3.5 CYCLIC PREFIX AND WINDOWING

Cyclic prefix is a key element of enabling the OFDM signal to operate reliably.

The cyclic prefix acts as a buffer region or guard interval to protect the OFDM signals from inter symbol interference. This can be an issue in some circumstances even with the much lower data rates that are transmitted in the multicarrier OFDM signal.

The basic concept behind the OFDM cyclic prefix is quite straightforward.

The cyclic prefix performs two main functions.

- The cyclic prefix provides a guard interval to eliminate inter symbol interference from the previous symbol.
- It repeats the end of the symbol so the linear convolution of a frequency- selective multipath channel can be modeled as circular convolution, which in turn may transform to the frequency domain via a discrete Fourier transform. This approach accommodates simple frequency domain processing, such as channel estimation and equalization.

The cyclic prefix is created so that each OFDM symbol is preceded by a copy of the end part of that same symbol.

Different OFDM cyclic prefix lengths are available in various systems.

For example within LTE a normal length and an extended length are available and after release a third extended length is also included, although not normally used.

Advantages of Cyclic prefix

- Provides robustness: The addition of the cyclic prefix adds robustness to the OFDM signal. The data that is retransmitted can be used if required.
- Reduces inter-symbol interference: The guard interval introduced by the cyclic prefix enables the effects of inter-symbol interference to be reduced.

Disadvantages

- Reduces data capacity: As the cyclic prefix re-transmits data that is already being transmitted, it takes up system capacity and reduces the overall data rate.

The use of a cyclic prefix is standard within OFDM and it enables the performance to be maintained even under conditions when levels of reflections and multipath propagation are high.

Cyclic Prefix is referred as the Guard time and cyclic extension

To eliminate inter symbol interference completely, a guard time is introduced for each OFDM symbol. The guard time is chosen larger than the expected delay spread, such that multipath components from one symbol cannot interfere with the next symbol. The guard time could consist of no signal at all.

When an OFDM receiver tries to demodulate the first subcarrier, it will encounter some interference from the second subcarrier, because within the FFT interval, there is no integer number of cycle's difference between subcarrier I and 2.

At the same time, there will be crosstalk from the first to the second subcarrier for the same reason.

To eliminate ICI, the OFDM symbol is cyclically extended in the guard time, this ensures that delayed replicas of the OFDM symbol always have an integer number of cycles within the FFT interval, as long as the delay is smaller than the guard time. As a result, multipath signals with delays smaller than the guard time cannot cause ICI.

Hence, the subcarriers are not orthogonal anymore, but the interference is still small enough to get a reasonable received constellation.

WINDOWING

In OFDM signals, sharp phase transitions caused by the modulation can be seen at the symbol boundaries. An OFDM signal consists of a number of unfiltered QAM subcarriers. As a result, the out-of-band spectrum decreases rather slowly, according to a sine function. For larger number of subcarriers, the spectrum goes down more rapidly in the beginning, which is caused by the fact that the side lobes are closer together. The spectrum for 256 subcarriers has a relatively large -40-dB bandwidth that is almost four times the -3-dB bandwidth.

To make the spectrum go down more rapidly, windowing can be applied to the individual OFDM symbols. Windowing an OFDM symbol makes the amplitude go smoothly to zero at the symbol boundaries.

A commonly used window type is the raised cosine window, which is defined as

$$w(t) = \begin{cases} 0.5 + 0.5\cos(\pi + t\pi/(\beta T_s)) & 0 \le t \le \beta T_s \\ 1.0 & \beta T_s \le t \le T_s \\ 0.5 + 0.5\cos((t - T_s)\pi/(\beta T_s)) & T_s \le t \le (1+\beta)T_s \end{cases}$$

Here, Tso, is the symbol interval, which is shorter than the total symbol duration because we allow adjacent symbols to partially overlap in the roll-off region.

In practice, the OFDM signal is generated as follows:

First, Nc input QAM values are padded with zeros to get N input samples that are used to calculate an IFFT. Then, the last Prefix, fix samples of the IFFT output are inserted at the start of the OFDM symbol, and the first Postfix samples are appended at the end. The OFDM symbol is then multiplied by a raised cosine window w (t) to more quickly reduce the power of out-of-band subcarriers.

The OFDM symbol is then added to the output of the previous OFDM symbol with a delay of Tso, such that there is an overlap region of βTso, where β is the roll off factor of the raised cosine window.

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 Power Ratio (PAPR) 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 N^2.

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.

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-An 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 orthogonally 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 (see Figure 3.4.3), while the latter effect causes interference to other users and thus decreases the cellular capacity of an OFDM system (see Figure 3.4.4). This means that in order to have constant adjacent channel interference we can trade off power amplifier performance against spectral efficiency (note that increased carrier separation decreases spectral efficiency).

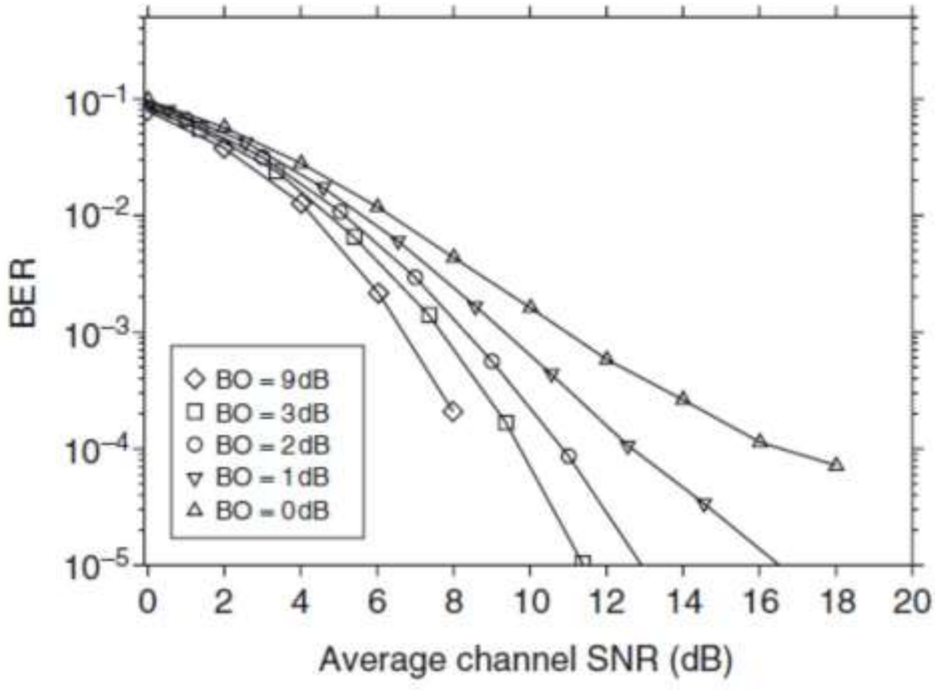


Fig 3.4.3 Bit error rate as a function of the signal-to-noise ratio, for different back off levels of the transmit amplifier.

[Source: "Wireless communications" by Andreas F.Molisch, Page-430]

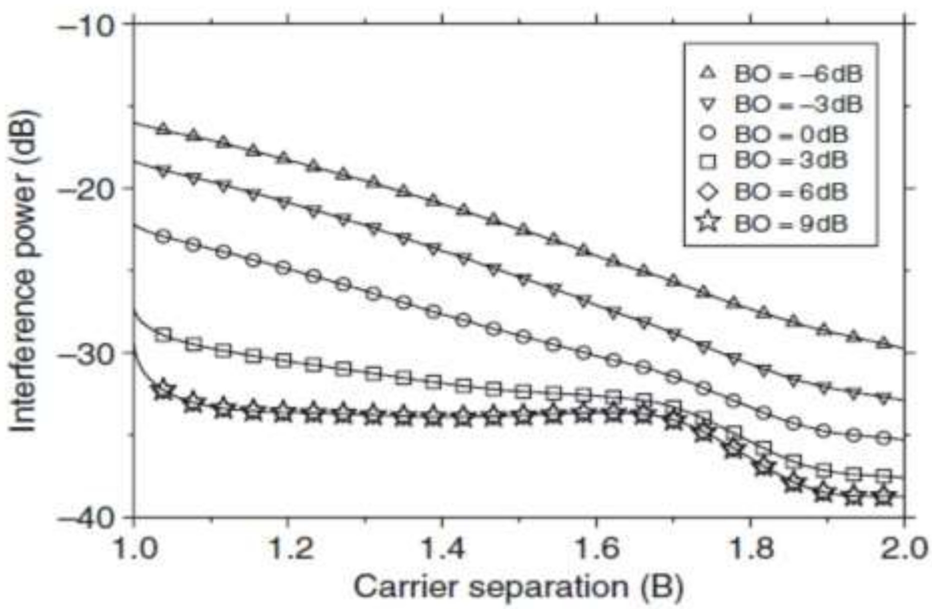


Fig 3.4.4 Interference power to adjacent bands (OFDM users), as a function of carrier separation

[Source: "Wireless communications" by Andreas F.Molisch, Page-430]

3. Use PAR reduction techniques.

1. Coding for PAR reduction: Under normal circumstances, each OFDM symbol can represent one of 2*N* code words (assuming BPSK modulation). Now, of these code words only a subset of size 2*K* 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 code word 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 bit stream into blocks of length *K*; (ii) select the associated code word of length *N*; (iii) transmit this code word via the OFDM modulator. The coding scheme can guarantee a certain value for the PAR.

