MICRO-461
Low-power Radio Design for the IoT

3. Wireless Communication

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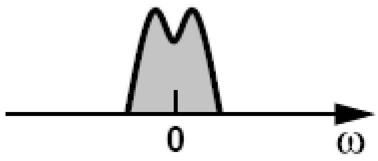
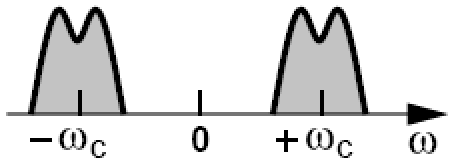
Outline

- **General Considerations**
- Analog Modulation
- Digital Modulation
- Power Efficiency of Modulation Schemes
- Noncoherent Detection

Baseband and Bandpass Waveforms



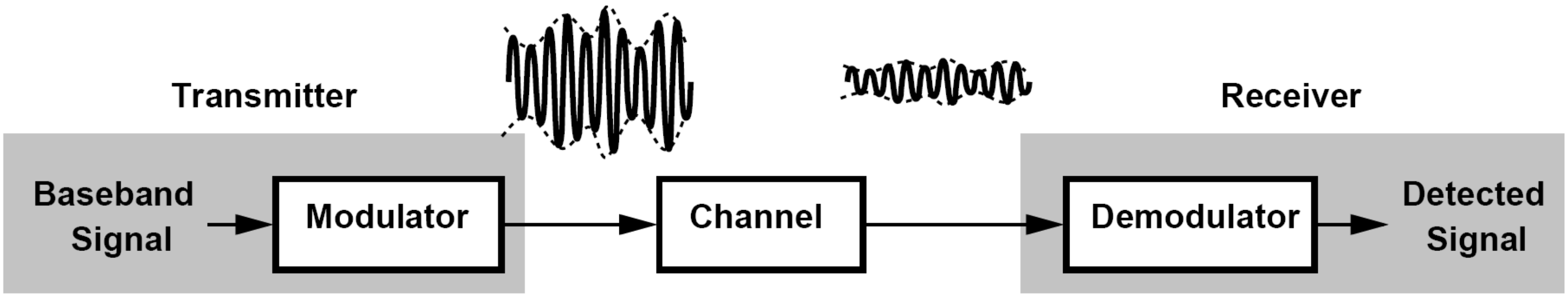
Baseband



Bandpass

- Transmitted waveform in RF communications is usually a high frequency carrier modulated by the signal to be transmitted, because
  - In wireless, the antenna size should be a significant fraction of the carrier wavelength to achieve a reasonable gain
  - Because of regulation, communication can only occur in certain parts of the spectrum
  - In certain cases, modulation allows simpler detection at the receive end in presence of nonidealities in the communication channel
- Two types of signals:
  - **Baseband** signal: nonzero in vicinity of $\omega=0$ and zero elsewhere
  - **Bandpass** signals: waveform with spectrum nonzero around a carrier frequency ω_c

Modulation and Demodulation



- Modulation converts a baseband signal into a bandpass counterpart by varying certain parameters of the sinusoidal carrier

$$x(t) = a(t) \cdot \cos\left[\omega_c t + \theta(t)\right]$$

- where $a(t)$ and $\theta(t)$ are usually real functions of time
- $\phi(t) = \omega_c t + \theta(t)$ is the total phase and $\theta(t)$ the excess phase
- The instantaneous frequency is the time derivative of the phase $\omega_c + d\theta/dt$ is the total frequency and $d\theta/dt$ is the excess frequency
- The inverse of modulation is demodulation (or detection) with the goal of extracting the baseband signal with maximum signal-to-noise ratio

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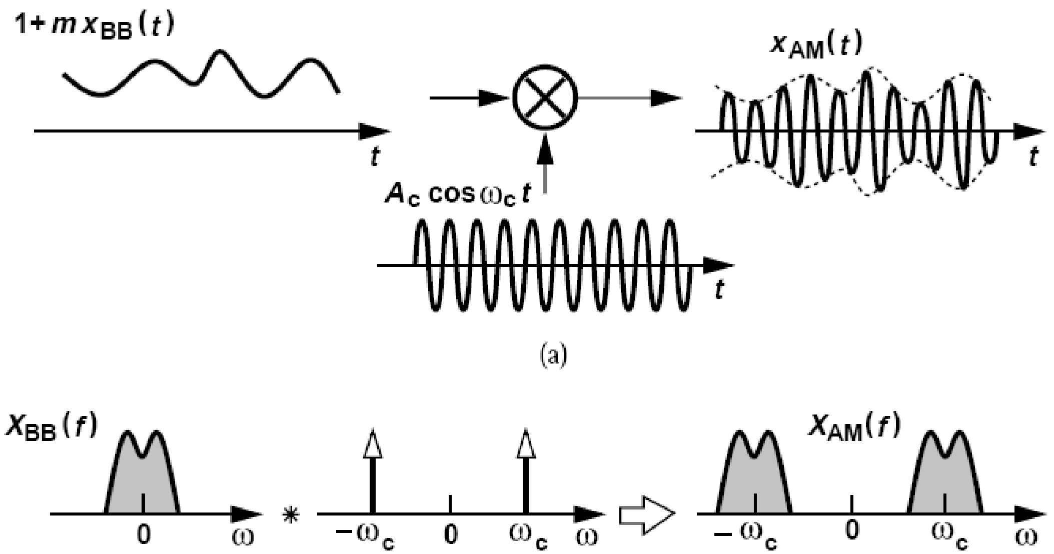
Analog Modulation – Amplitude Modulation

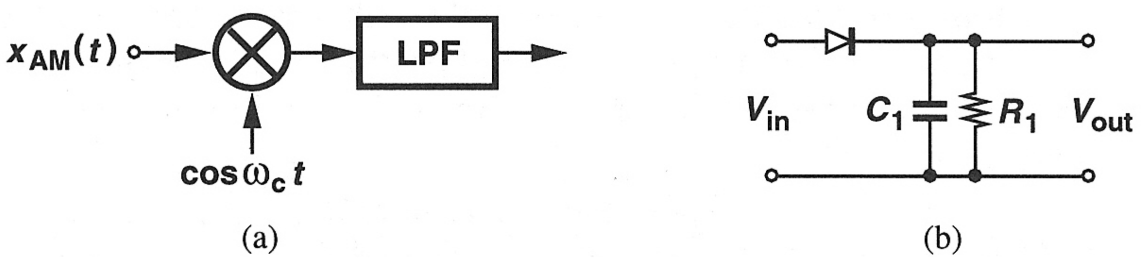


- Amplitude modulation (AM) waveform

$$x_{AM}(t) = A_c\left[1 + m \cdot x_{BB}(t)\right]\cos\omega_c t$$

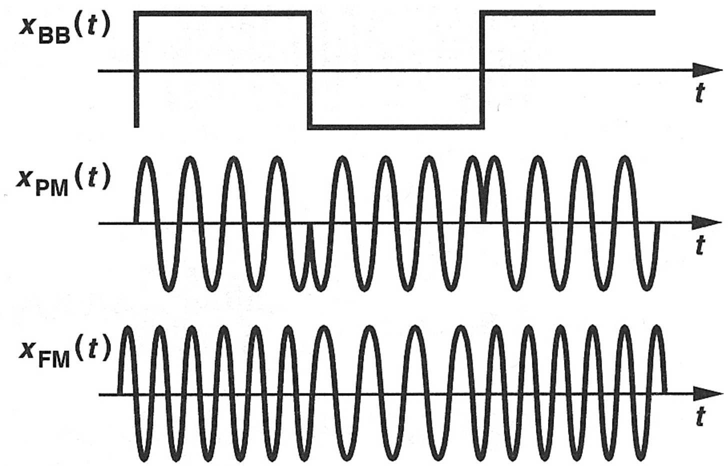
- where m is the **modulation index**
- The bandwidth of $x_{AM}(t)$ is twice the bandwidth of $x_{BB}(t)$
- The signal $A_c \cos\omega_c t$ is generated by a local oscillator (LO)

Analog Modulation – Amplitude Modulation



- The AM signal can be demodulated using a multiplier and a LP filter (envelope detection)
- Except for broadcast radios and sound in television, AM finds limited use in today's wireless systems
- This is because carrying information in the amplitude both makes the signal susceptible to noise and requires a highly linear power amplifier in the transmitter

Analog Modulation – Phase and Frequency Modulations

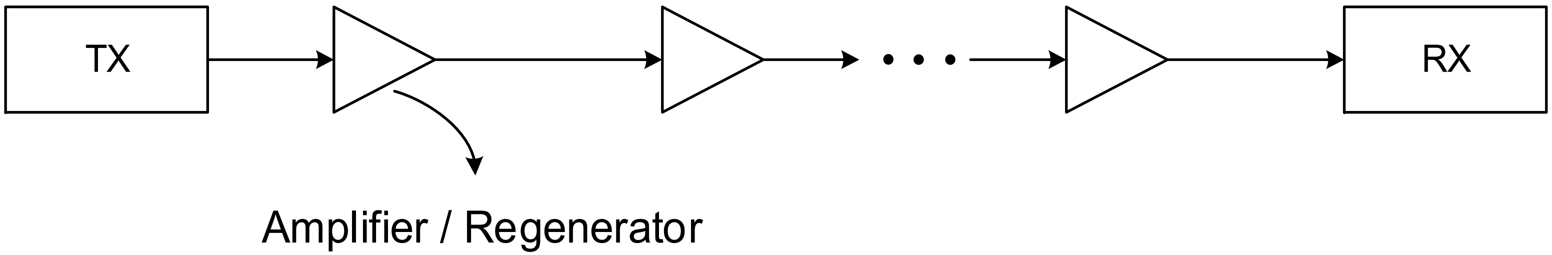


- Phase or Frequency modulation (PM or FM) waveform

$$x_{PM}(t) = A_c \cos\left[\omega_c t + \theta(t)\right] \quad \text{with} \quad \theta(t) = \begin{cases} m \cdot x_{BB}(t) & \text{for PM} \\ m \cdot \int_{-\infty}^{t} x_{BB}(t)dt & \text{for FM} \end{cases}$$

- where m is the **modulation index**

Analog vs. Digital



- Imagine a system in which you transmit a signal over a vey large distance
- Due to very large attenuation amplifiers are put between the transmitter and receiver
- If the transmitted signal is **analog**, each amplifier degrades the signal quality (adds noise), the larger the distance the worse the signal quality
- If the transmitted signal is **digital** (a discrete set of symbols is used for transmission), it can be fully reconstructed at each regeneration stage
- It is possible to reconstruct the TX signal on the RX side without the loss of quality

