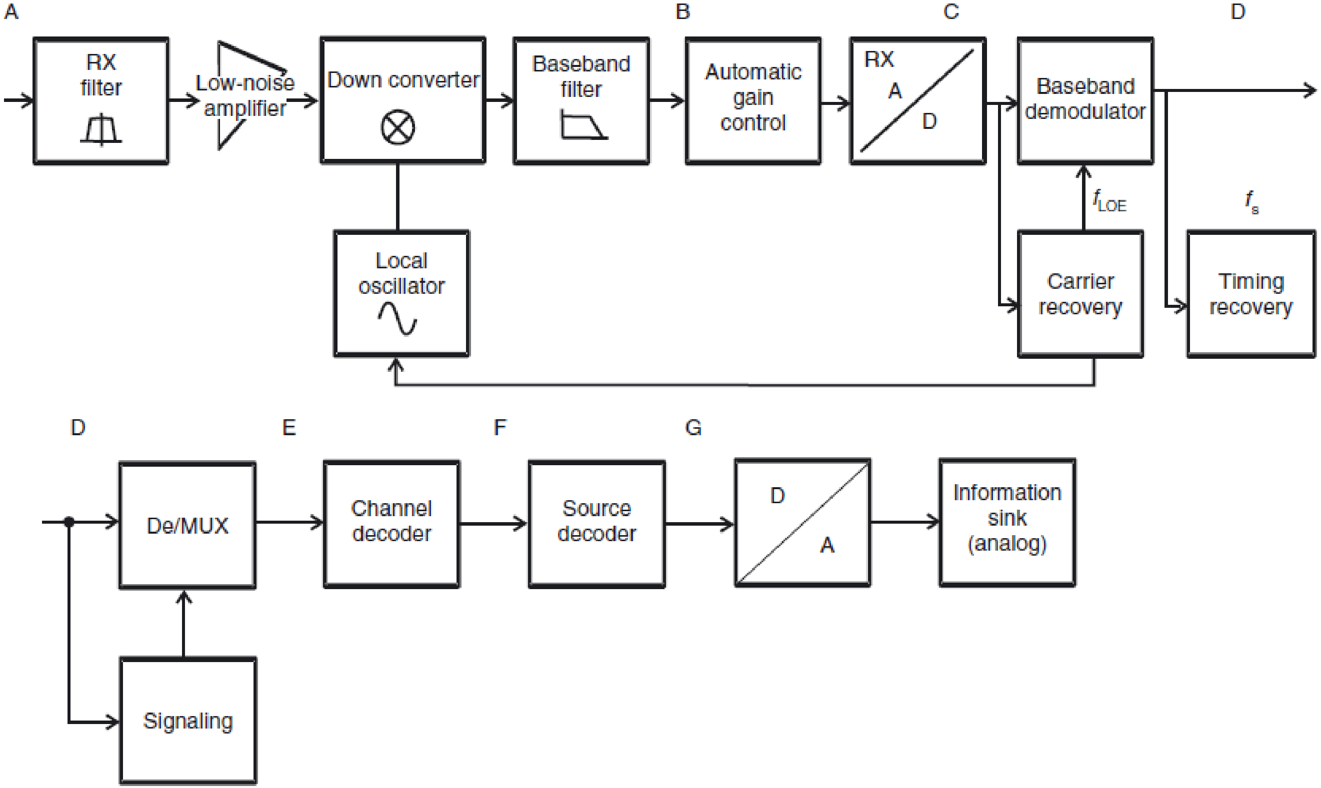


Figure 10.3 Block diagram of a digital receiver chain for mobile communications. MUX, multiplexing.

- The (analog) propagation channel attenuates the signal, and leads to delay and frequency dispersion.
- The RX filter performs a rough selection of the received band. The bandwidth of the filter corresponds to the total bandwidth assigned to a specific service, and can thus cover multiple communications channels belonging to the same service.
- The low-noise amplifier amplifies the signal, so that the noise added by later components of the RX chain has less effect on the Signal-to-Noise Ratio (SNR).
- The RX LO provides sinusoidal signals corresponding to possible signals at the TX LO. The frequency of the LO can be fine-tuned by a carrier recovery algorithm to make sure that the LOs at the TX and the RX produce oscillations with the same frequency and phase.
- The RX down converter converts the received signal into baseband. In baseband, the signal is thus available as a complex analog signal.
- The RX low-pass filter provides a selection of desired frequency bands for one specific user. It eliminates adjacent channel interference as well as noise. The filter should influence the desired signal as little as possible.
- The Automatic Gain Control (AGC) amplifies the signal such that its level is well adjusted to the quantization at the subsequent ADC.
- The RX ADC converts the analog signal into values that are discrete in time and amplitude. The required resolution of the ADC is determined essentially by the dynamics of the subsequent signal processing.

• Carrier recovery determines the frequency and phase of the carrier of the received signal, and uses it to adjust the RX LO.

• The baseband demodulator obtains soft-decision data from digitized baseband data, and hands them over to the decoder. The baseband demodulator can be an optimum, coherent demodulator, or a simpler differential or incoherent demodulator. This stage can also include further signal processing like equalization.

•Symbol-timing recovery uses demodulated data to determine an estimate of the duration of symbols, and uses it to fine-tune sampling intervals.

• The decoder uses soft estimates from the demodulator to find the original (digital) source data. In the most simple case of an uncoded system, the decoder is just a hard-decision (threshold) device.

• Signaling recovery identifies the parts of the data that represent signaling information and controls the subsequent demultiplexer.