2. Correction by multiplicative function: Here we 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

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

A less radical method is to multiply the signal by a Gaussian function centered at times when the level exceeds the threshold:

$$\hat{s}(t) = s(t)\left[1 - \sum_n \max\left(0, \frac{|s_k| - A_0}{|s_k|}\right) \exp\left(-\frac{t^2}{2\sigma_t^2}\right)\right]$$

Multiplication by a Gaussian function of variance a_t^2 in the time domain implies convolution with a Gaussian function in the frequency domain with variance $a_f^2 = 1/(2\pi a_t^2)$. Thus, the amount Of out-of-band interference can be influenced by the judicious choice of a_t^2.

3.6 ERROR PERFORMANCE IN FADING CHANNELS

When signal propagate for long distance its signal strength decreases .It is called fading.

Signal fading depends on 1. Nature of the transmitted signal 2. Characteristics of the channel

Slow flat fading channels change much slower than the applied modulation.

Flat fading channels cause a gain variation in the transmitted signal S (t).

Then the received signal r (t) can be written as

$$r(t) = \propto(t)\,\text{expel}(-j8(t))s(t) + n(t) \qquad 0 \le t \le T$$

$\propto(t)$ - gain of the channel

8(t) – phase shift of the channel n

(t) – additive white Gaussian noise

The probability of error is

$$P_{\alpha} = \int_0^{\infty} P_O(x)\,P(x)\,dx$$

$$P(x) = \frac{1}{T}\,\text{expel}\left(-\frac{X)}{T}\right) \qquad x > 0$$

$T = \frac{Eb}{N_0}\alpha^2$ is the average value of SNR.

$\propto^2 = 1$ for the unity gain fading channel.

Frequency selective fading caused by multipath time delay spread causes ISI which results in irreducible BER floor for mobile systems.

The error floor is a frequency selective channel which is caused by the errors due to the ISI which interfere with the signal component at the receiver sampling instants.

This occurs when

a).the main signal component is removed through multipath cancellations.

b). non zero value causes inter symbol interference

c) .the sampling time of a receiver is shifted as a result of delay spread

Normalized RMS delay spread $d = \frac{an}{T_s}$

Based on the results of simulation we can say that, for small delay spreads flat fading is dominant which causes errors.

For large delay spread, timing errors and ISI are the dominant error mechanisms.

Error performance of $\frac{G}{4}$ DPSK **in fading and interference**:

Here the channel can be modelled as a frequency selective 2 ray channel with AWGN and co channel interference.

In a slow fading channel, the multipath dispersion and Doppler spread are negligible and errors are caused by fading and co channel interference.

If C/I > 20 dB, the errors are due to fading and interference has little effect.

If C/I < 20 dB, then interference dominates link performance.

____

3.3 MINIMUM SHIFT KEYING (MSK)

Minimum shift keying (MSK) is a special type of continuous phase. Frequency shift keying (CPFSK) where in the peak frequency deviation is equal to
1/4 the bit rate. In other words, MSK is continuous phase FSK with a modulation index of 0.5.

A modulation index of 0.5 corresponds to the minimum frequency spacing that allows two FSK signals to be coherently orthogonal, and the name minimum shift keying implies the minimum frequency separation (i.e. bandwidth) that allows orthogonal detection.

MSK is a spectrally efficient modulation scheme and is particularly attractive for use in mobile radio communication systems. It possesses properties such as constant envelope, spectral efficiency, good bit error rate performance, and self- synchronizing capability.

An MSK signal can be thought of as a special form of OQPSK where the baseband rectangular pulses are replaced with half-sinusoidal pulses.

Consider the OQPSK signal with the bit streams offset.

If half-sinusoidal pulses are used instead of rectangular pulses, the modified signal can be defined as MSK and for an N-bit stream is given by

$$S_{MSK}(t) = \sum_{i=0}^{N-1} m_I(t)p(t-2iT_b)\cos 2\pi f_c t + \sum_{i=0}^{N-1} m_Q(t)p(t-2iT_b-T_b)\sin 2\pi f_c t$$

$$\text{where } p(t) = \begin{cases} \sin\left(\frac{\pi t}{2T_b}\right) & 0 \le t \le 2T_b \\ 0 & elsewhere \end{cases}$$

And where mi (t) and mms (t) are the "odd" and "even" bits of the bipolar data stream which have values of ± 1 and which feed the in-phase and quadrature arms of the modulator at a rate of Rib/2.

One version of MSK uses only positive half-sinusoids as the basic pulse shape, another version uses alternating positive and negative half-sinusoids as the basic pulse shape.

MSK can be modified in trigonometric identities as,

$$S_{MSK}(t) = \sqrt{\frac{2E_b}{T_b}} \cos\left[2\pi f_c t - m_I(t) m_Q(t) \frac{\pi t}{2T_b} + \phi_k\right]$$

Where wk. is 0 or it depends on whether M1 (t) is 1 or -1.

From the above equation it can be deduced that MSK has a constant amplitude. Phase continuity at the bit transition periods is ensured by choosing the carrier frequency to be an integral multiple of one fourth the bit rate, 1 /4T.

MSK Power Spectrum

For MSK, the baseband pulse shaping function is given by

$$p(t) = \begin{cases} \cos\left(\frac{\pi t}{2T}\right) & |t| < T \\ 0 & \text{elsewhere} \end{cases}$$

The normalized power spectral density for MSK is given by

$$P_{MSK} = \frac{16}{\pi^2}\left(\frac{\cos 2\pi(f + f_c)T}{1.16 f^2 T^2}\right)^2 + \frac{16}{\pi^2}\left(\frac{\cos 2\pi(f - f_c)T}{1.16 f^2 T^2}\right)^2$$

The MSK spectrum has lower side lobes than QPSK and OQPSK.

Ninety-nine percent of the MSK power is contained within a bandwidth B = 1.2/T, while for QPSK and OQPSK, the 99 percent bandwidth B is equal to 8/T.

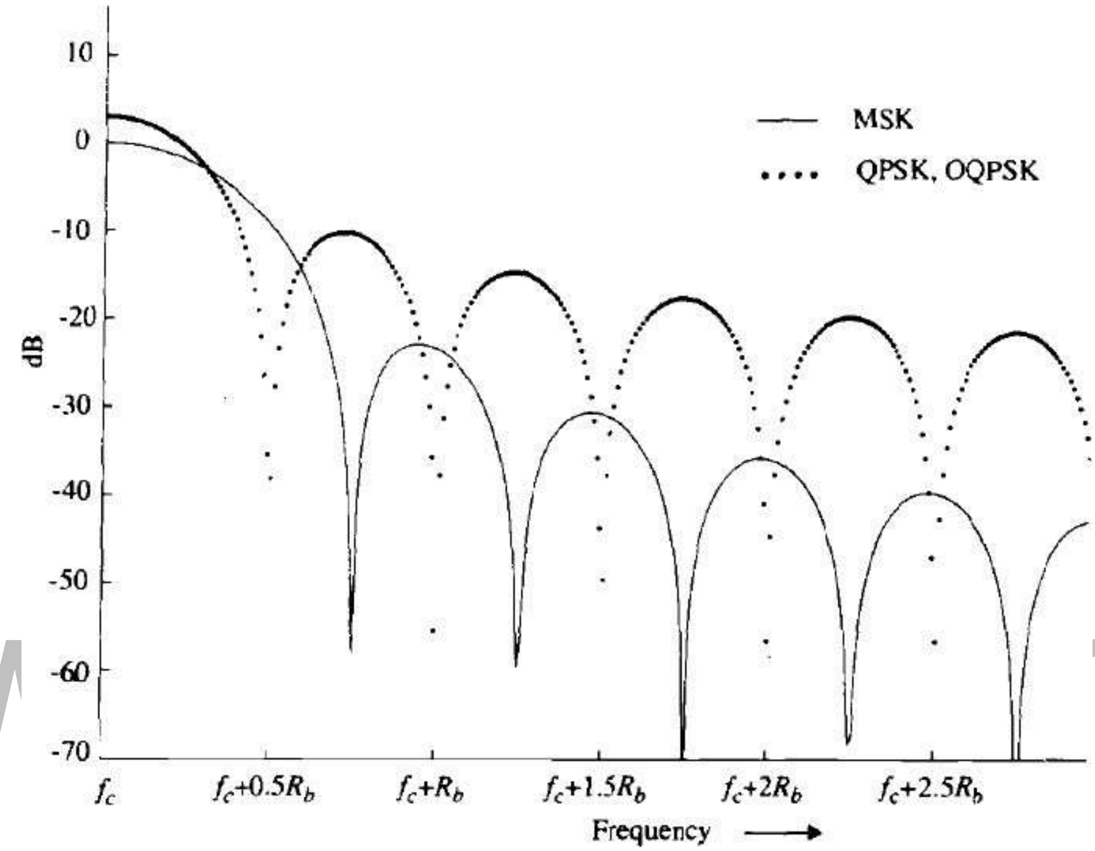


Fig 3.3.1: MSK spectrum with side lobes

[Source: "Wireless communications" by Theodore S. Rappaport, Page-261]

The faster roll off of the MSK spectrum is due to the fact that smoother pulse functions are used.