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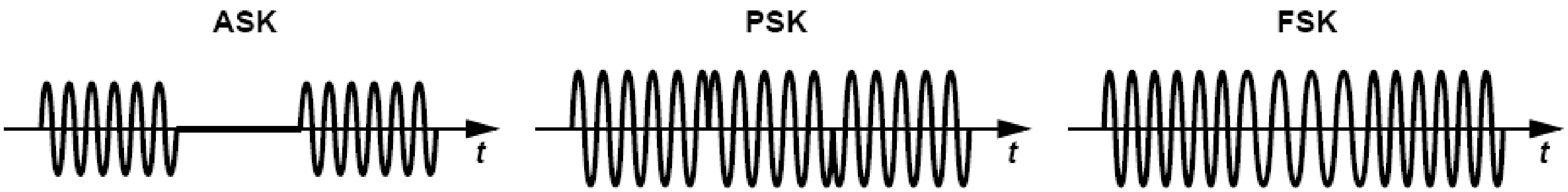
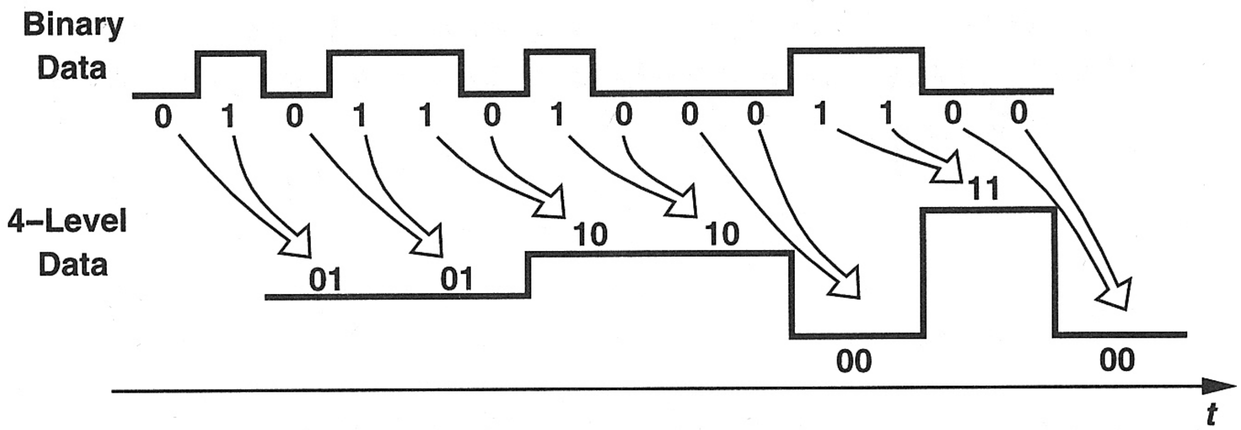
Digital Modulation



- Carrier modulated by a digital baseband signal
- Digital counter parts of AM, PM and FM
  - Amplitude shift keying (ASK)
  - Phase shift keying (PSK) and
  - Frequency shift keying (FSK)
- In RF, PSK and FSK more widely used because of a their lower sensitivity to amplitude noise

Binary and M-ary Signaling



- Digital waveform $x_{BB}(t) = \sum_n b_n \cdot p\left(t - nT_b\right)$
  - Binary: b_n can take two distinct discrete values (+1 and -1 or +1 and 0)
  - M-ary: b_n can take M distinct discrete levels
- In the example below ($M = 4$), the original bit stream is converted into groups of 2 bits and each group is converted into 1 of 4 possible levels
- Fewer transitions per unit time, but higher amplitude resolution in the receiver
- Symbol rate is ½ the bit rate

Basis Functions

- Binary FSK

$$x_{FSK}(t) = \begin{cases} A_c \cdot \cos \omega_1 t & \text{if } b_n = 0 \\ A_c \cdot \cos \omega_2 t & \text{if } b_n = 1 \end{cases}$$

- Can also be represented by a linear combination of two **orthogonal basis functions** $\phi_1(t)$ and $\phi_2(t)$
- Where $x_{FSK}(t) = a_1 \cdot \varphi_1(t) + a_2 \cdot \varphi_2(t) = \begin{bmatrix} a_1 & a_2 \end{bmatrix} \cdot \begin{bmatrix} \varphi_1(t) & \varphi_2(t) \end{bmatrix}$

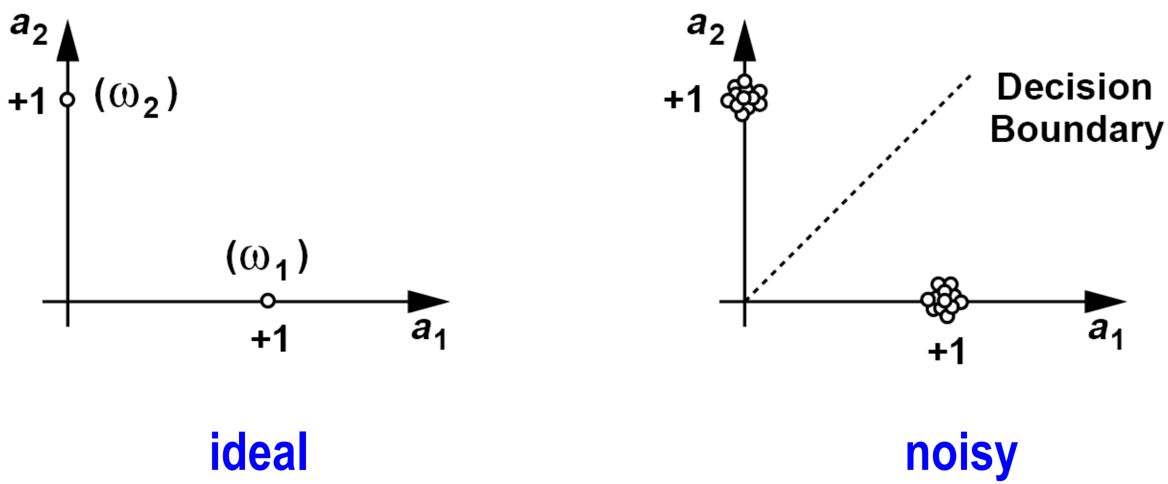
  where $\varphi_1(t) = A_c \cdot \cos \omega_1 t$ and $\varphi_2(t) = A_c \cdot \cos \omega_2 t$
- More generally each symbol in a digital modulated waveform is represented by a linear combination of orthogonal basis functions

$$x(t) = \sum_{n=1}^{N} \alpha_n \cdot \varphi_n(t) \quad \text{with} \quad \int_0^{T_S} \varphi_m(t)\varphi_k(t)dt = 0 \quad \text{for } m \neq k$$

- where T_S is the **symbol period** and N the **dimension** of the set of basis functions

Signal Constellations

- Modulated waveforms can be visualized in terms of the a_1 and a_2 **coefficients** ignoring the basis functions
- a_1 and a_2 can be plotted in a Cartesian coordinates resulting in the **signal constellation** (or signal space)



- Such plots proves valuable in understanding various effects (such as noise, distortion, ...) in digital modems

Signal Constellations

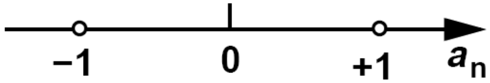
- Another example is the ASK modulation

$$x_{ASK}(t) = \begin{cases} A_c \cdot \cos \omega_c t & \text{if } b_n = 1 \\ 0 & \text{if } b_n = 0 \end{cases} \quad \text{or} \quad x_{ASK}(t) = a_1 \cdot \varphi_1(t)$$

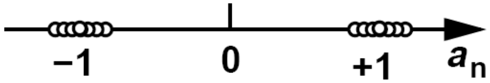
- Where

$$a_1(t) = \begin{cases} A_c & \text{if } b_n = 1 \\ 0 & \text{if } b_n = 0 \end{cases} \quad \text{and} \quad \varphi_1(t) = \cos \omega_c t$$

- This scheme has **one dimension** with the constellation shown below



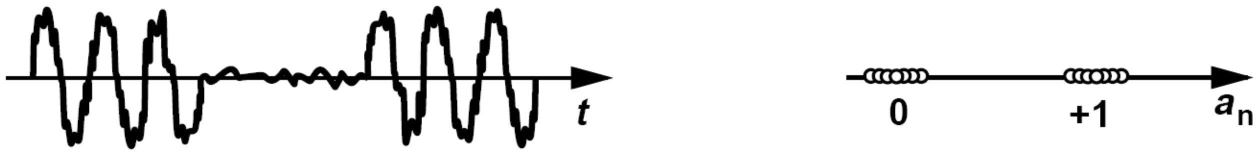
ideal



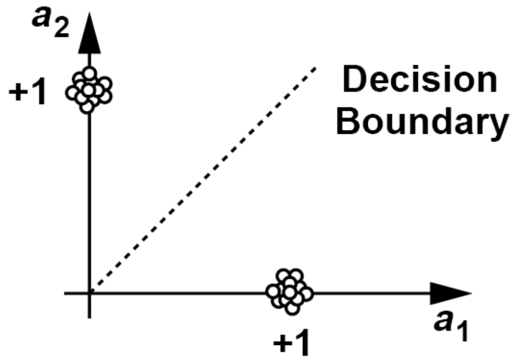
noisy

Effect of Noise on Signal Constellation

- The effect of noise on an ASK and FSK signals and their corresponding constellations are shown below

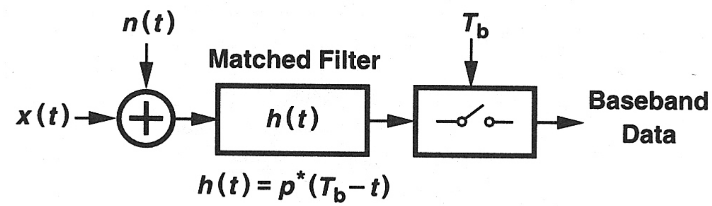


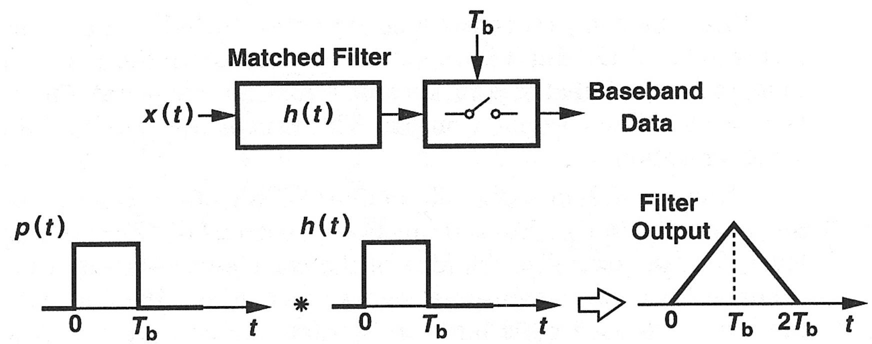
ASK with noise



FSK with noise

Optimum Detection – Matched Filter





- The SNR is maximized when the impulse response $h(t)$ of the filter is given by

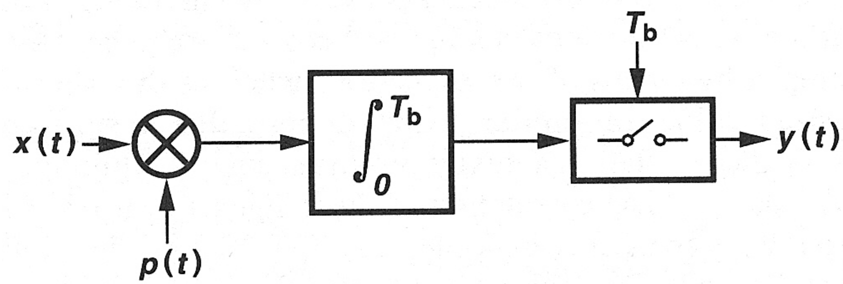
$$h(t) = p^{*}\left(T_b - t\right)$$

- where T_b is the bit duration
- If the matched filter input is a pulse $p(t)$ and additive white noise with PSD $N_0/2$, the maximum SNR is given by

$$SNR_{\max} = \frac{2E_p}{N_0} \quad \text{with} \quad E_p = \int_{-\infty}^{+\infty} |p(t)|^2 \, dt$$