The demultiplexer separates the user data and signaling information and reverses possible time compression of the TX multiplexer. Note that the demultiplexer can also be placed earlier in the transmission scheme; its optimum placement depends on the specific multiplexing and multiaccess scheme.

• The source decoder reconstructs the source signal from the rules of source coding. If the source data are digital, the output signal is transferred to the data sink. Otherwise, the data are transferred to the DAC, which converts the transmitted information into an analog signal, and hands it over to the information sink.

Quadrature-Phase Shift Keying

- A Quadrature-Phase Shift Keying (QPSK)-modulated signal is a PAM where the signal carries 1 bit per symbol interval on both the in- phase and quadrature-phase component. The original data stream is split into two streams, *b1i and b2i*

$$\left.\begin{aligned} b1_i &= b_{2i} \\ b2_i &= b_{2i+1} \end{aligned}\right\}$$

$$R_S = 1/T_S = R_B/2 = 1/(2T_B)$$

Let us first consider the situation where basis pulses are rectangular pulses

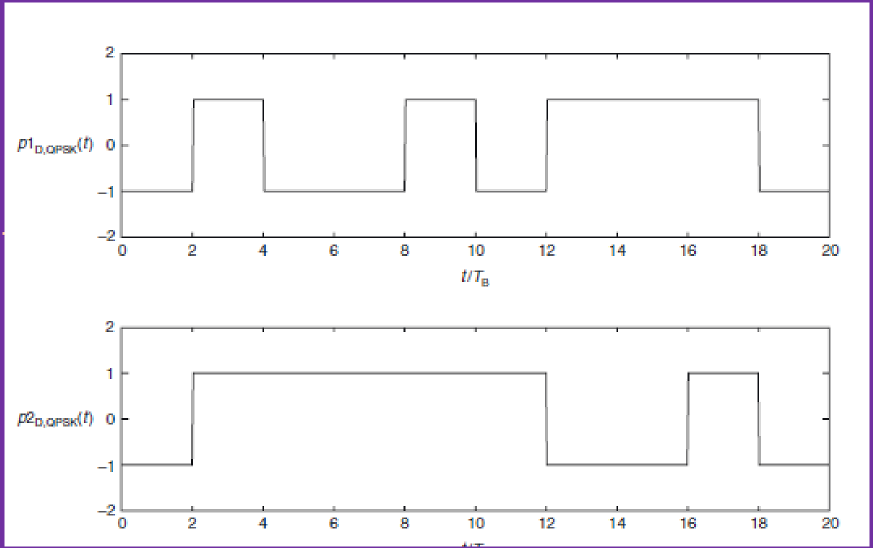
$$\left.\begin{aligned} p1_D(t) &= \sum_{i=-\infty}^{\infty} b1_i g(t - iT_S) = b1_i * g(t) \\ p2_D(t) &= \sum_{i=-\infty}^{\infty} b2_i g(t - iT_S) = b2_i * g(t) \end{aligned}\right\}$$

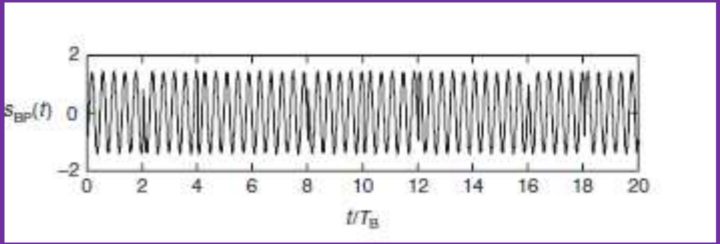
When interpreting QPSK as a PAM, the bandpass signal

$$s_{BP}(t) = \sqrt{E_B/T_B}[p1_D(t)\cos(2\pi f_c t) - p2_D(t)\sin(2\pi f_c t)]$$

The baseband signal is

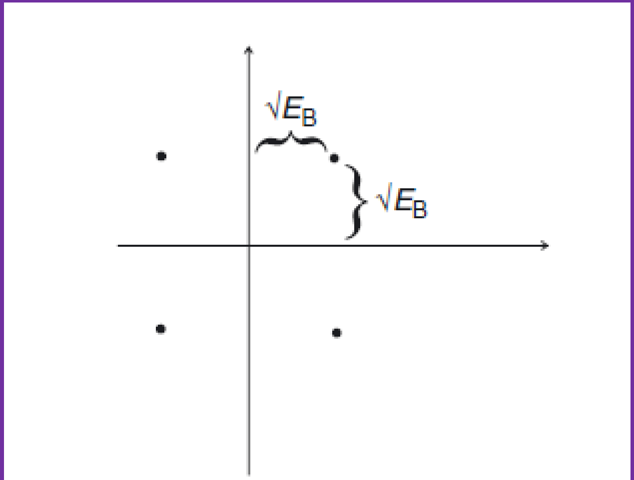
$$s_{LP}(t) = [p1_D(t) + jp2_D(t)]\sqrt{E_B/T_B}$$





the low-pass signal can be written as

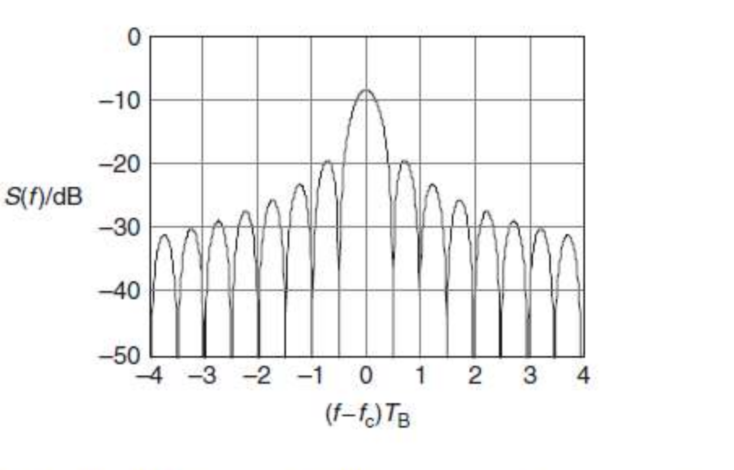
$$\Phi_S(t) = \pi \cdot \left[\frac{1}{2} \cdot p2_D(t) - \frac{1}{4} \cdot p1_D(t) \cdot p2_D(t)\right]$$