Figure 3.3.1, shows that the main lobe of MSK is wider than that of QPSK and OQPSK, and hence when compared in terms of first null bandwidth, MSK is less spectrally efficient than the phase-shift keying techniques.

MSK Transmitter and Receiver

MSK transmitter is called MSK modulator as shown in figure 3.3.2. Multiplying a carrier signal with coos [πt/2T] produces two phase-coherent signals at fc+ 1 /4T and fc- I /4T.

These two FSK signals are separated using two narrow band pass filters and combined to form the in-phase and quadrature carrier components x (t) and y (t), respectively. These carriers are multiplied with the odd and even bit streams, rn1 (t) and mms (t), to produce the MSK modulated signal SMSK (t).

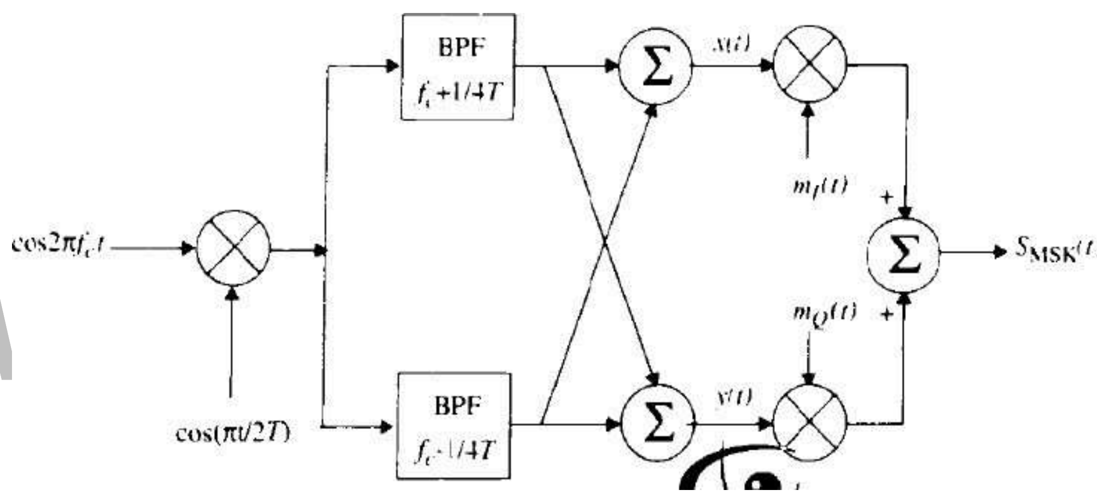


Fig 3.3.2: MSK transmitter

[Source: "Wireless communications" by Theodore S. Rappaport, Page-262]

The block diagram of an MSK receiver is shown in Figure 3.3.3.

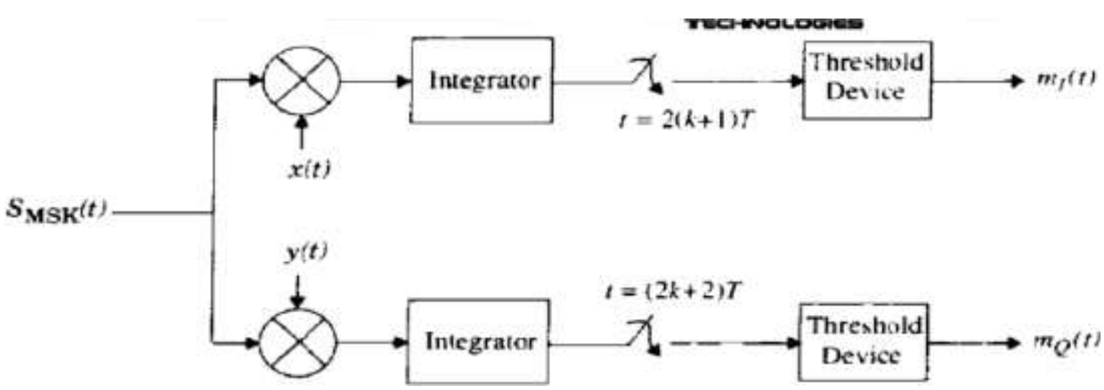


Fig 3.3.3: MSK Receiver

[Source: "Wireless communications" by Theodore S. Rappaport, Page-262]

The received signal SMSK (t) (in the absence of noise and interference) is multiplied by the respective in-phase and quadrature carriers x (t) and y (t).

The output of the multipliers are integrated over two bit periods and dumped to a decision circuit at the end of each two bit periods. Based on the level of the signal at the output of the integrator, the threshold detector decides whether the signal is a 0 or a 1. The output data streams correspond to m1 (t) and mms (t), which are offset combined to obtain the demodulated signal.

GAUSSIAN MINIMUM SHIFT KEYING

GMSK is a simple binary modulation scheme which is the derivative of MSK.

In GMSK, the side lobe levels of the spectrum are further reduced by passing the modulating NRZ data waveform through a pre modulation Gaussian pulse-shaping filter

.

Baseband Gaussian pulse shaping smooth's the phase trajectory of the MSK signal and hence stabilizes the instantaneous frequency variations over time. This has the effect of reducing the side lobe levels in the transmitted spectrum.

Pre modulation Gaussian filtering converts the full response message signal (where each baseband symbol occupies a single bit period T) into a partial response scheme where each transmitted symbol spans several bit periods.

GMSK can be coherently detected just as an MSK signal, or non-coherently detected as simple FSK. In practice, GMSK is most attractive for its excellent power efficiency (due to the constant envelope) and its excellent spectral efficiency.

The pre modulation Gaussian filtering introduces ISI in transmitted signal

The GMSK pre modulation filter has an impulse response given by

$$h_G(t) = \frac{\sqrt{\pi}}{\alpha}\exp\left(-\frac{\pi^2}{\alpha^2}t^2\right)$$

And the transfer function given by

$$H_G(f) = \exp(-\alpha^2 f^2)$$

$$\alpha = \frac{\sqrt{ln2}}{\sqrt{2}B} = \frac{0.5887}{B}$$

And the GMSK filter may be completely defined from B and the baseband symbol duration T.

The power spectrum of MSK, which is equivalent to GMSK with a BT product of infinity, is also shown for comparison purposes. It is clearly seen from the graph that as the BT product decreases, the side lobe levels fall off very rapid]y.

For example, for a BT=0.5, the peak of the second lobe is more than 30dB below the main lobe, whereas for simple MSK, the second lobe is only 20 dB below main lobe. Reducing BT increases the irreducible error rate produced by the low pass filter due to ISI.

Power spectrum of GMSK is shown in figure 3.3.4.