- where E_p is the signal energy

Optimum Detection – Correlator



- The output of the matched filter at the end of the bit period is given by

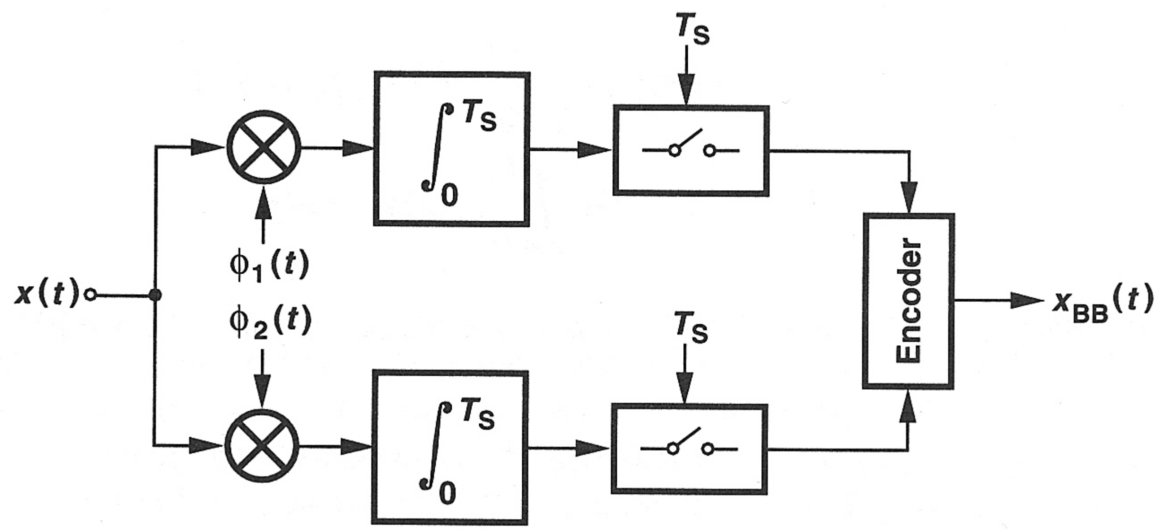
$$y(T_b) = \int_{-\infty}^{+\infty} x(\tau)h(T_b - \tau)d\tau = \int_{-\infty}^{+\infty} x(\tau)p(\tau)d\tau$$

- where $x(t) = p(t) + n(t)$. If $p(t)$ is zero outside $[0, T_b]$, then

$$y(T_b) = \int_0^{T_b} x(\tau)p(\tau)d\tau$$

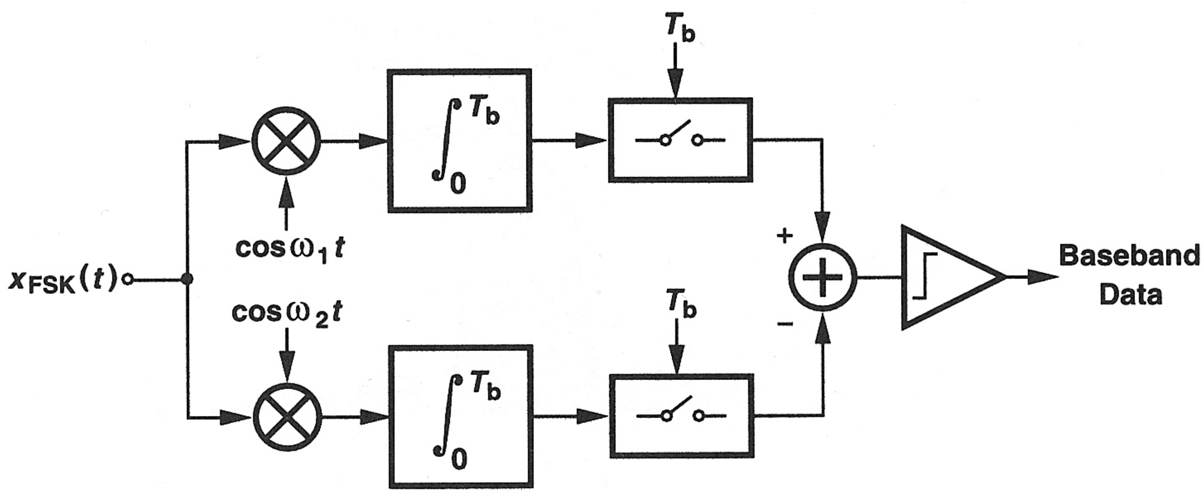
- which suggests that a correlation function must be performed between the input (signal + noise) and the known pulse shape $p(t)$ for one bit period

Optimum Detection – Correlator



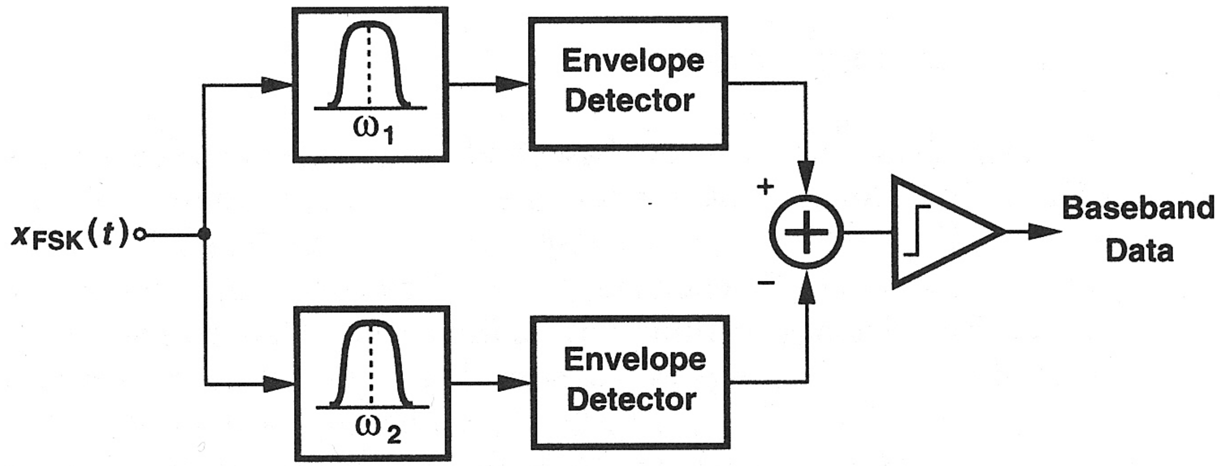
- The above discussion can be generalized to the case of a two-dimensional signal space such as FSK for example
- The correlators serve to find how similar $x_{FSK}(t)$ is to $\cos(\omega_1 t)$ or $\cos(\omega_2 t)$ by integrating $x_{FSK}(t)\cos(\omega_1 t)$ and $x_{FSK}(t)\cos(\omega_2 t)$ over one bit period

Coherent and Noncoherent Detection



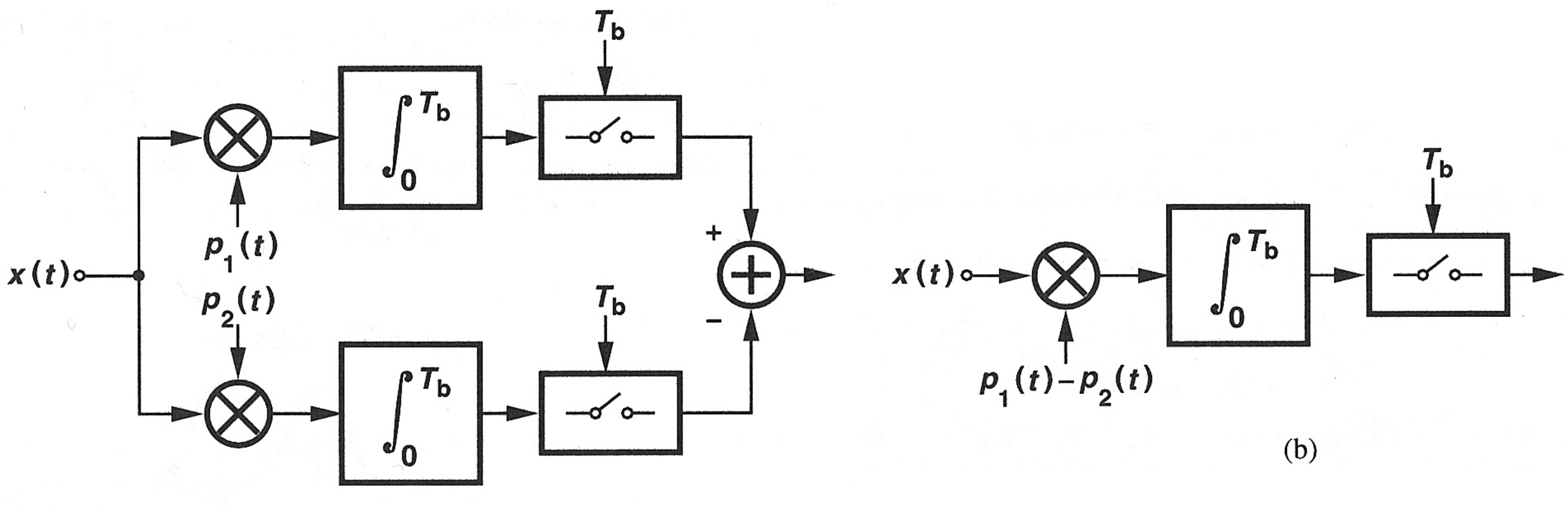
- **Coherent detection** require **phase synchronization** between the received waveform and the one synthesized within the receiver (usually matched filter detection schemes require phase synchronization and are hence coherent detection)
- Can be done by extraction of the carrier from the received waveform (using for example a **Costas loop**)

Coherent and Noncoherent Detection



- Some modulated waveforms can be demodulated using **noncoherent detection**
- A simple example is the FSK detector shown below
- This circuit uses two narrowband filters centered around ω_1 and ω_2 followed by envelope detectors
- Noncoherent detection detectors are more widely used in RF due to their **lower complexity**

Binary Modulation

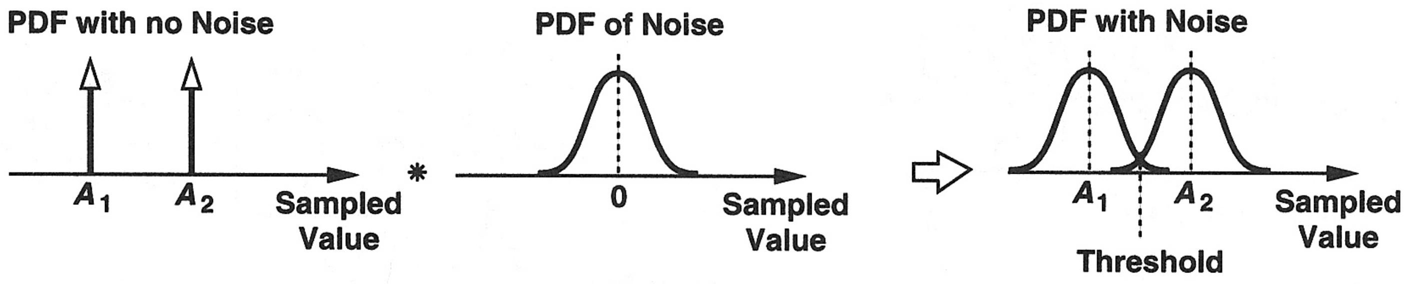


(a)