- The spectral efficiency of QPSK is twice the efficiency of BPSK, since both the in-phase and the quadrature-phase components are exploited for the transmission of information.
- This means that when considering the 90% energy bandwidth, the efficiency is 1.1 bit/s/Hz, while for the 99% energy bandwidth, it is 0.1 bit/s/Hz

π/4-Differential Quadrature-Phase Shift Keying

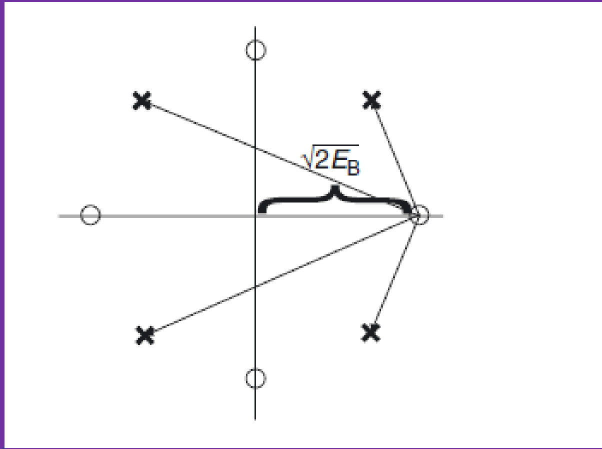
- Even though QPSK is nominally a constant envelope format, it has amplitude dips at bit transitions; this can also be seen by the fact that the trajectories in the I–Q diagram pass through the origin for some of the bit transitions.
- The duration of the dips is longer when non-rectangular basis pulses are used. Such variations of the signal envelope are undesirable, because they make the design of suitable amplifiers more difficult. One possibility for reducing these problems lies in the use of *π/4-DQPSK (π/4 differential quadrature-phase shift keying).*
- The principle of *π/4-DQPSK can be understood from the signal space diagram of DQPSK*.
- There exist *two sets of signal constellations: (0, 90, 180, 270◦) and (45, 135, 225,* 315◦).
- All symbols with an even temporal index *i are chosen from the first set, while all symbols* with odd index are chosen from the second set.
- In other words: whenever *t is an integer multiple of* the symbol duration, the transmit phase is increased by *π/4, in addition to the change of phase due* to the transmit symbol. Therefore, transitions between subsequent signal constellations can never pass through the origin in physical terms, this means smaller fluctuations of the envelope.

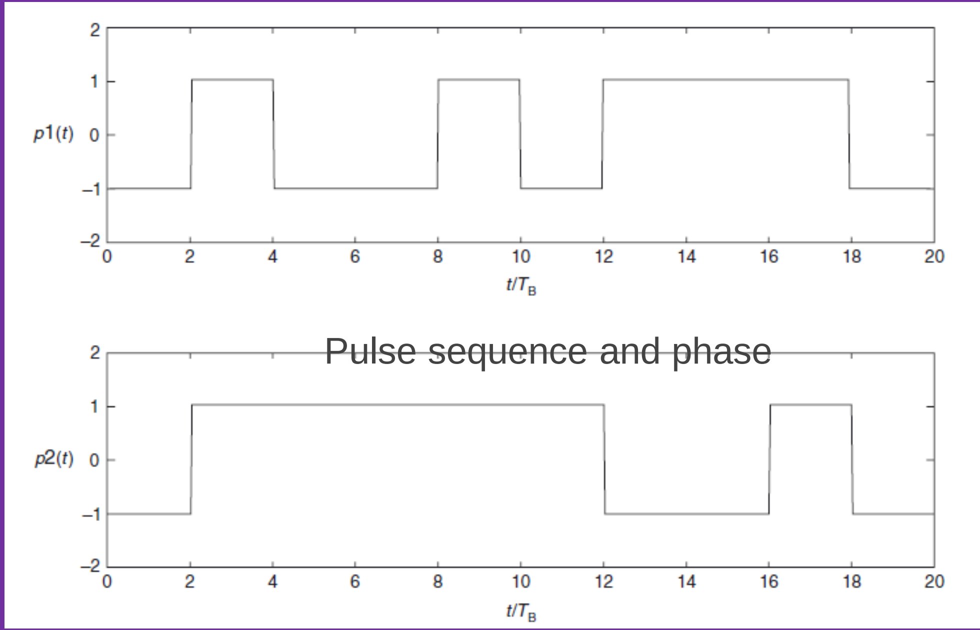


Normalized power-spectral density of quadrature-phase shift keying.