GMSK Bit Error Rate

The bit error probability is a function of BT, since the pulse shaping impacts ISI. The bit error probability for GMSK is given by

$$P_e = Q\left\{\sqrt{\frac{2\gamma E_b}{N_0}}\right\}$$

Where **γ** is a constant related to BT by

y = { o. 68 †or GMSK witM BT = o. 25
o. 85 †or MSK (BT) = ∞

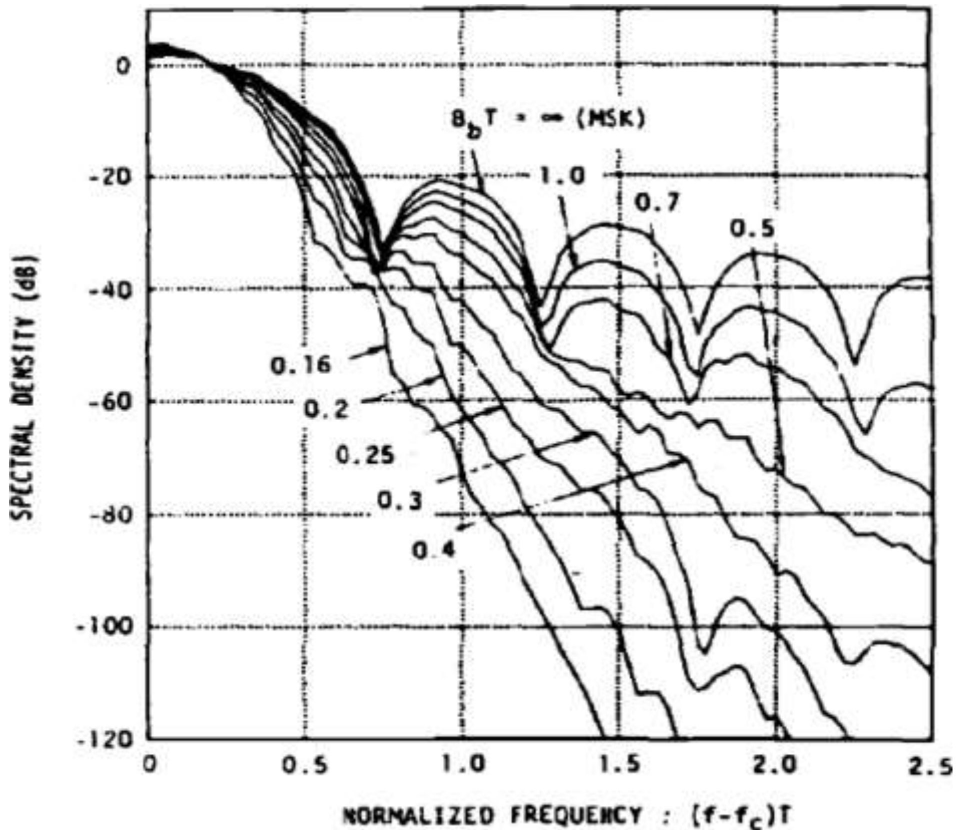


Fig 3.3.4: Power spectrum of GMSK

[Source : "Wireless communications" by Theodore S. Rappaport,Page-264]

GMSK Transmitter and Receiver

The simplest way to generate a GMSK signal is to pass a NRZ message bit stream through a Gaussian baseband filter having an impulse response followed by an FM modulator.

This modulation technique is shown in Figure 3.3.5 and is currently used in a variety of analog and digital implementations for the U.S. Cellular Digital Packet Data (CDPD) system as well as for the Global System for Mobile (GSM) system.

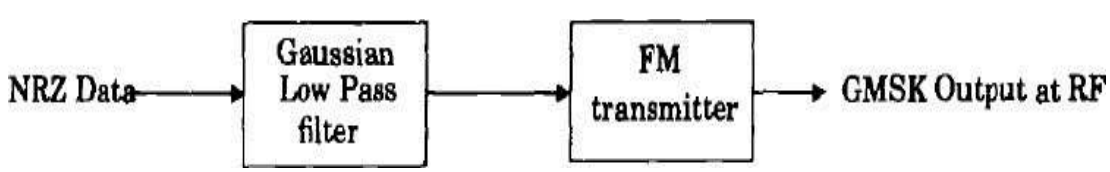


Fig3.3.5: GMSK transmitter using direct FM generation

[Source:" Wireless communications" by Theodore S. Rappaport, Page-265]

GMSK signals can be detected using orthogonal coherent detectors as sh3.3.6.

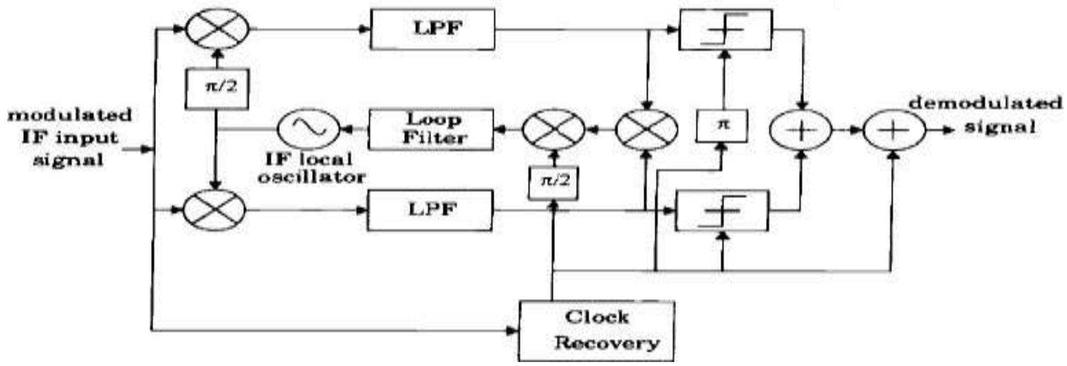
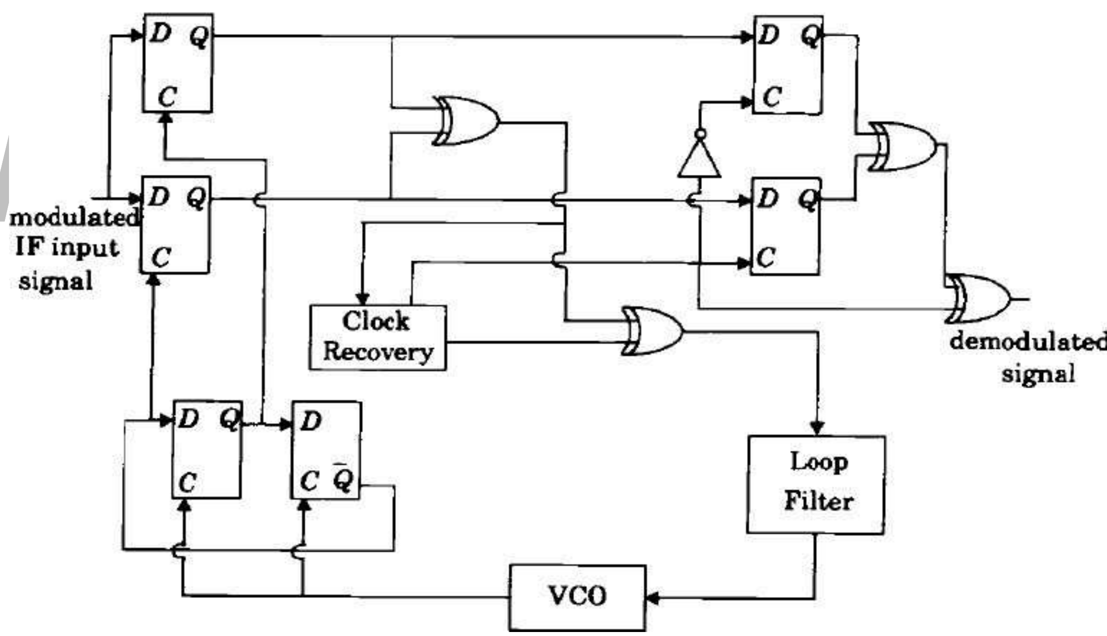


Fig 3.3.6: GMSK using orthogonal coherent detectors

[Source: "Wireless communications" by Theodore S. Rappaport, Page-266]

This type of receiver can be easily implemented using digital logic as shown in (figure.3.3.7). The two D flip-flops act as a quadrature product demodulator and the XOR gates act as baseband multipliers. The mutually orthogonal reference carriers are generated using two D flip- flops, and the VCO center frequency is set equal to four times the carrier center frequency.



A non-optimum, but highly effective method of detecting GMSK signal is to simply sample the output of an FM demodulator.

Fig 3.3.7: Digital logic circuit for GMSK demodulation

[Source: "Wireless communications" by Theodore S. Rappaport, Page-266]

3.4 ORTHOGONAL FREQUENCY DIVISION MULTIPLEXING

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.

Principle of Orthogonal Frequency Division Multiplexing

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.

Figure 3.4.1 shows this principle in the frequency domain. 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 expel (−j**2**πfnt) and integrating over symbol duration), the data streams of any two subcarriers will not interfere.

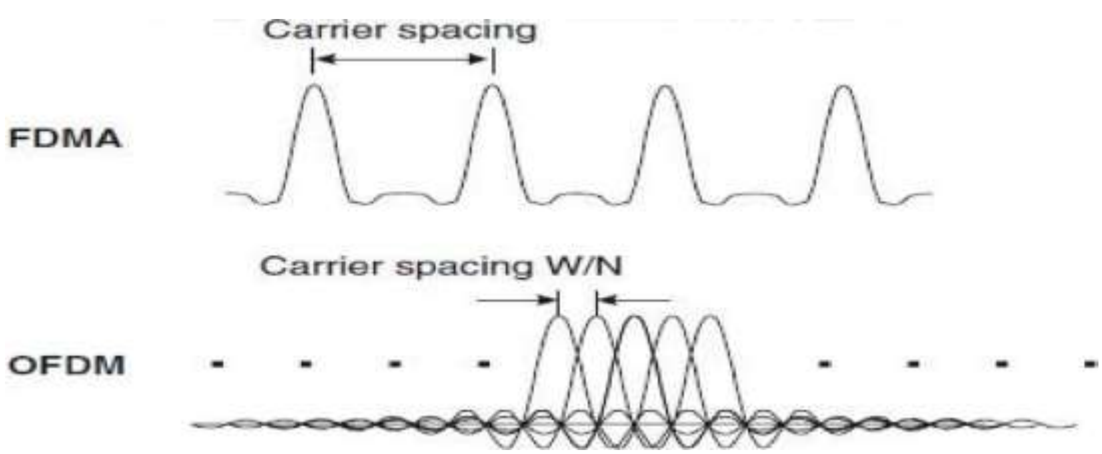


Fig3.4.1: Principle behind OFDM

[Source : "Wireless communications" by Andreas F.Molisch,Page-418]

Implementation of Transceivers

OFDM can be interpreted in two ways: One is an "analog" interpretation following from the picture of Figure 3.4.2.