(b)

- Digital modulation with binary baseband waveforms: ASK (OOK), PSK (BPSK) and FSK (BFSK)
- Binary data represented by two real **pulse shapes** $p_1(t)$ and $p_2(t)$
  - For example in BFSK $p_1(t) = A_c \cos(\omega_1 t)$ and $p_2(t) = A_c \cos(\omega_2 t)$ for $0 < t < T_b$

Bit Error Rate (BER)



PDF=Probability density function

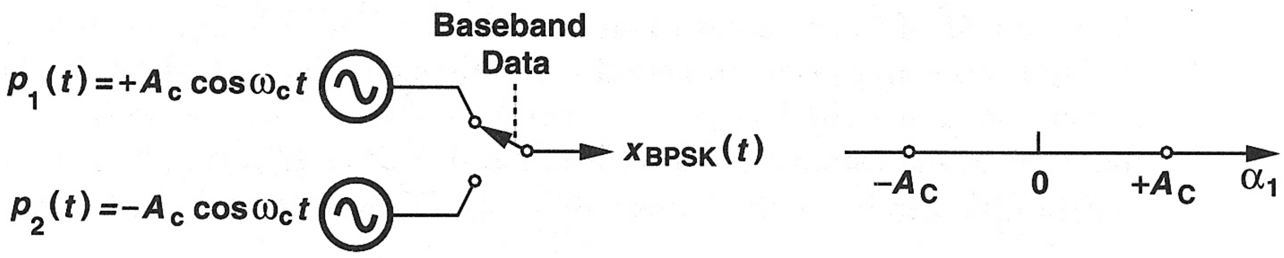
- The max SNR is then given by

$$SNR_{max} = \frac{2E_d}{N_0} \quad \text{with} \quad E_d = \int_{-\infty}^{+\infty} \left| p_1(t) - p_2(t) \right|^2 dt$$

- which is maximum for $p_1(t) = -p_2(t)$
- The bit error rate (BER) assuming additive white gaussian noise (AWGN), ideal matched filter detection and equiprobable states

$$P_e = Q\left(\sqrt{\frac{E_d}{2N_0}}\right) \quad \text{with} \quad Q(x) = \frac{1}{\sqrt{2\pi}} \int_x^{\infty} e^{-\frac{u^2}{2}} du \cong \frac{e^{-\frac{x^2}{2}}}{x\sqrt{2\pi}} \quad \text{for } x > 3$$

Binary Phase Shift Keying (BPSK)



(a) (b)

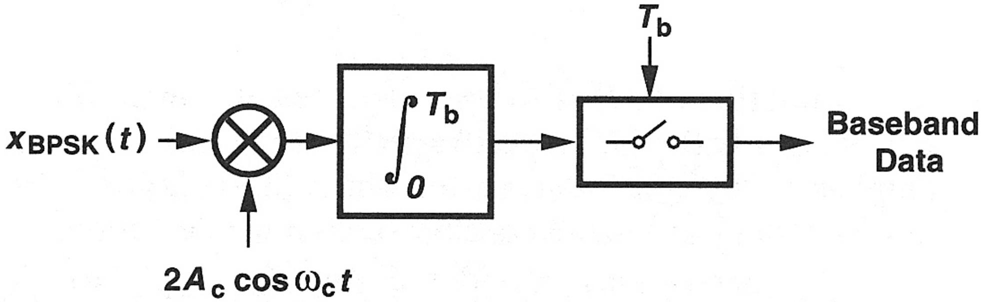
- In BPSK, the binary baseband signal is given by

$$x_{BPSK}(t) = A_c \cdot \cos(\omega_c t + \phi) \quad \text{where} \quad \phi = \begin{cases} 0 & \text{if } b_n = 0 \\ \pi & \text{if } b_n = 1 \end{cases}$$

- which can also be written as

$$x_{BPSK}(t) = \alpha \cdot \cos(\omega_c t) \quad \text{where} \quad \alpha = \begin{cases} +A_c & \text{if } b_n = 1 \\ -A_c & \text{if } b_n = 0 \end{cases}$$

Binary Phase Shift Keying (BPSK)



(c)

- Since $p_1(t) = -p_2(t)$, the correlating signal in the detector is then

$$p_1(t) - p_2(t) = 2p_1(t) = 2A_c \cdot \cos(\omega_c t)$$

BPSK – BER

- The signal energy E_d of a BPSK signal is given by

$$E_d = \int_{-\infty}^{+\infty} \left|p_1(t) - p_2(t)\right|^2 dt = \int_0^{T_b} \left(2A_c \cdot \cos(\omega_c t)\right)^2 dt = 2A_c^2 T_b$$

- and hence the BER is then given by

$$P_e = Q\left(\sqrt{\frac{E_d}{2N_0}}\right) = Q\left(\sqrt{\frac{A_c^2 T_b}{N_0}}\right)$$

- To make fair comparisons with other modulation schemes, we define the average **energy per bit** E_b as

$$E_b = \frac{A_c^2 T_b}{2} \quad \text{and hence} \quad P_{e,BPSK} = Q\left(\sqrt{\frac{2E_b}{N_0}}\right)$$

- Note that E_b contains both the amplitude and the period of the signal, the error rate can be lowered by increasing the signal power (i.e. increasing A_c) or decreasing the data rate (i.e. increasing T_b)

Binary Frequency Shift Keying (BFSK)

- In BFSK, the binary baseband signal is given by

$$x_{FSK}(t) = \alpha_1 \cdot \cos(\omega_1 t) + \alpha_2 \cdot \cos(\omega_2 t) \quad \text{where} \quad [\alpha_1 \quad \alpha_2] = [0 \quad A_c] \quad \text{or} \quad [A_c \quad 0]$$

- For the two basis functions to be orthogonal over one bit period, we need

$$\int_0^{T_b} \cos(\omega_1 t) \cdot \cos(\omega_2 t) \cdot dt = 0$$

- Assuming $\omega_1 + \omega_2 \gg \omega_1 - \omega_2$, the above equation reduces to

$$\frac{\sin((\omega_1 - \omega_2) \cdot T_b)}{\omega_1 - \omega_2} = 0 \quad \text{thus} \quad (\omega_1 - \omega_2) \cdot T_b = n\pi$$

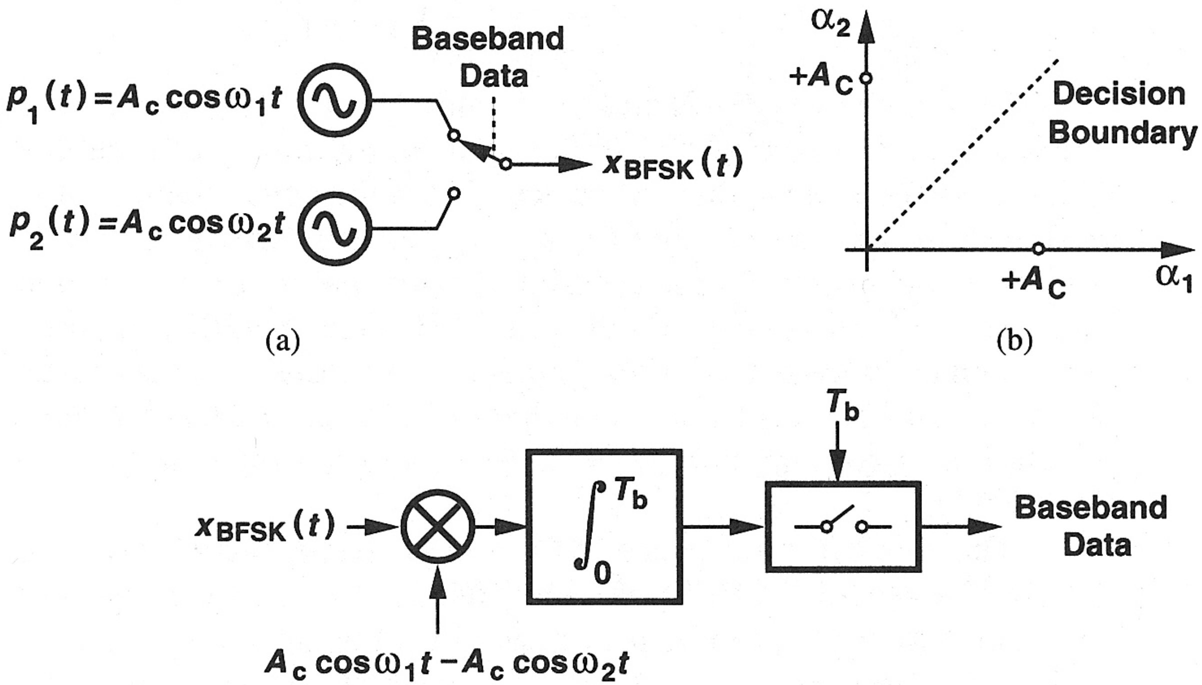
- For minimum spacing between ω_1 and ω_2

$$(\omega_1 - \omega_2) \cdot T_b = \pi \quad \text{or} \quad f_1 - f_2 \triangleq \Delta f = \frac{1}{2T_b}$$

- The FSK signal bandwidth is approximately given by Carson's rule

$$B_T \cong 2\left(\Delta f + \frac{1}{T_b}\right) \quad \text{for orthogonal FSK} \quad B_T \cong \frac{3}{T_b}$$

BFSK



(a) (b) (c)

- BFSK is 2-dimensional scheme with

$$p_1(t) = A_c \cdot \cos(\omega_1 t) \quad \text{and} \quad p_2(t) = A_c \cdot \cos(\omega_2 t)$$

- The corresponding constellation is shown above

BFSK – BER

- Since $p_1(t)$ and $p_2(t)$ are orthogonal

$$E_d = \int_{-\infty}^{+\infty} \left(p_1^2(t) + p_2^2(t) \right) dt = A_c^2 T_b$$

- and hence the BER is then given by

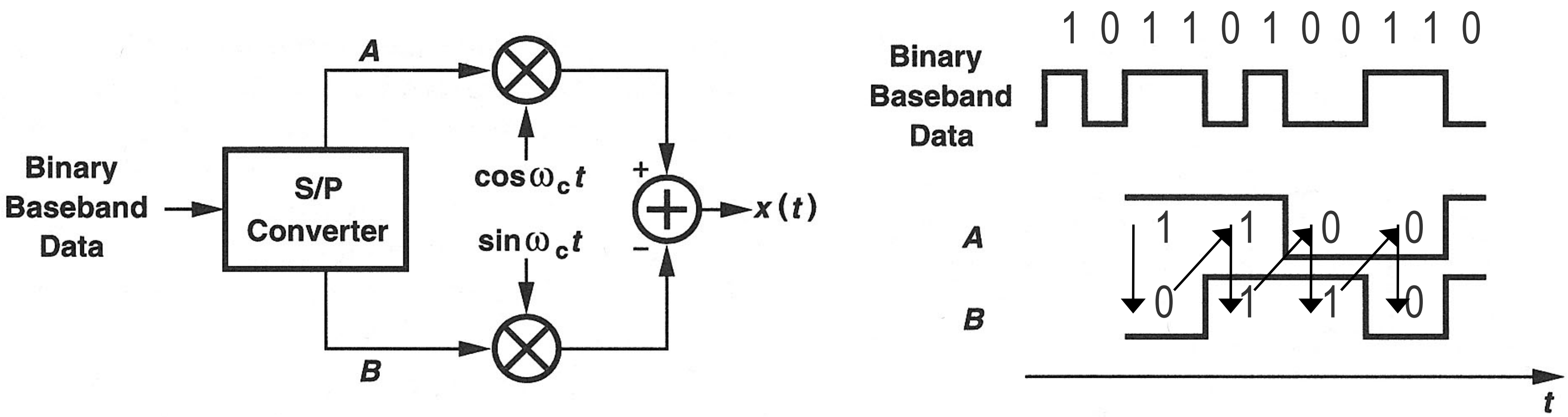
$$P_e = Q\left(\sqrt{\frac{A_c^2 T_b}{2N_0}} \right)$$