The signal phase is given by

$$\Phi_{\mathrm{S}}(t) = \pi\left[\frac{1}{2}p2_{\mathrm{D}}(t) - \frac{1}{4}p1_{\mathrm{D}}(t)p2_{\mathrm{D}}(t) + \frac{1}{4}\left\lfloor\frac{t}{T_{\mathrm{S}}}\right\rfloor\right]$$



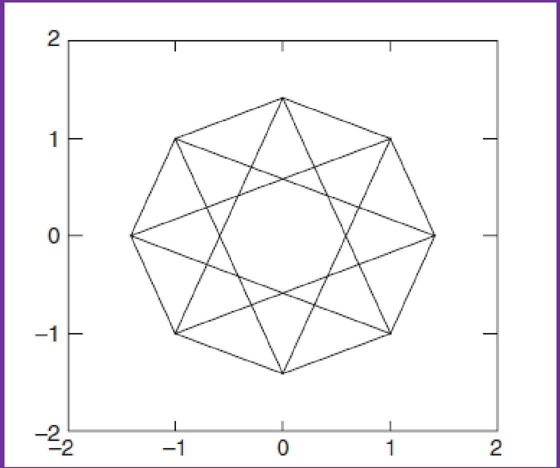


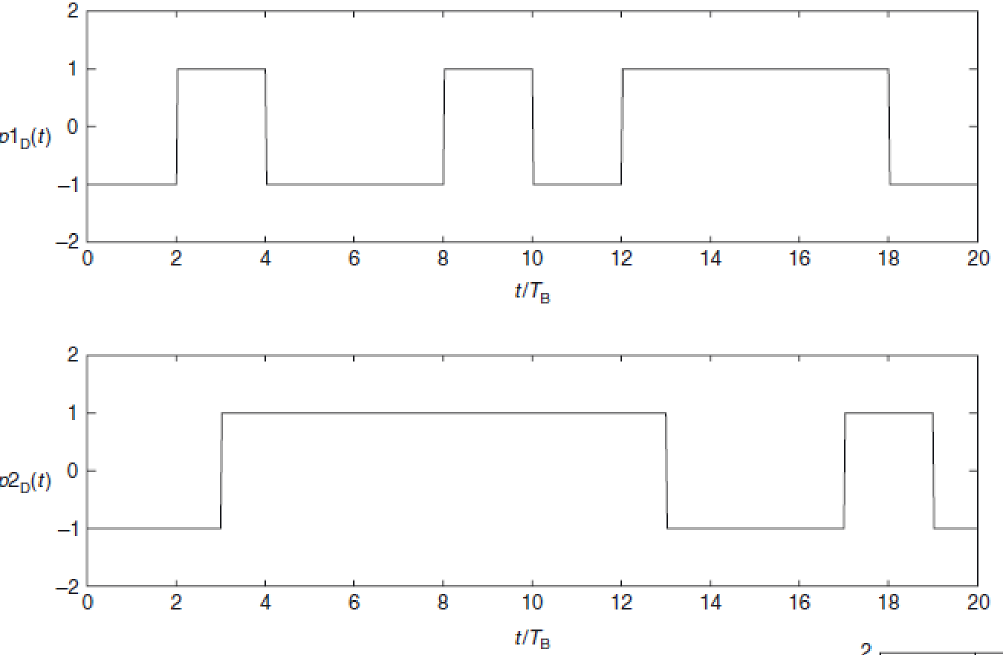
Pulse sequence and phase

Offset Quadrature-Phase Shift Keying

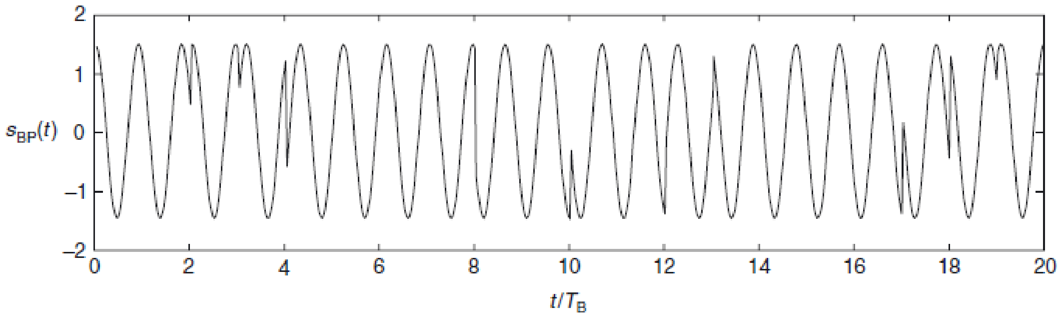
- Another way of improving the peak-to-average ratio in QPSK is to make sure that bit transitions for the in-phase and the quadrature-phase components occur at different time instants.
- This method is called OQPSK (offset QPSK). The bitstreams modulating the in-phase and quadrature-phase components are offset half a symbol duration with respect to each other so that transitions for the in-phase component occur at integer multiples of the symbol duration (even integer multiples of the bit duration), while quadrature component transitions occur half a symbol duration (1-bit duration) later. Thus, the transmit pulse streams are

$$\left.\begin{aligned} p1_{\mathrm{D}}(t) &= \sum_{i=-\infty}^{\infty} b1_i g(t - iT_{\mathrm{S}}) = b1_i * g(t) \\ p2_{\mathrm{D}}(t) &= \sum_{i=-\infty}^{\infty} b2_i g\left(t - \left(i + \tfrac{1}{2}\right) T_{\mathrm{S}}\right) = b2_i * g\left(t - \frac{T_{\mathrm{S}}}{2}\right) \end{aligned}\right\}$$





Pulse sequence and phase



Minimum Shift Keying

Minimum Shift Keying (MSK) is one of the most important modulation formats for wireless communications. However, it can be interpreted in different ways

The first interpretation is as CPFSK with a modulation index

$$h_{\mathrm{mod}} = 0.5, \quad f_{\mathrm{mod}} = 1/4T$$

Alternatively, we can interpret MSK as *Offset QAM (OQAM) with basis pulses that are sinusoidal* half-waves extending over a duration of 2*TB*

$$g(t) = \sin(2\pi f_{\mathrm{mod}}(t + T_{\mathrm{B}}))g_{\mathrm{R}}(t, 2T_{\mathrm{B}})$$