First split the original data stream into N parallel data streams, each of which has a lower data rate. And have a number of local oscillators (LOs) available, each of which oscillates at a frequency fun = now/N, where n = 0, 1, N − 1.

Each of the parallel data streams then modulates one of the carriers. This picture allows an easy understanding of the principle.

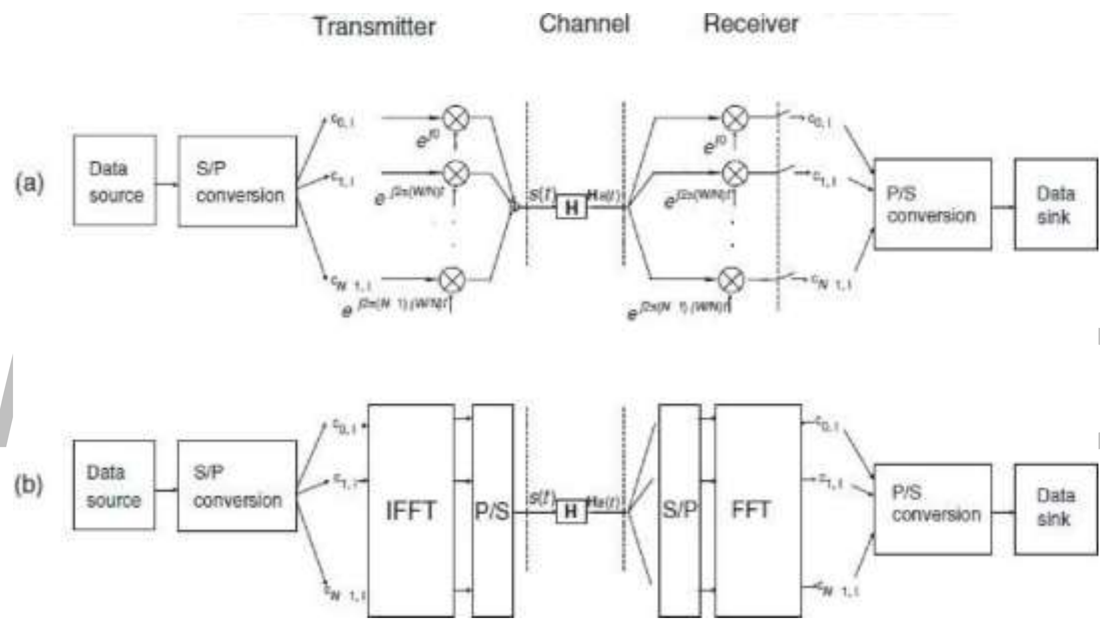


Fig 3.4.2. Transceiver structures for OFDM

[Source: "Wireless communications" by Andreas F.Molisch, Page-419]

An alternative implementation is digital. It first divides the transmit data into blocks of N symbols. Each block of data is subjected to an Inverse Fast Fourier Transformation (IFFT), and then transmitted.

This approach is much easier to implement with integrated circuits.

Let us first consider the analog interpretation. Let the complex transmit symbol at time instant i on the nth carrier be candid.

The transmit signal is then:

$$s(t) = \sum_{i=-\infty}^{\infty} s_i(t) = \sum_{i=-\infty}^{\infty} \sum_{n=0}^{N-1} c_{n,i} g_n(t - iT_S)$$

Where the basis pulse gnu (t) is a normalized, frequency-shifted rectangular pulse:

$$g_n(t) = \begin{cases} \frac{1}{\sqrt{T_S}} \exp\left(j2\pi n \frac{t}{T_S}\right) & \text{for } 0 < t < T_S \\ 0 & \text{otherwise} \end{cases}$$

Consider the signal only for i = 0, and sample it at instances tk = kTs/N:

$$s_k = s(t_k) = \frac{1}{\sqrt{T_S}} \sum_{n=0}^{N-1} c_{n,0} \exp\left(j2\pi n \frac{k}{N}\right)$$

This is the inverse Discrete Fourier Transform (DFT) of the transmit symbols.

Therefore, the transmitter can be realized by performing an Inverse Discrete Fourier Transform (IDFT) on the block of transmit symbols (the block size must equal the number of subcarriers).

In almost all practical cases, the number of samples N is chosen to be a power of 2, and the IDFT is realized as an IFFT.

Note that the input to this IFFT is made up of N samples (the symbols for the different subcarriers), and therefore the output from the IFFT also consists of N values. These N values now have to be transmitted, one after the other, as temporal samples – this is the reason why we have a P/S (Parallel to Serial) conversion directly after the IFFT.

At the receiver, sample the received signal, write a block of N samples into a vector – i.e., an S/P (Serial to Parallel) conversion – and perform an FFT on this vector.

The result is an estimate $\overline{C_n}$ of the original data C_{an}.

3.2 PRINCIPLES OF OFFSET QPSK

QUADRATURE-PHASE SHIFT KEYING

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

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

B1i and b2i each of which has a data rate that is half that of the original data stream:

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

Let us first consider the situation where basis pulses are rectangular pulses, g(t)=gray(t, TS).

Then we can give an interpretation of QPSK as either a phase modulation or as a PAM.

We first define two sequences of 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 band pass signal (as in figure 3.2.2) reads

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

Normalization is done in such a way that the energy within one symbol interval is_

$\int_0^{T_S} s_{BP}(t)^2 dt =$ **$2E_B$**, E_B is the energy expended on transmission of a bit.

Data streams of in-phase and quadrature-phase components in quadrature-phase shift keying as shown in Fig 3.2.1.

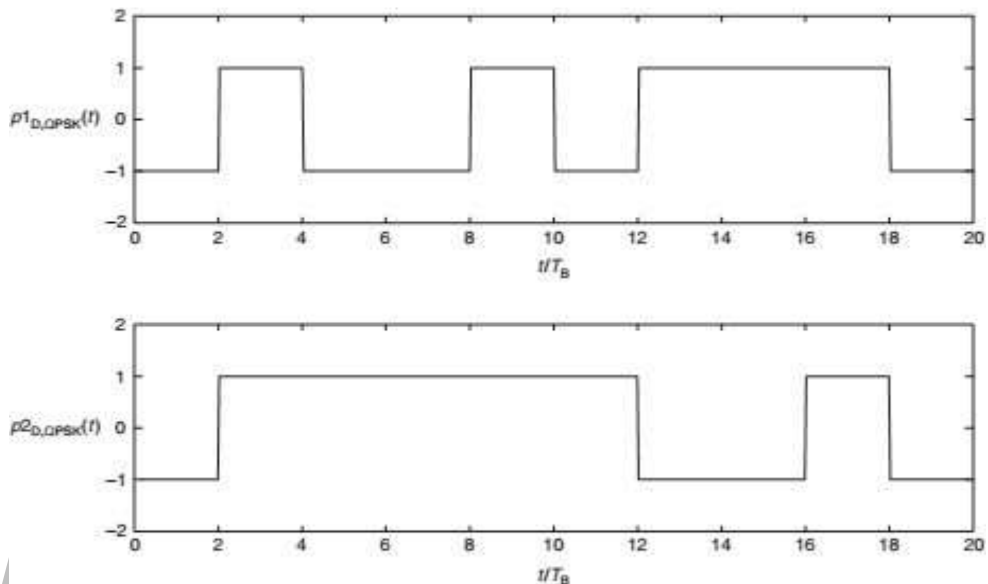


Fig 3.2.1: Data streams in –in phase and quadrature phase components
[Source: "Wireless communications" by Andreas F.Molisch, Page-200]

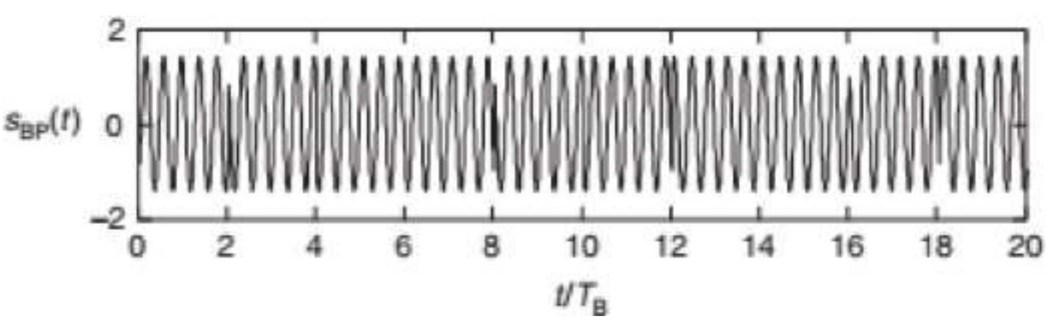


Fig 3.2.2: Quadrature-phase shift keying signal as a function of time
[Source: "Wireless communications" by Andreas F.Molisch, Page-201]

The baseband signal is

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

When interpreting QPSK as a phase modulation, the low pass signal can be written as

$\frac{F}{2}$ expel((t)) and ja$_{ck}$
E
B
T$_B$

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

It is clear from this representation that the signal is constant envelope, except for the transitions at t = its.