- As in BFSK the average energy per bit is given by

$$E_b = \frac{A_c^2 T_b}{2} \quad \text{and hence} \quad P_{e,BFSK} = Q\left(\sqrt{\frac{E_b}{N_0}} \right)$$

- For the same BER, the bit energy in BFSK is twice that in BPSK (3dB advantage of BPSK over BFSK)
- Nevertheless BFSK is widely used for low data rate applications where E_b can be maximized by allowing a long T_b

Quadrature Modulation – Principle

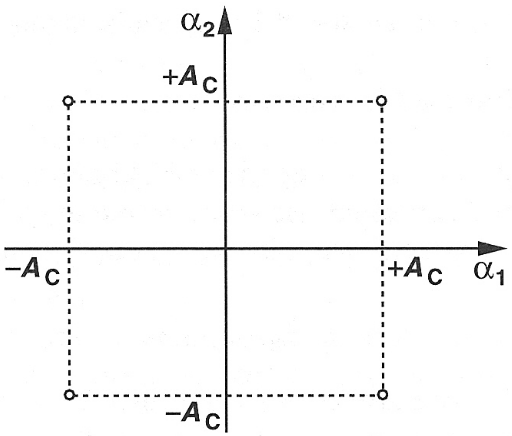


- In many applications it is beneficial to subdivide a binary data stream into pairs of two bits and represent each pair with one of four levels performing modulation
- For example bits b_m and b_{m+1} can be impressed upon a single carrier as

$$x(t) = b_m A_c \cos\left(\omega_c t\right) - b_{m+1} A_c \sin\left(\omega_c t\right)$$

- This is possible because $\cos(\omega_c t)$ and $\sin(\omega_c t)$ are orthogonal

Quadrature Modulation – Constellation

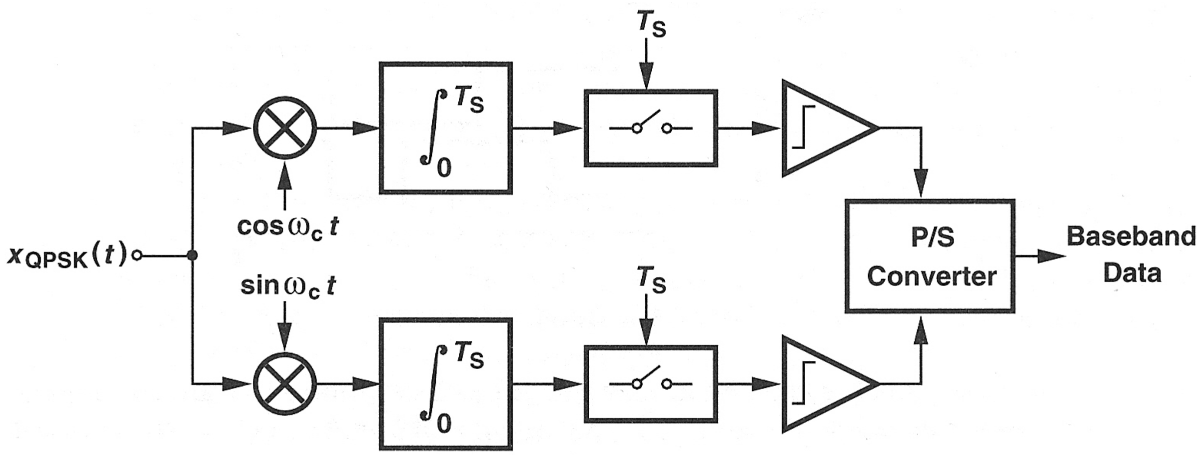


- Assuming b_m and b_{m+1} are rectangular pulses with height ±1

$$x(t) = \alpha_1 \cos(\omega_c t) + \alpha_2 \sin(\omega_c t)$$

- Where α_1 and α_2 can take on a value of $+A_c$ or $-A_c$
- Two broad categories: **quadrature phase shift keying** (QPSK) and **minimum shift keying** (MSK)
- QPSK includes offset QPSK (OQPSK) and π/4-QPSK and MSK includes Gaussian MSK (GMSK)

Quadrature Phase Shift Keying (QPSK)

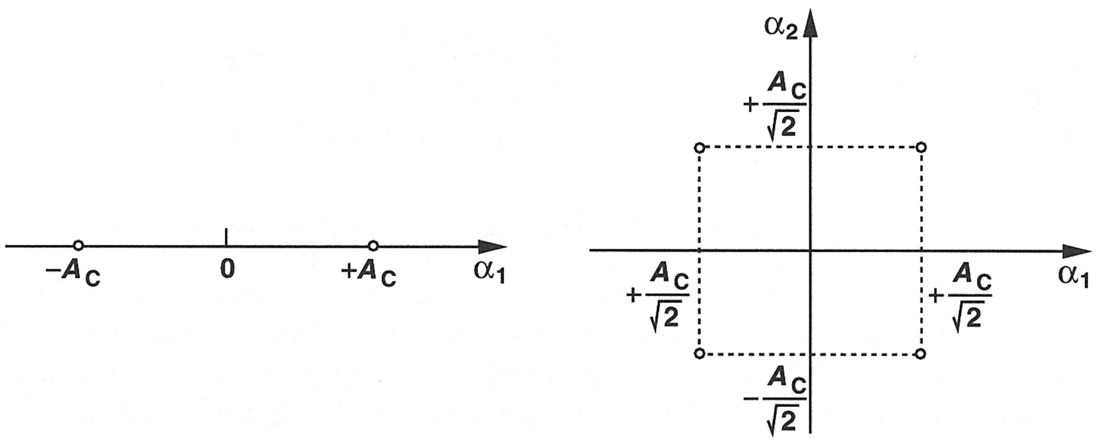


- In analogy with BPSK, QPSK uses one of four phases of a sinusoid according to the symbol

$$x(t) = \sqrt{2} A_c \cos\left(\omega_c t + k\frac{\pi}{4}\right) \quad \text{with } k = 1,3,5,7$$

- Coherent detection can be performed as shown below
- Determines the most likely value of the symbol from the set [±1 ±1]

QPSK – Constellation and BER



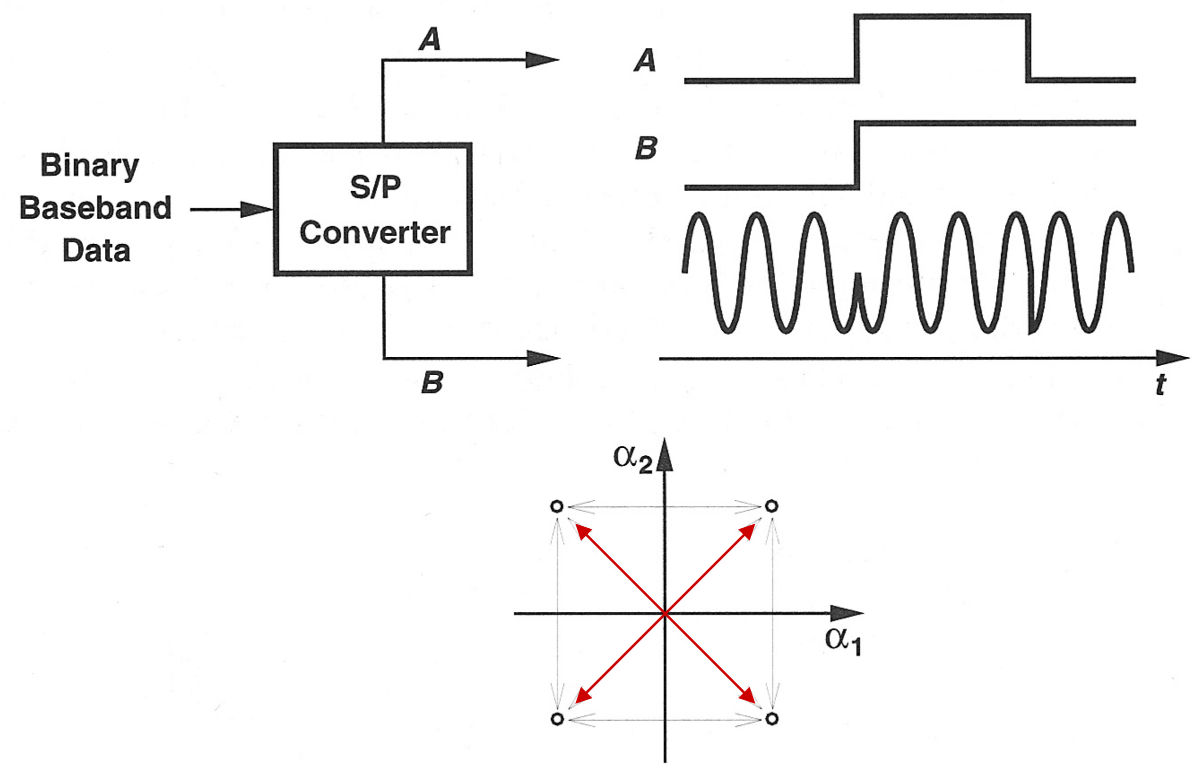
QPSK seems to have a higher BER because the points are located closer to each other

- For fair comparison between BPSK and QPSK it should be assumed that the transmitters produce the same average output power of $A_c^2/2$

$$x_{BPSK}(t) = \pm A_c \cos(\omega_c t) \quad \text{and} \quad x_{QPSK}(t) = A_c \cos\left(\omega_c t + k\frac{\pi}{4}\right) = \pm\frac{A_c}{\sqrt{2}}\cos(\omega_c t) \pm \frac{A_c}{\sqrt{2}}\sin(\omega_c t)$$

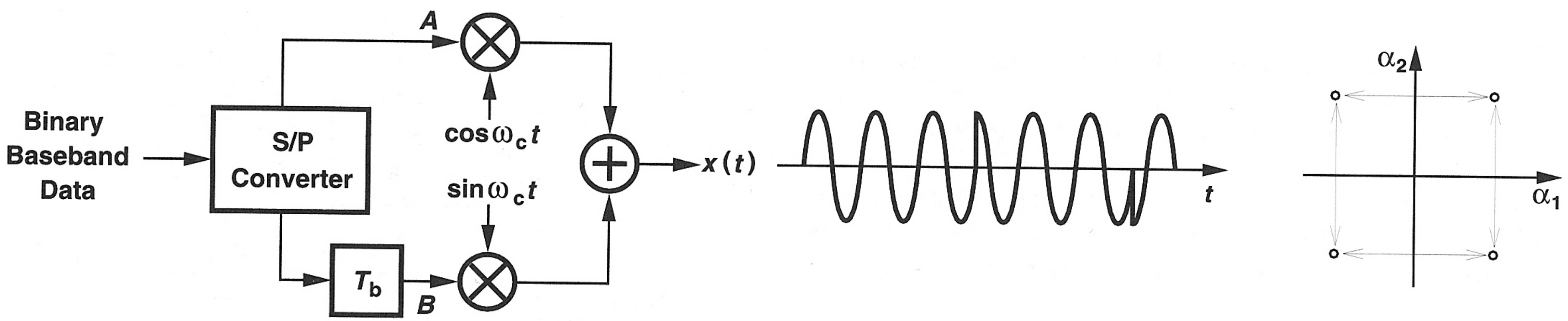
- Fair comparison also requires identical bit rates (identical binary data period T_b)
- Bit energy of BPSK is half the symbol energy of QPSK
- BPSK and QPSK have nearly the same BER $P_{e,QPSK} = P_{e,BPSK} = Q\left(\sqrt{\frac{2E_b}{N_0}}\right)$