Due to the use of smoother basis functions, the spectrum decreases faster than that of “regular” OQPSK:

$$S(f) = \frac{16T_{\mathrm{B}}}{\pi^2}\left(\frac{\cos(2\pi f T_{\mathrm{B}})}{1 - 16f^2T_{\mathrm{B}}^2}\right)^2$$

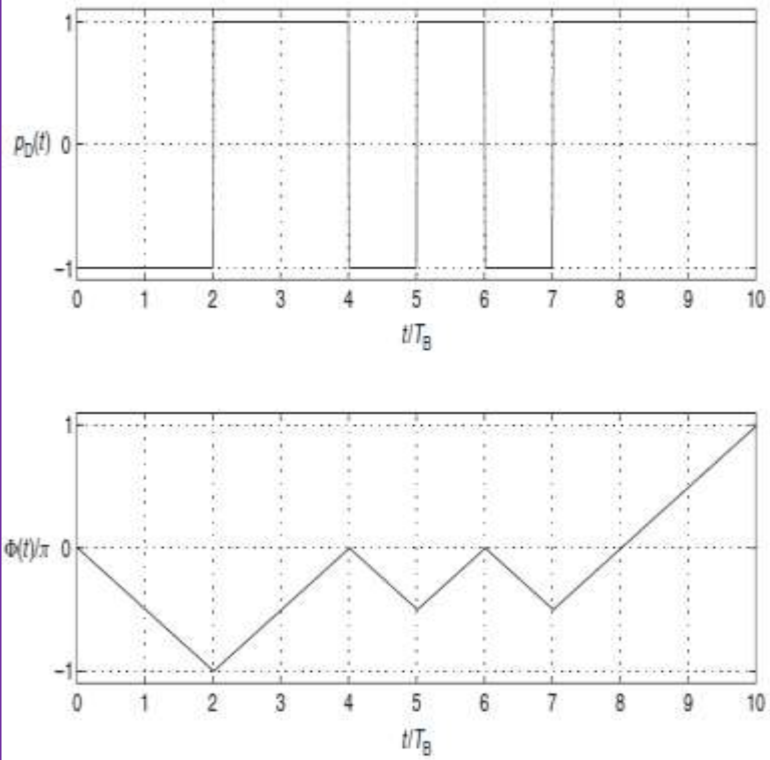
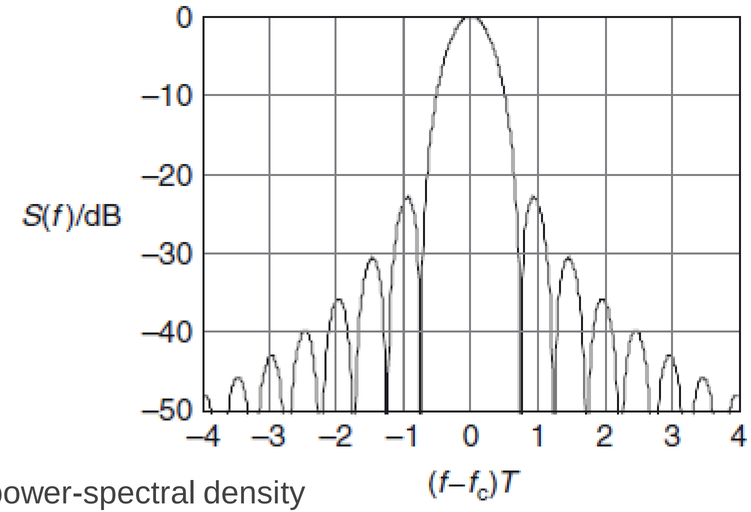
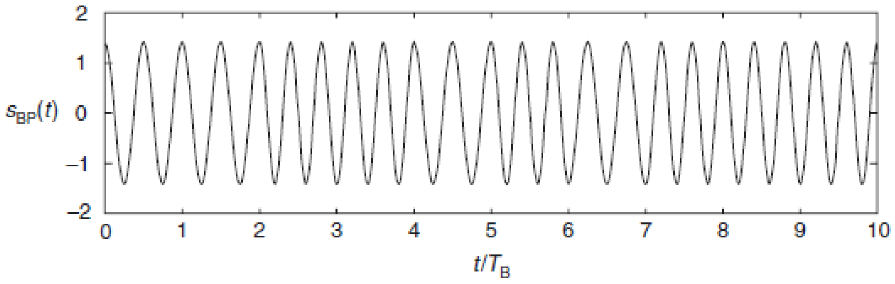


Figure 11.32 Phase pulse and phase as function of time for minimum shift keying signal.