Signal space diagram of quadrature-phase shift keying (Fig 3.2.3).

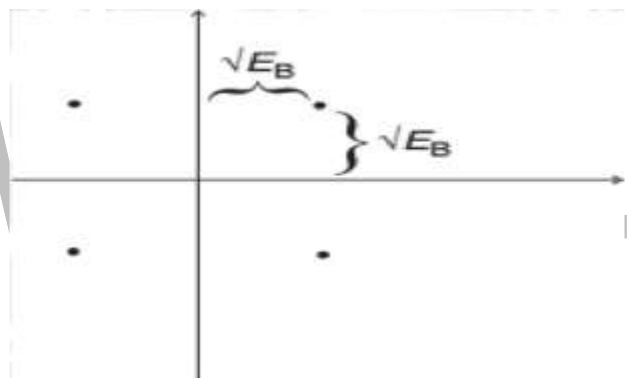


Fig 3.2.3: Signal space diagram of quadrature-phase shift keying
[Source: "Wireless communications" by Andreas F.Molisch, Page-201]

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.

Normalized power-spectral density of quadrature-phase shift keying.(Fig 3.2.4).

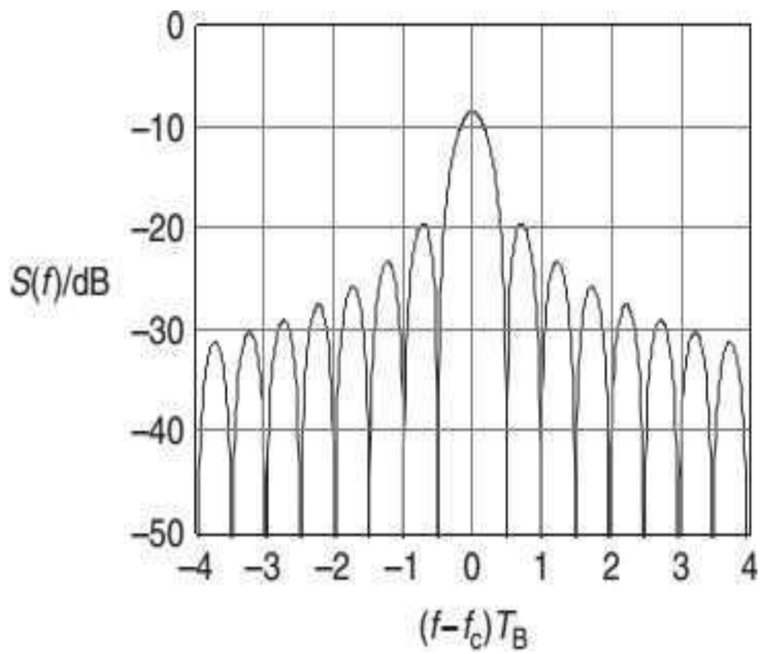


Fig 3.2.4 : Normalized PSD of quadrature-phase shift keying

[Source : "Wireless communications" by Andreas F.Molisch,Page-202]

………………………………………………………………..

OFFSET QUADRATURE-PHASE SHIFT KEYING

The method 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 bit streams modulating the in-phase and quadrature-phase components are offset half a symbol duration with respect to each other (shown in Figure 3.2.5) , 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

$$\left.\begin{aligned} p1_D(t) &= \sum_{i=-\infty}^{\infty} b1_i g(t - iT_S) = b1_i * g(t) \\ p2_D(t) &= \sum_{i=-\infty}^{\infty} b2_i g\left(t - \left(i + \tfrac{1}{2}\right) T_S\right) = b2_i * g\left(t - \frac{T_S}{2}\right) \end{aligned}\right\}$$

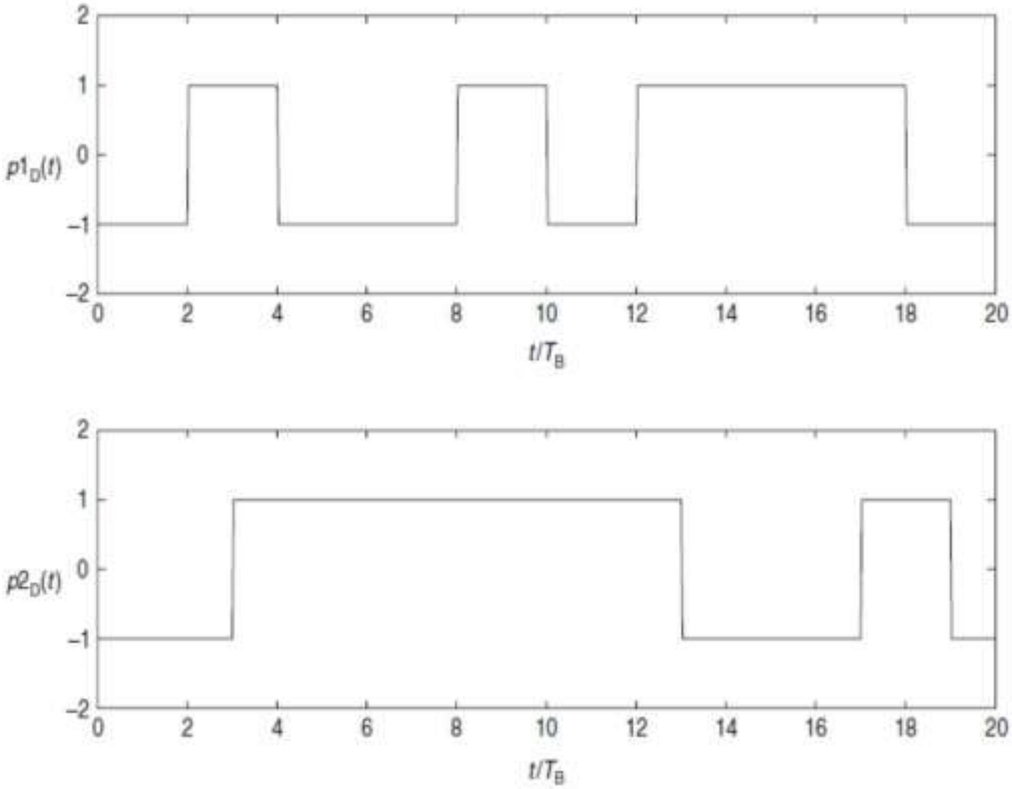


Fig 3.2.5: Sequence of basis pulses for offset QPSK

[Source : "Wireless communications" by Andreas F.Molisch,Page-206]

These data streams can again be used for interpretation as PAM or as phase modulation.

The resulting band pass signal is shown in figure 3.2.6.

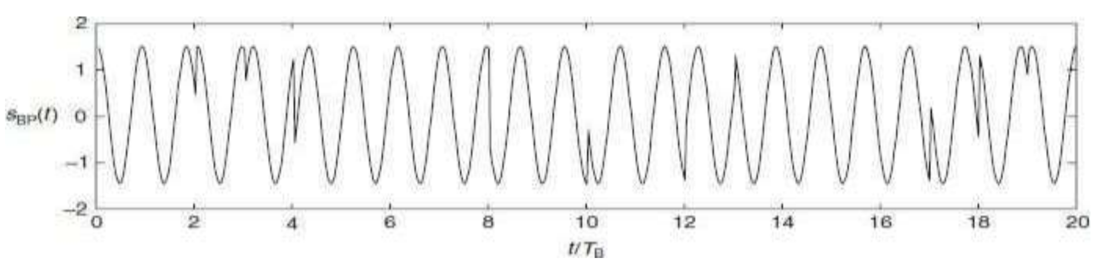


Fig3.2.6: Offset QPSK signal as a function of time

[Source :" Wireless communications" by Andreas F.Molisch,Page-206]

The representation in the I–Q diagram (Figure 3.2.7) makes clear that there are no transitions passing through the origin of the coordinate system; thus this modulation format takes care of envelope fluctuations.