QPSK – Phase Changes



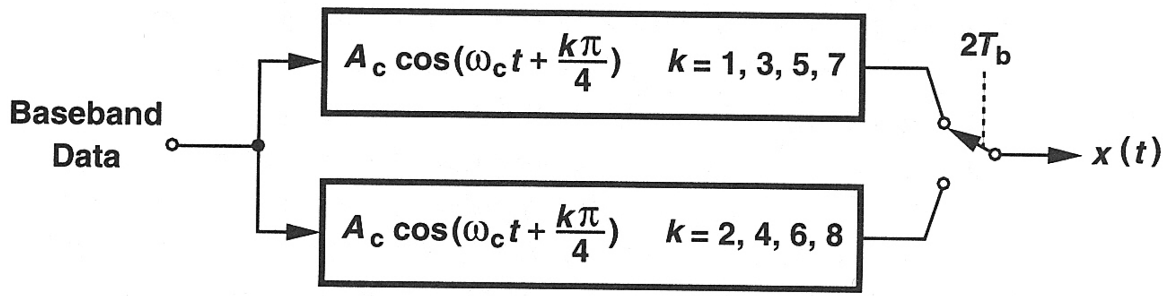
- Abrupt phase changes can occur in QPSK
- They are undesirable since they generate envelope variation after filtering (requires linear PA)

Offset-QPSK (OQPSK) – Principle



- The phase change of 180° of QPSK can be reduced to 90° by introducing a time offset of ½ symbol period between the two data streams
- OQPSK does not allow for differential encoding (encoding the phase change) which is required for noncoherent detection

π/4-QPSK – Principle



- In contrast to OQPSK, π/4-QPSK allows for differential encoding
- The π/4-QPSK signal consist of two QPSK schemes, one rotated by π/4

$$x_1(t) = A_c \cos\left(\omega_c t + k\frac{\pi}{4}\right) \quad \text{for } k \text{ odd,}$$

$$x_2(t) = A_c \cos\left(\omega_c t + k\frac{\pi}{4}\right) \quad \text{for } k \text{ even.}$$

π/4-QPSK – Generation

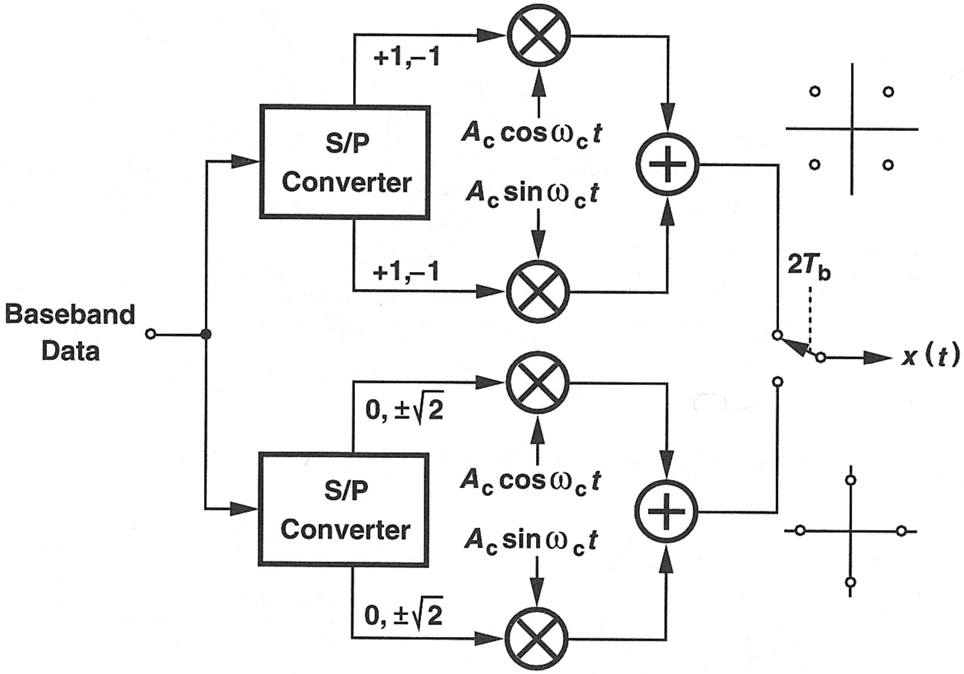


Figure 3.40 Generation of $\pi/4$-QPSK signals.

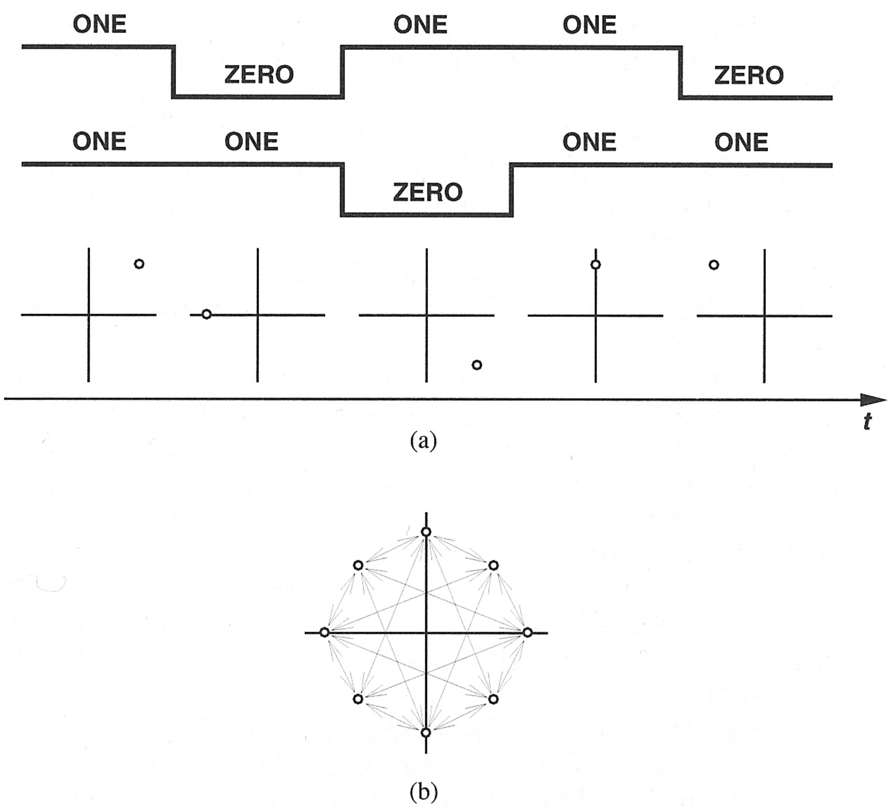


Figure 3.41 (a) Evolution of $\pi/4$-QPSK in time domain, (b) possible phase

Raised-Cosine Pulse Shaping

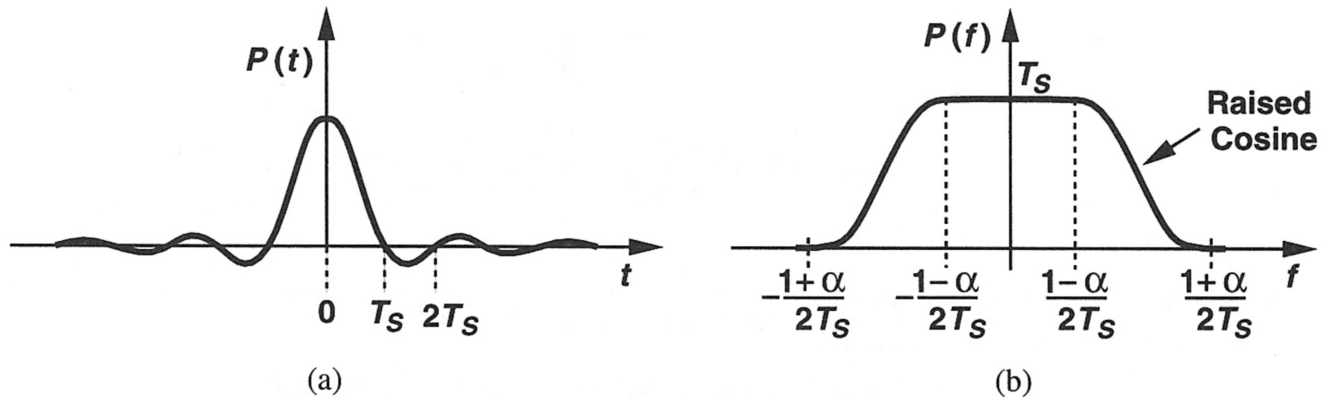


Figure 2.16 Raised-cosine pulse: (a) in time domain, (b) in frequency domain.

- It was assumed that the baseband bits are represented by rectangular pulses
- For **better spectrum efficiency**, more band-limited pulses should be used (such as **raised-cosine**)

$$p(t) = \frac{\sin(\pi t/T_S)}{\pi t/T_S} \cdot \frac{\cos(\pi \alpha t/T_S)}{1-(2\alpha t/T_S)^2}$$

$$P(f) = \begin{cases} T_S & 0 < |f| < \frac{1-\alpha}{2T_S} \\ \frac{T_S}{2}\left[1+\cos\left(\frac{\pi T_S}{\alpha}\left(|f| - \frac{1-\alpha}{2T_S}\right)\right)\right] & \frac{1-\alpha}{2T_S} < |f| < \frac{1+\alpha}{2T_S} \\ 0 & \frac{1+\alpha}{2T_S} < |f| \end{cases}$$

Raised-Cosine Pulse Shaping

- Raised-cosine signaling can be obtained by filtering each bit (represented by a Dirac impulse) by a filter having an impulse response equal to $p(t)$

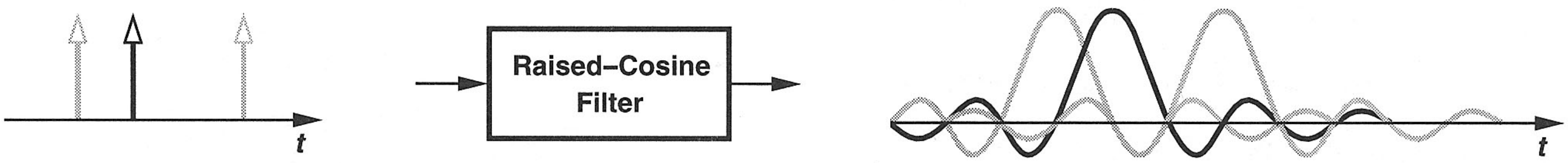


Figure 2.17 Raised-cosine filtering.