power-spectral density



Gaussian Minimum Shift Keying

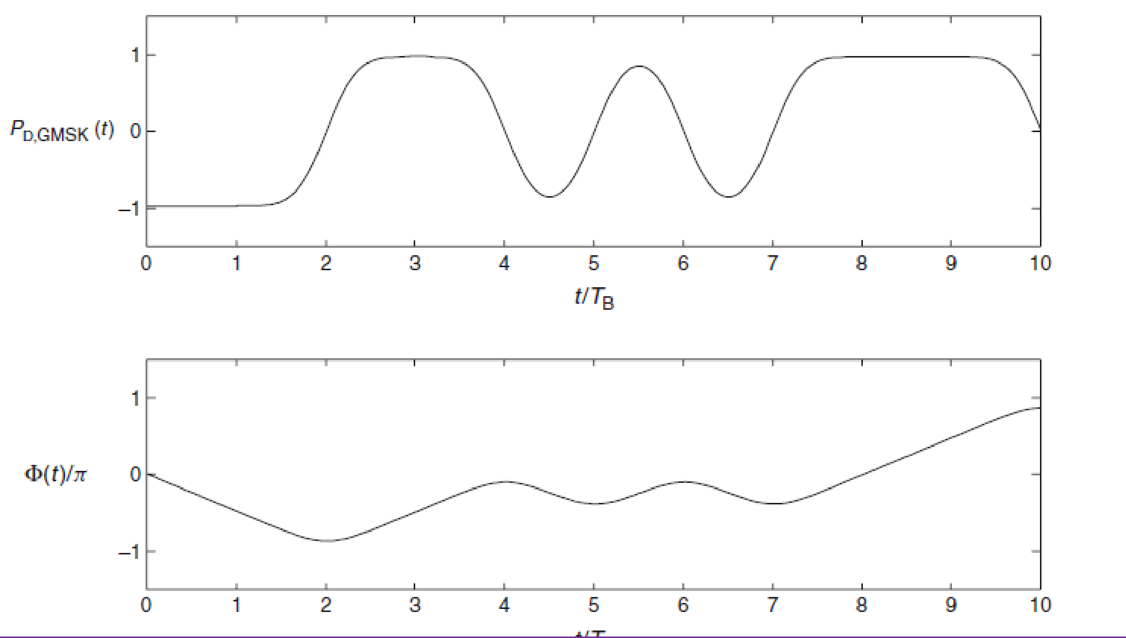
GMSK (Gaussian MSK) is CPFSK with modulation index hmod = 0.5 and Gaussian phase basis pulses:

$$\tilde{g}(t) = g_G(t, T_B, B_G T)$$

Thus the sequence of transmit phase pulses is

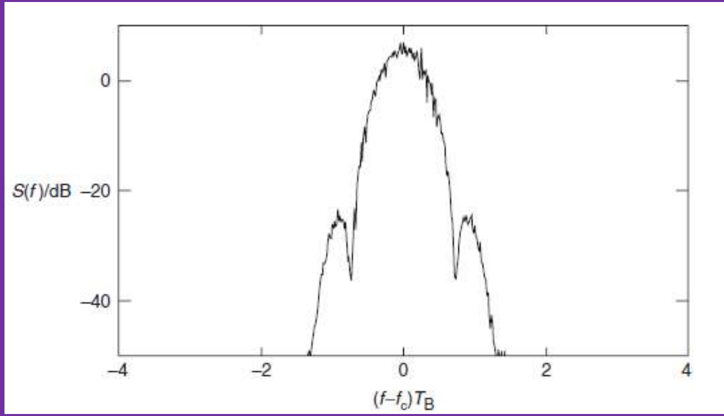
$$p_D(t) = \sum_{i=-\infty}^{\infty} b_i \tilde{g}(t - iT_B) = b_i * \tilde{g}(t)$$

GMSK is the modulation format most widely used in Europe. It is applied in the cellular Global System for Mobile communications (GSM) standard (with *BGT = 0.3) and the cordless standard* Digital Enhanced Cordless Telecommunications



Pulse sequence and phase of Gaussian minimum shift keying signal

Gaussian minimum shift keying power-spectral density



DIGITAL SIGNALING - 2

UNIT 03 – DIGITAL SIGNALING FOR FADING CHANNELS– LECTURE 02

Window Parameters

- *Window parameters* precisely calculate the *interference quotient QT and the delay window WQ. They are a measure for the* percentage of energy of the average PDP arriving within a certain delay interval.
- In contrast to the delay spread and coherence bandwidth, the window parameters need to be defined in the context of specific systems. The *interference quotient QT is the ratio between the signal power arriving within a time window* of duration *T* , *relative to the power arriving outside that window. The delay window characterizes* the self-interference due to delay dispersion.
-

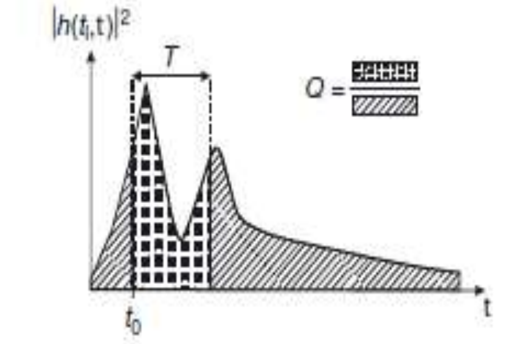


Figure 6.10 Definition of window parameters.
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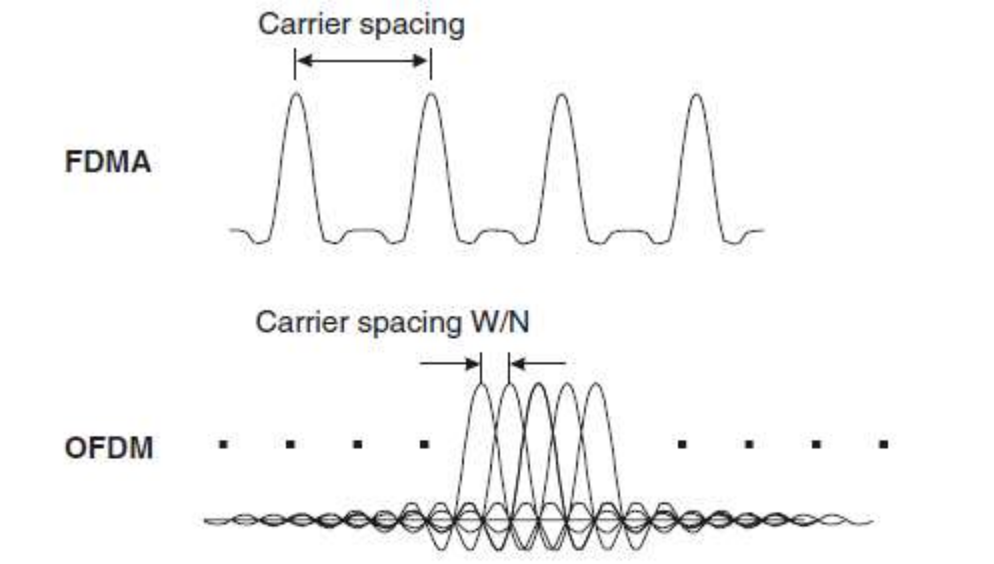
Orthogonal Frequency Division Multiplexing (OFDM)