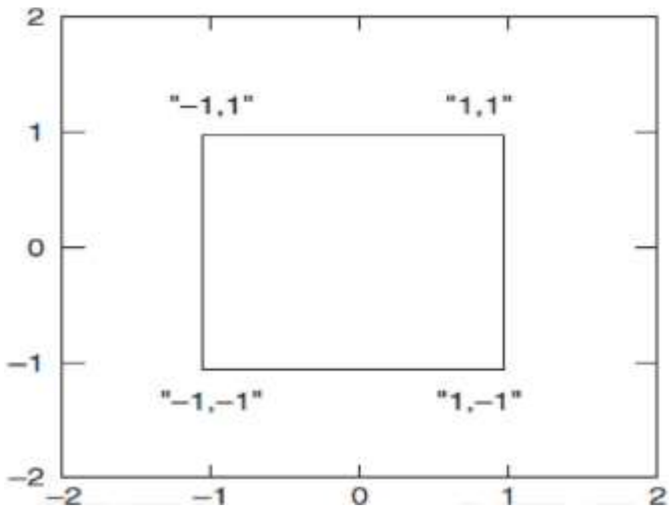


Fig 3.2.7: I–Q diagram for offset QPSK signal
[Source: "Wireless communications" by Andreas F.Molisch, Page-207]

I–Q diagram (above) for offset quadrature-phase shift keying with rectangular basis functions. Also shown are the four points of the normalized signal space diagram, (1, 1), (1, −1), (−1, −1), (−1, 1). Smoother basis pulses, like raised cosine pulses are used to improve spectral efficiency.

Figure 3.2.8, shows the I–Q diagram for offset quadrature amplitude modulation with raised cosine basis pulses.

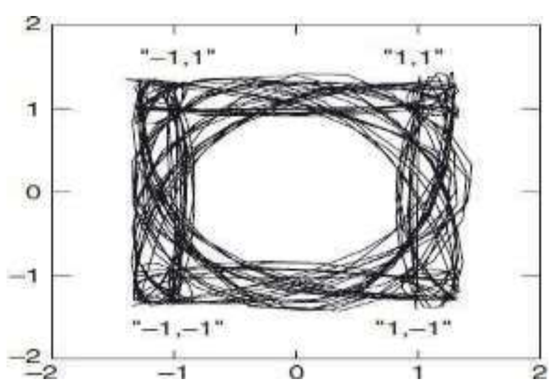


Fig3.2.8: I–Q diagram for offset QPSK signal with raised cosine basis pulses
[Source: "Wireless communications" by Andreas F.Molisch, Page-207]

Π/4-DIFFERENTIAL QUADRATURE-PHASE SHIFT KEYING

QPSK has amplitude dips at bit transitions; 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 $\pi/4$-DQPSK ($\pi/4$ differential quadrature-phase shift keying).

This modulation format had great importance for second-generation cellphones – it was used in several American standards (IS-54, IS-136, PWT), as well as the Japanese cellphone (JDC) and cordless (PHS) standards, and the European trunk radio standard (TETRA).

The principle of π/4-DQPSK can be understood from the signal space diagram of DQPSK (Figure3.2.9).

There exist two sets of signal constellations: (0, 90, 180, 270◦) and (45, 135, 225, 315◦).

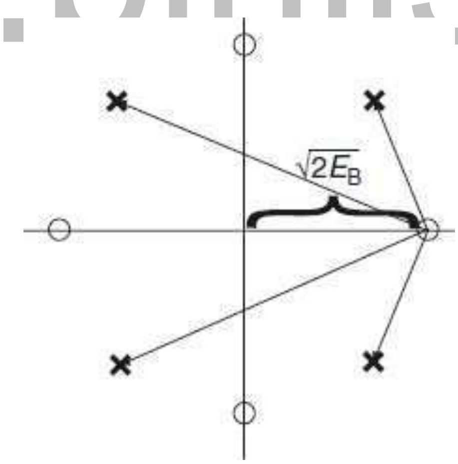


Fig 3.2.9: Allowed transitions in the signal space diagram

[Source: "Wireless communications" by Andreas F.Molisch, Page-204]

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 (Figure; in physical terms, this means smaller fluctuations of the envelope.
Figure 3.2.10 shows the Sequence of basis pulses for π/4 differential quadrature-phase shift keying.

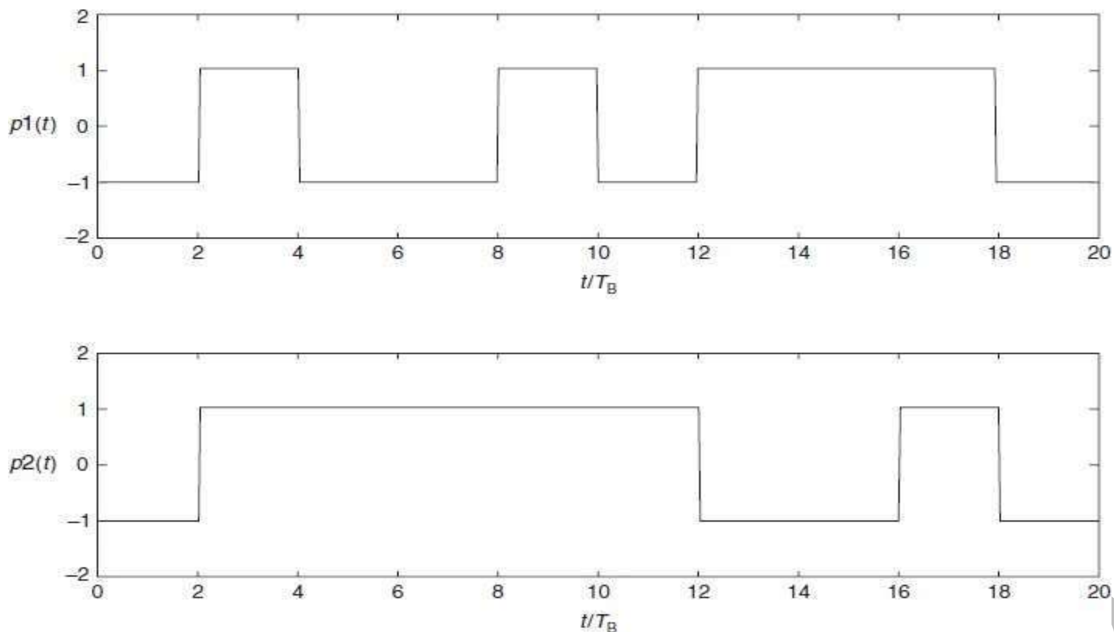


Fig 3.2.10. Basis pulses for π/4 differential PS

[Source: "Wireless communications" by Andreas F.Molisch, Page-204]

The signal phase is given by

$$\Phi_s(t) = \pi \left[\frac{1}{2} p2_D(t) - \frac{1}{4} p1_D(t) p2_D(t) + \frac{1}{4} \left\lfloor \frac{t}{T_S} \right\rfloor \right]$$

Where x denotes the largest integer smaller or equal to x. comparing this with Eq. equation below,

We can see the change in phase at each integer multiple of TS.

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

Figure (below) shows the data sequences, and the resulting band pass signals when using rectangular or raised cosine basis pulses.

Π/4 differential quadrature-phase shift keying signals as function of time for rectangular basis functions (a) and raised cosine basis pulses (b) are shown in figure 3.2.11.

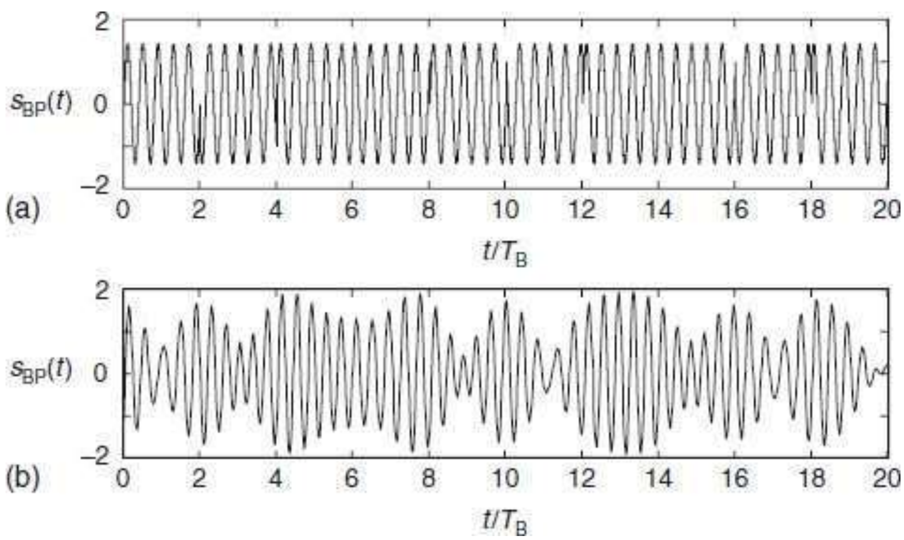


Fig 3.2.11. π/4 differential quadrature-phase shift keying signals as function of time for rectangular basis functions (a) and raised cosine basis pulses (b).