- In practice the raised-cosine filter is decomposed into two sections one placed in the transmitter and the other in the receiver

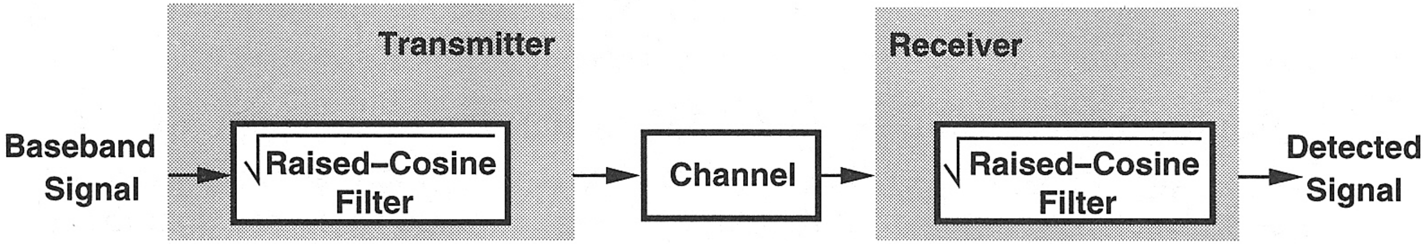
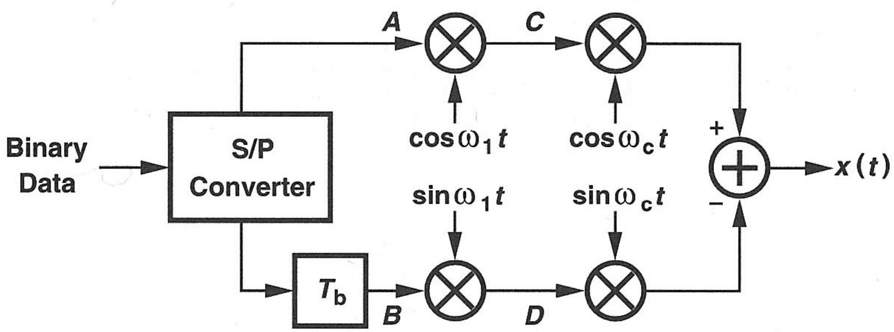
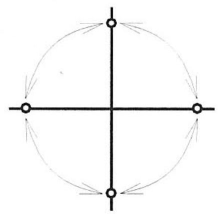
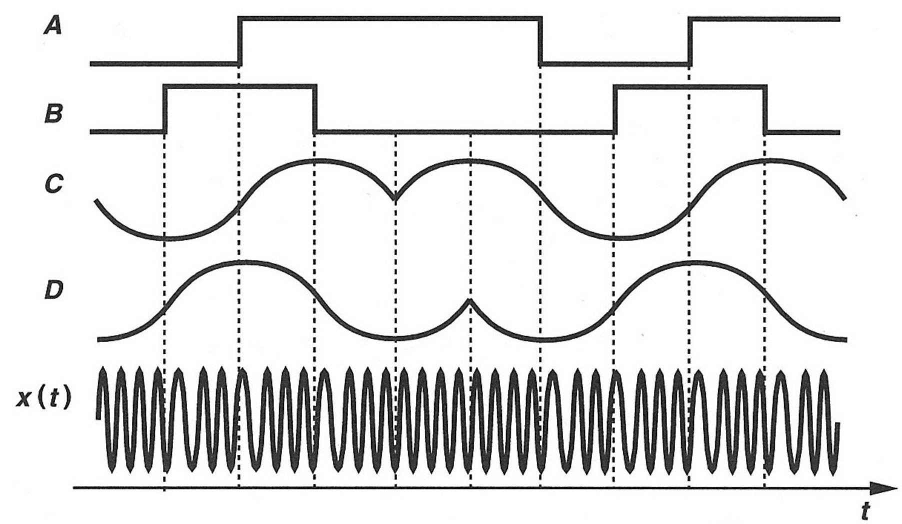


Figure 3.42 Decomposition of a raised-cosine filter into two sections.

Minimum Shift Keying (MSK)





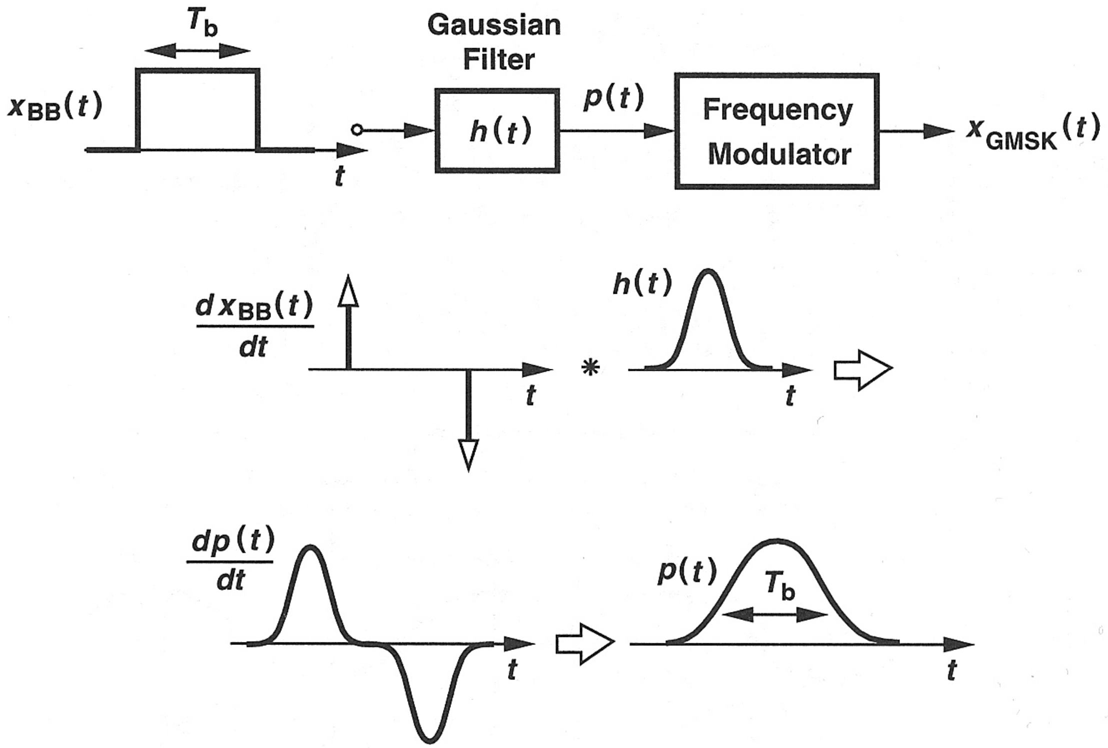


- MSK is continuous phase modulation to avoid sharp phase transitions
- OQPSK with half-cosines instead of rectangular pulses

$$x(t) = a_m \cos(\omega_1 t)\cos(\omega_c t) - a_{m+1} \sin(\omega_1 t)\sin(\omega_c t)$$

- Where a_m and a_{m+1} are rectangular pulses toggling between +1 and −1

Gaussian Minimum Shift Keying (GMSK)



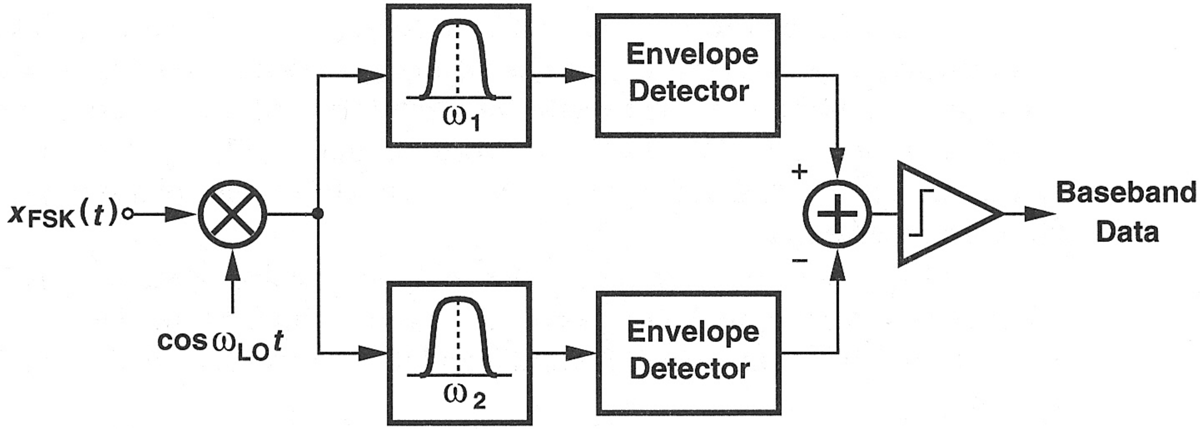
- Pulse shape obtained by passing the baseband rectangular pulse through a Gaussian impulse response

$$h(t) = \exp\left(-\alpha t^2\right)$$

Noncoherent Detection

- Matched filters (example of which are **coherent detectors** or **correlators**) provide the highest SNR and hence the **lowest BER**
- Coherent detector require that the phase of the local oscillator (LO) in the receiver is equal to that of the received signal which requires **carrier recovery**
- **Phase alignment** necessitate substantial circuit complexity and becomes especially difficult at low signal levels in the presence of interferers and signal fading
- For this reason many RF systems employ **noncoherent detection** despite its somewhat inferior performance

Noncoherent FSK Detection

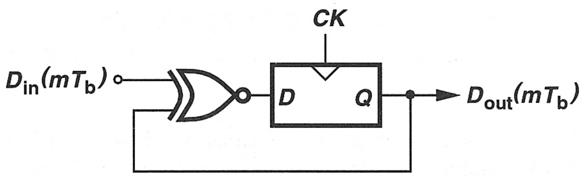


- It can be shown that the BER is given by

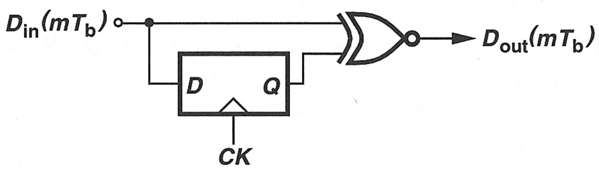
$$P_e = \frac{1}{2}\exp\frac{-E_b}{2T_b B_p N_0} = \frac{1}{2}\exp\frac{-E_b}{2N_0} \quad \text{for } B_p = \frac{1}{T_b}$$

- Where B_p is the bandwidth of each bandpass filter and E_b is the bit energy
- For error rates on the order of 10^{-3} it requires an E_b/N_0 that is only 1.5dB greater than that of an ideal coherent FSK detection

Differential Phase Shift Keying



(a)



(b)

Input Data		0	1	1	1	0	0	1	1	0	1
Encoded Data	1	0	0	0	0	1	0	0	0	1	1
Decoded Data		0	1	1	1	0	0	1	1	0	1

(c)

$$P_e = \frac{1}{2}\exp\frac{-E_b}{N_0} \quad \text{where } E_b = \frac{A_c^2 T_b}{2}$$