- Orthogonal Frequency Division Multiplexing (OFDM) is a modulation scheme that is especially suited for high-data-rate transmission in delay-dispersive environments.
- It converts a high-rate data stream into a number of low-rate streams that are transmitted over parallel, narrowband channels that can be easily equalized.
- OFDM splits a high-rate data stream into N parallel streams, which are then transmitted by modulating N distinct carriers (henceforth called subcarriers or tones).
- Symbol duration on each subcarrier thus becomes larger by a factor of N. In order for the receiver to be able to separate signals carried by different subcarriers, they have to be orthogonal.
- Conventional Frequency Division Multiple Access (FDMA), can achieve this by having large (frequency) spacing between carriers.

- Due to the rectangular shape of pulses in the time domain, the spectrum of each modulated carrier has a sin*(x)/x shape.*
- *The spectra of* different modulated carriers overlap, but each carrier is in the spectral nulls of all other carriers.
- Therefore, as long as the receiver does the appropriate demodulation (multiplying by exp$(-j2\pi f n t)$ and integrating over symbol duration), the data streams of any two subcarriers will not interfere.

$$\int_{iT_S}^{(i+1)T_S} \exp(j2\pi f_k t)\exp(-j2\pi f_n t)\,dt = \delta_{nk}$$



Cyclic Prefix

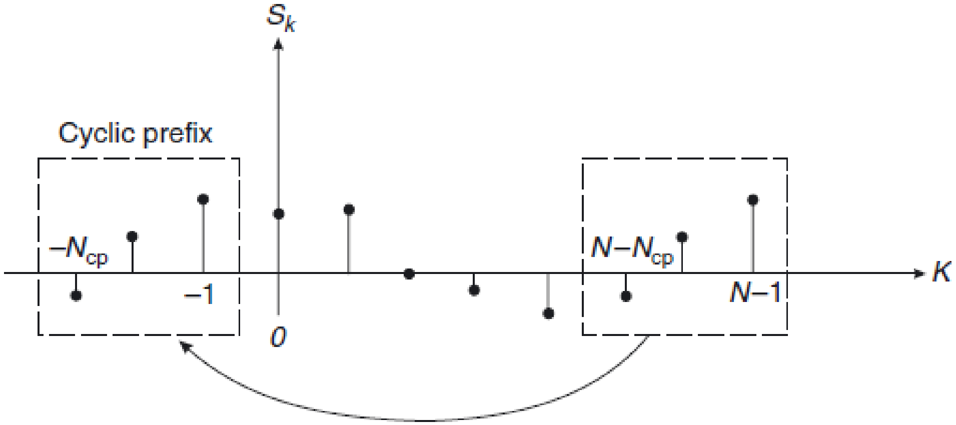
Let us first define a new base function for transmission:

$$g_n(t) = \exp\left[j2\pi n\frac{W}{N}t\right] \quad \text{for} - T_{cp} < t < \hat{T}_S$$

- When transmitting any data stream over a delay-dispersive channel, the arriving signal is the linear convolution of the transmitted signal with the channel impulse response.
- The cyclic prefix converts this *linear convolution into a cyclical convolution.*
- *During* the time $-T_{cp} < t < -T_{cp} + \tau_{max}$, *where* τ_{max} *is the maximum excess delay of the channel, the* received signal suffers from "real" InterSymbol Interference (ISI), as echoes of the last part of the preceding symbol interfere with the desired symbol.
- This "regular" ISI is eliminated by discarding the received signal during this time interval

In the receiver, there is a bank of filters that are matched to the basis functions *without the cyclic prefix:*

$$\bar{g}_n(t) = \begin{cases} g_n^*(\hat{T}_S - t) & \text{for } 0 < t < \hat{T}_S \\ 0 & \text{otherwise} \end{cases}$$

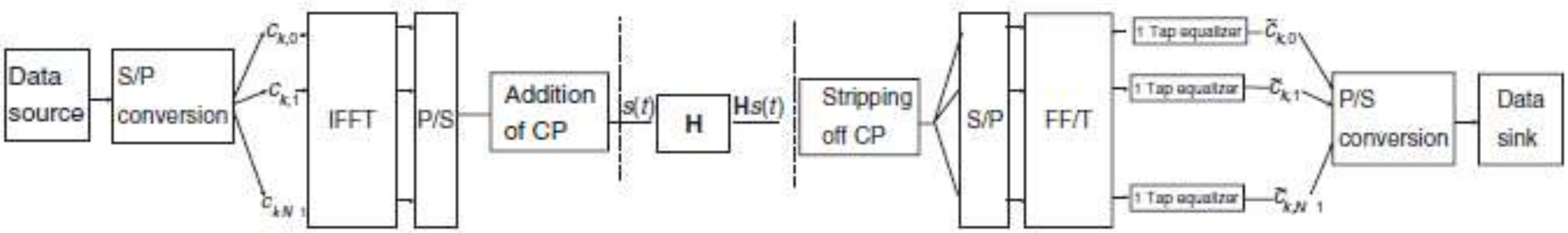


This operation removes the first part of the received signal (of duration *Tcp) from the detection* process; the matched filtering of the remainder can be realized as an FFT operation. The signal at the output of the matched filter is thus convolution of the transmit signal with the channel impulse response and the receive filter:

$$r_{n,0} = \int_0^{\hat{T}_s} \left[\int_0^{T_{cp}} h(t,\tau) \left(\sum_{k=0}^{N-1} c_{k,0} g_k(t-\tau) \right) d\tau \right] g_n^*(t)\, dt + n_n$$