[Source: "Wireless communications" by Andreas F.Molisch, Page-205]

I-Q diagram for π/4 -differential quadrature-phase shift keying signal with rectangular basis functions is shown in figure 3.2.12.

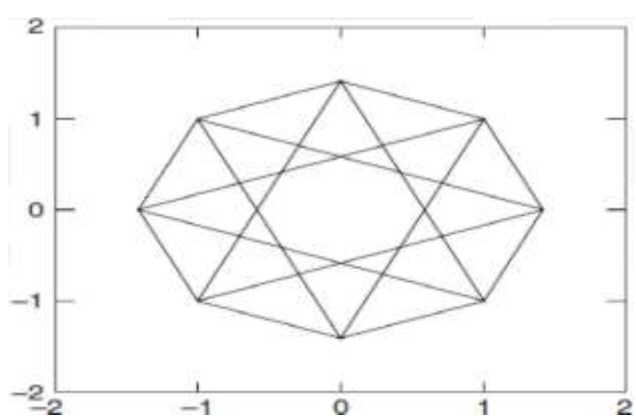


Fig 3.2.12: I-Q diagram of π/4 differential QPSK.

[Source: "Wireless communications" by Andreas F.Molisch, Page-205]

UNIT 3- DIGITAL SIGNALING FOR FADING CHANNELS

3.1 STRUCTURE OF WIRELESS COMMUNICATION LINK

The aim 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.

The transmitter (TX) can then add redundancy in the form of a forward error correction code, in order to make it more resistant to errors introduced by the channel.

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 sent through the propagation channel, which attenuates and distorts it, and adds noise.

At the receiver (RX), the signal is received by one or more antennas. The different users are separated (e.g., by receiving signals only at a single frequency).

If the channel is delay dispersive, then an equalizer can be used to reverse that dispersion, and eliminate inter symbol interference.

After wards, the signal is demodulated, and a channel decoder eliminates (most of) the errors that are present in the resulting bit stream.

The figure 3.1.1. Show a more detailed block diagram of a digital TX and RX that concentrate on the hardware aspects and the interfaces between analog and digital components:

The TX Digital to Analog Converter (DAC) generates a pair of analog, discrete amplitude voltages corresponding to the real and imaginary part of the transmit symbols, respectively.

The analog low-pass filter in the TX eliminates the (inevitable) spectral components outside the desired transmission bandwidth.

These components are created by the out- of band emission of an (ideal) baseband modulator, which stem from the properties of the chosen modulation format.

Furthermore, imperfections of the baseband modulator and imperfections of the DAC lead to additional spurious emissions that have to be suppressed by the TX filter.

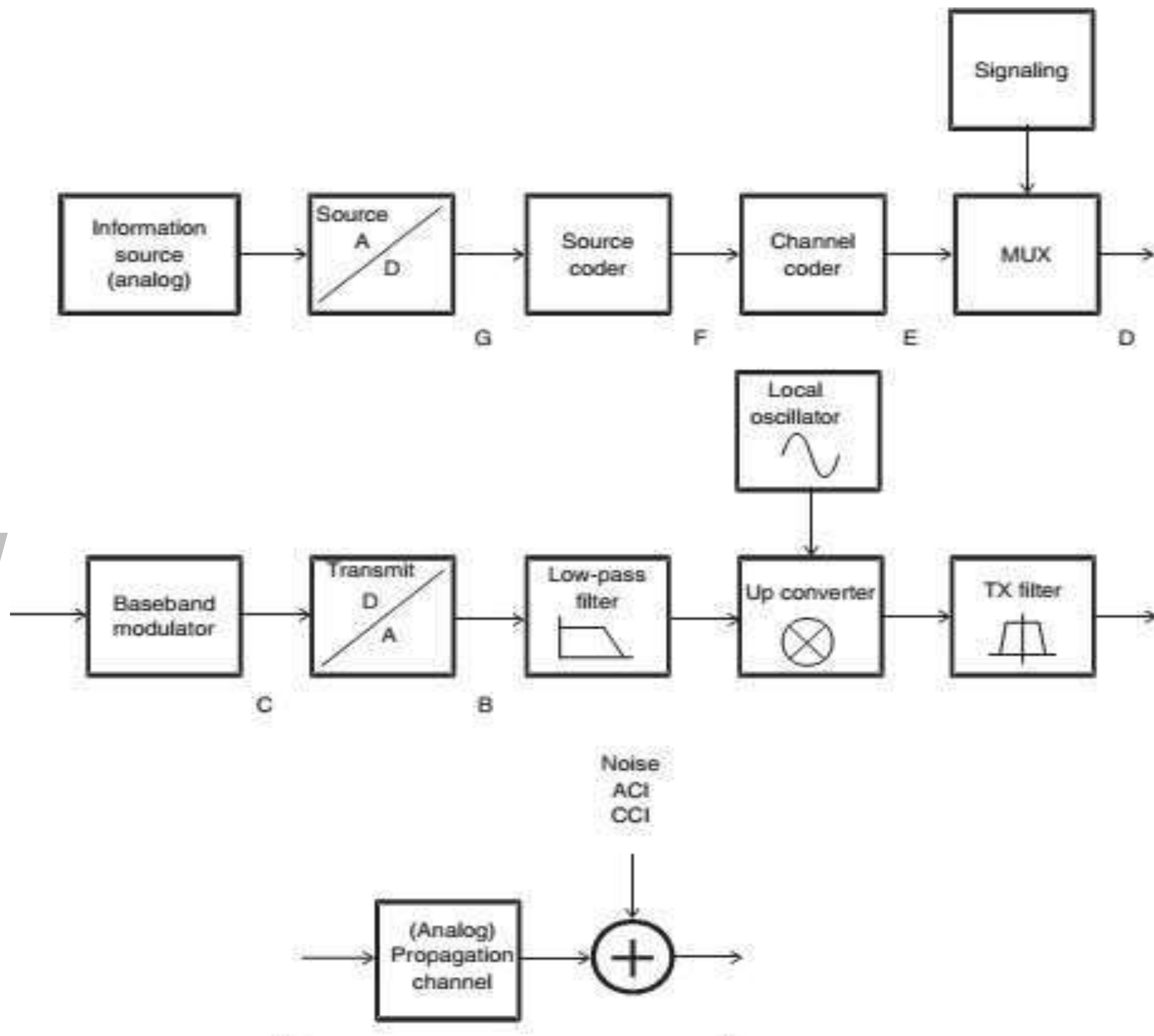


Fig 3.1.1: Digital receiver chain of wireless communication

[Source: "Wireless communications" by Andreas F.Molisch, Page-183]

The TX Local Oscillator (LO) provides a UN modulated sinusoidal signal, corresponding to one of the admissible center frequencies of the considered system.

The requirements for frequency stability, phase noise, and switching speed between different frequencies depend on the modulation and multi access method.

The **up converter** converts the analog, filtered baseband signal to a pass band signal by mixing it with the LO signal. Up conversion can occur in a single step, or in several steps. Finally, amplification in the Radio Frequency (RF) domain is required.

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, up conversion can lead to the creation of additional out-of-band components.

Especially, nonlinearities of mixers and amplifiers lead to inter modulation products and "spectral regrowth" – i.e., creation of additional out-of-band emissions.

The (analog) propagation channel attenuates the signal, and leads to delay and frequency dispersion. Furthermore, the environment adds noise (Additive White Gaussian Noise – AWGN) and co-channel interference.

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). Further amplification occurs in the subsequent steps of down conversion.

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 (see below), 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 (in one or several steps) 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 (in contrast to the RX band pass filter that selects the frequency range in which the service operates).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. The sampling rate is of limited importance as long as the conditions of the sampling theorem are fulfilled. Oversampling increases the requirements for the ADC, but simplifies 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.

If there are multiple antennas, then the RX either selects the signal from one of them for further processing or the signals from all of the antennas have to be processed (filtering, amplification, down conversion).

In the latter case, those baseband signals are then either combined before being fed into a conventional baseband demodulator or they are fed directly into a “joint” demodulator that can make use of information from the different antenna elements.

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 simplest case of a un coded system, the decoder is just a hard-decision (threshold) device.

For convolutional codes, Maximum Likelihood Sequence Estimators (MLSEs, such as the Viterbi decoder) are used.

Signaling recovery identifies the parts of the data that represent signaling information and controls the subsequent DE multiplexer.

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