The inner integral can be written as $$\exp\left[j2\pi t k \frac{W}{N} \right] \int_0^{T_{cp}} h(\tau) \exp\left(-j2\pi \tau k \frac{W}{N} \right) d\tau = g_k(t) H\left(k \frac{W}{N} \right)$$

- The original data stream is S/P converted.
- Each block of *N data symbols is subjected to an IFFT,* and then the last *NTcp/TS samples are prepended.*
- *The resulting signal is modulated onto a (single)* carrier and transmitted over a channel, which distorts the signal and adds noise.
- At the receiver, the signal is partitioned into blocks.
- For each block, the cyclic prefix is stripped off, and the remainder is subjected to an FFT.
- The resulting samples (which can be interpreted as the samples in the frequency domain) are “equalized” by means of one-tap equalization – i.e., division by the complex channel attenuation – on each carrier.



Peak-to-Average Power Ratio

- One of the major problems of OFDM is that the peak amplitude of the emitted signal can be considerably higher than the average amplitude.
- This *Peak-to-Average Ratio (PAR) issue originates from* the fact that an OFDM signal is the superposition of *N sinusoidal signals on different subcarriers.*
- On average the emitted power is linearly proportional to *N. However, sometimes, the signals on the* subcarriers add up constructively, so that the *amplitude of the signal is proportional to N, and the* power thus goes with *N2. We can thus anticipate the (worst case) power PAR to increase linearly* with the number of subcarriers.
- If the number of subcarriers is large, we can invoke the central limit theorem to show that the distribution of the amplitudes of in-phase components is Gaussian, with a standard deviation $\sigma = 1/\sqrt{2}$ (and similarly for the quadrature components) such that mean power is unity. Since both in-phase and quadrature components are Gaussian, the absolute amplitude is Rayleigh distributed .
- Knowing the amplitude distribution, it is easy to compute the probability that the instantaneous amplitude will lie above a given threshold, and similarly for power.

There are three main methods to deal with the Peak-to-Average Power Ratio (PAPR):

1. Put a power amplifier into the transmitter that can amplify linearly up to the possible *peak* value of the transmit signal. This is usually not practical, as it requires expensive and power-consuming class-A amplifiers. The larger the number of subcarriers *N, the more difficult* this solution becomes.
2. Use a nonlinear amplifier, and accept the fact that amplifier characteristics will lead to distortions in the output signal. Those nonlinear distortions destroy orthogonality between subcarriers, and also lead to increased out-of-band emissions (*spectral regrowth – similar to third-order intermodulation* products – such that the power emitted outside the nominal band is increased). The first effect increases the BER of the desired signal, while the latter effect causes interference to other users and thus decreases the cellular capacity of an OFDM system. This means that in order to have constant adjacent channel interference we can trade off power amplifier performance against spectral efficiency
3. Use PAPR reduction techniques.

Peak-to-Average Ratio Reduction Techniques

- **Coding for PAR reduction**: under normal circumstances, each OFDM symbol can represent one of 2N codewords (assuming BPSK modulation). Now, of these codewords only a subset of size 2K is acceptable in the sense that its PAR is lower than a given threshold. Both the transmitter and the receiver know the mapping between a bit combination of length K, and the codeword of length N that is chosen to represent it, and which has an admissible PAR.
- The transmission scheme is thus the following:
  - (i) parse the incoming bitstream into blocks of length K;
  - (ii) select the associated codeword of length N;
  - (iii) transmit this codeword via the OFDM modulator.
  - The coding scheme can guarantee a certain value for the PAR. It also has some coding gain, though this gain is smaller than for codes that are solely dedicated to error correction.

Phase adjustments: this scheme first defines an ensemble of phase adjustment vectors φl, l = 1, . . . , L, that are known to both the transmitter and receiver; each vector has N entries {φn}l . The transmitter then multiplies the OFDM symbol to be transmitted cn by each of these phase vectors to get

$$\{\hat{c}_n\}_l = c_n \exp[j(\phi_n)_l]$$

Correction by multiplicative function: another approach is to multiply the OFDM signal by a time-dependent function whenever the peak value is very high. The simplest example for such an approach is the clipping we mentioned in the previous subsection: if the signal attains a level sk >A0, it is multiplied by a factor A0/sk. In other words, the transmit signal becomes

$$\hat{s}(t) = s(t)\left[1 - \sum_k \max\left(0, \frac{|s_k| - A_0}{|s_k|}\right)\right]$$

Correction by additive function:

- in a similar spirit, we can choose an additive, instead of a multiplicative, correction function.
- The correction function should be smooth enough not to introduce significant out-of-band interference.
- Furthermore, the correction function acts as additional pseudo noise, and thus increases the BER of the system

THANKS!

Does anyone have any questions?

Rama.athilingam@gmail.com
9095100240