- Simple PSK waveforms cannot be detected noncoherently (requires time origin)
- However noncoherent detection is possible if the information is coded in the phase change from one bit (or symbol) to the next (does not require time origin)
- Differential PSK (DPSK) $D_{out}\left[(m+1)T_b\right] = \overline{D_{in}(mT_b) \oplus D_{out}(mT_b)}$
- For BER=10^{-3} DPSK requires approximately 3dB higher SNR than coherent PSK

Outline

- General Considerations
- Analog Modulation
- Digital Modulation
- **Power Efficiency of Modulation Schemes**
- Noncoherent Detection

Power Efficiency of Modulation Schemes

- Modulated waveform $x(t) = A(t)\cos\left(\omega_c t + \varphi(t)\right)$
- Constant envelope if $A(t) = const.$, otherwise variable envelope
  - Can define peak to average power ratio (PAPR) for variable envelope signals
- 3rd-order memoryless nonlinearity with $A(t) = A_c$

$$
\begin{aligned}
y(t) &= \alpha_3 x^3(t) + \ldots \\
&= \alpha_3 A_c^3 \cos^3\left(\omega_c t + \varphi(t)\right) + \ldots \\
&= \frac{\alpha_3 A_c^3}{4}\cos\left(3\omega_c t + 3\varphi(t)\right) + \frac{3\alpha_3 A_c^3}{4}\cos\left(\omega_c t + \varphi(t)\right)
\end{aligned}
$$

- For BER=10^{-3} DPSK requires approximately 3dB higher SNR than coherent PSK
- Bandwidth of original signal typically much smaller than ω_c
- From Carson's rule, the bandwidth occupied by third harmonic remains small

Spectral Regrowth

- Consider a variable-envelope signal applied to the same nonlinear system

$$x(t) = I(t)\cos(\omega_c t) + Q(t)\sin(\omega_c t)$$

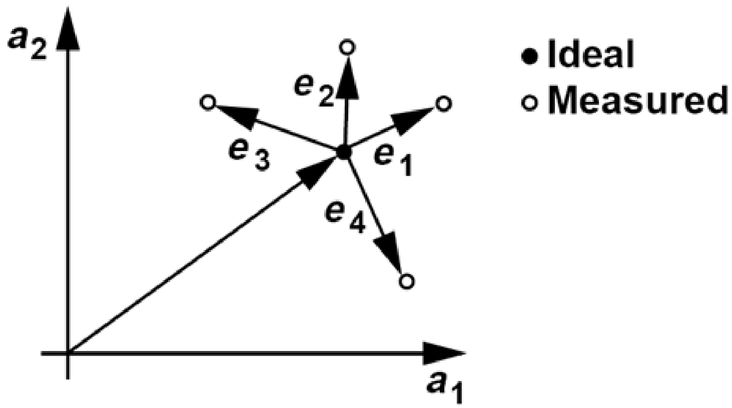
- where $I(t)$ and $Q(t)$ are the baseband in-phase and quadrature components

$$y(t) = \alpha_3 \left[I(t)\cos(\omega_c t) + Q(t)\sin(\omega_c t) \right]^3$$

$$= \alpha_3 I^3(t) \frac{\cos(3\omega_c t) + 3\cos(\omega_c t)}{4} - \alpha_3 Q^3(t) \frac{-\cos(3\omega_c t) + 3\sin(\omega_c t)}{4} + \ldots$$

- Spectrum contains the spectra of $I^3(t)$ and $Q^3(t)$ centered around ω_c
- Since these components generally exhibit a broader spectrum than do $I(t)$ and $Q(t)$ the spectrum grows when a variable envelope signal passes through a nonlinear system

Error Vector Magnitude



$$\mathrm{EVM}_1 = \frac{1}{V_{rms}}\sqrt{\frac{1}{N}\sum_{j=1}^{N} e_j^2},$$

$$\mathrm{EVM}_2 = \frac{1}{P_{avg}} \cdot \frac{1}{N}\sum_{j=1}^{N} e_j^2,$$

- Transmitter nonlinearity and other non-ideal effects cause the constellation points to deviate from their ideal values
- **Error vector magnitude (EVM)** provides a quantitative measure of impairments that corrupt the transmitted signal

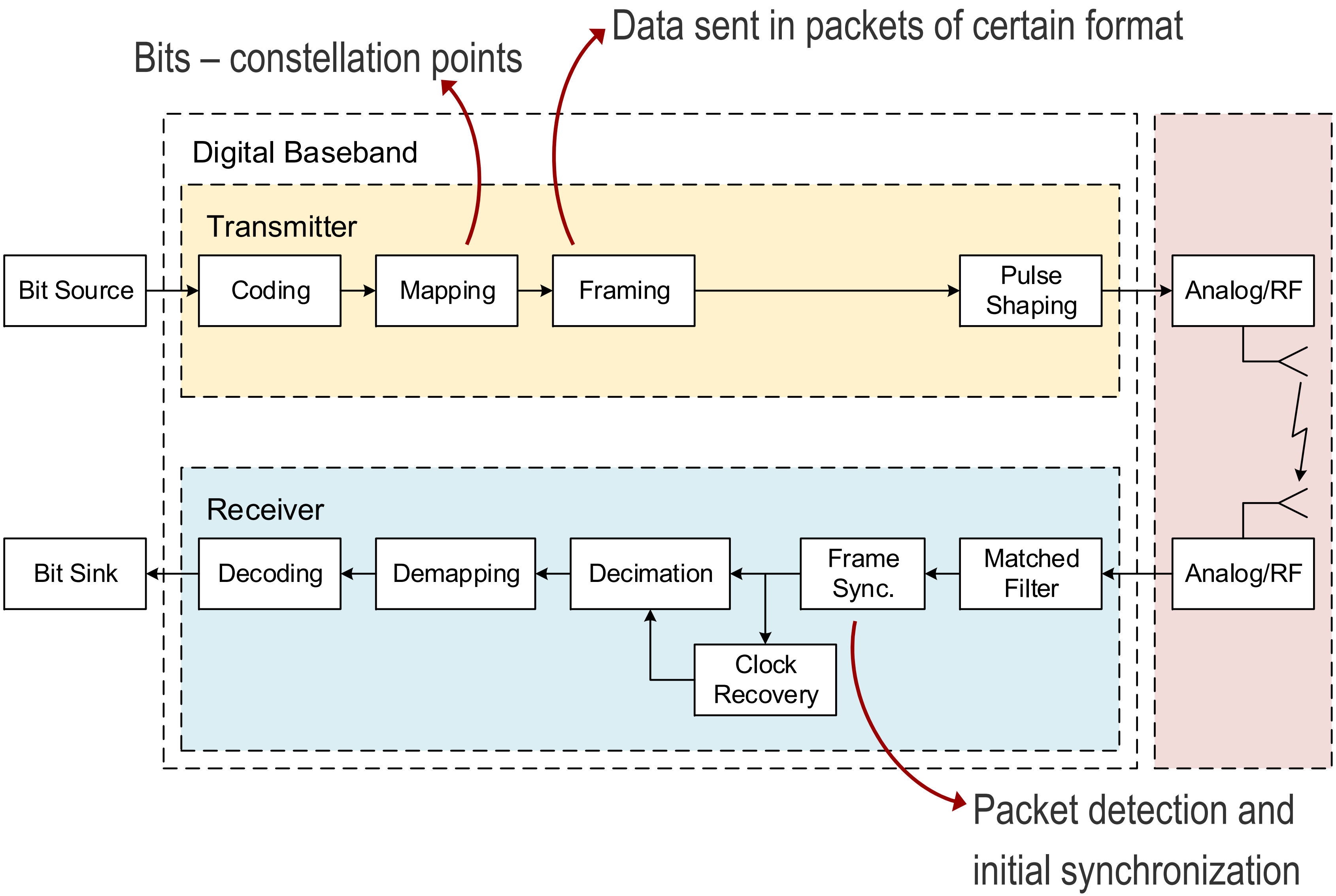
Outline

- General Considerations
- Analog Modulation
- Digital Modulation
- Power Efficiency of Modulation Schemes
- **Noncoherent Detection**

Performance of Modulation Techniques

- The performance of a modulation technique for RF applications is determined by three parameters:
  - The signal-to-noise ratio (**SNR**) or bit-error-rate (**BER**) in case of digital modulation, defining the quality of the modulation scheme
  - The **spectral efficiency** measuring the bandwidth required by the modulation scheme
  - The **power efficiency** depending on the type of power amplifier required by the modulation technique
  - Some modulated waveforms (constant envelope) can be processed by means of a nonlinear amplifier whereas some other require linear amplifier

Generic Transceiver



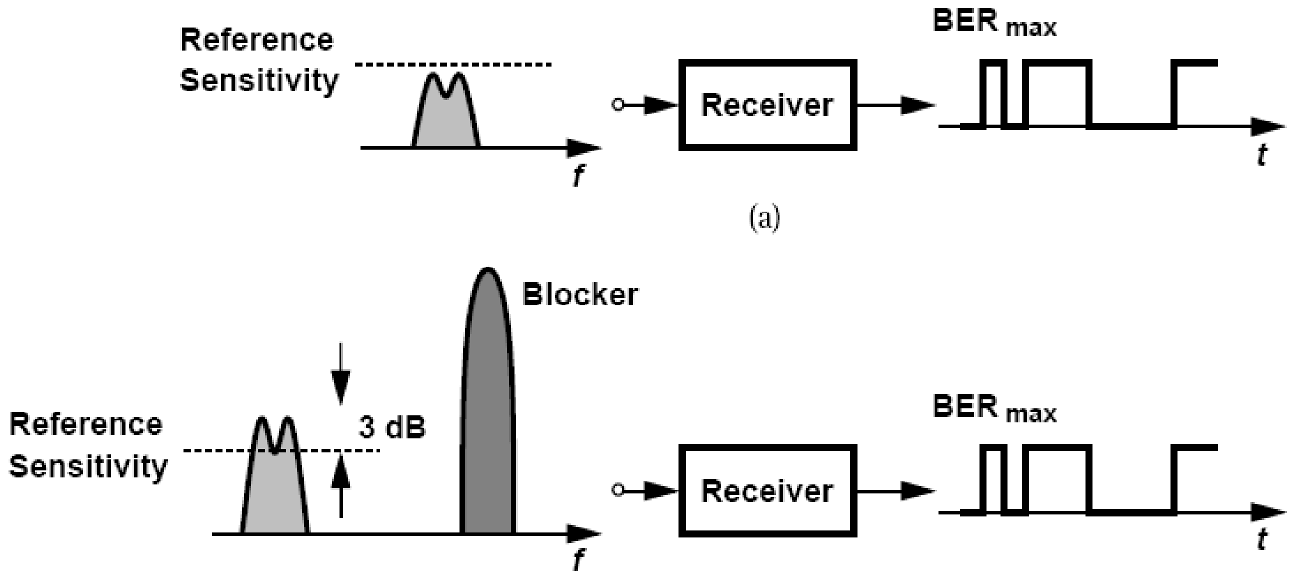
Wireless Standards: Specifications

- Frequency bands and channelization
  - Each standard defines frequency bands in which the communication takes place
- Data Rates
  - The standard specifies the data rates that must be supported
- Type of modulation
  - Types of supported modulations as well as pulse shaping are defined by the standard
- Duplexing/Multiplexing Methods
  - Most systems support some kind of frequency (FDD or FDMA) or time (TDD or TDMA) multiple access techniques
  - Some systems employ code division multiple access (CDMA)
    - Direct sequence or frequency hopping

Wireless Standards: Specifications

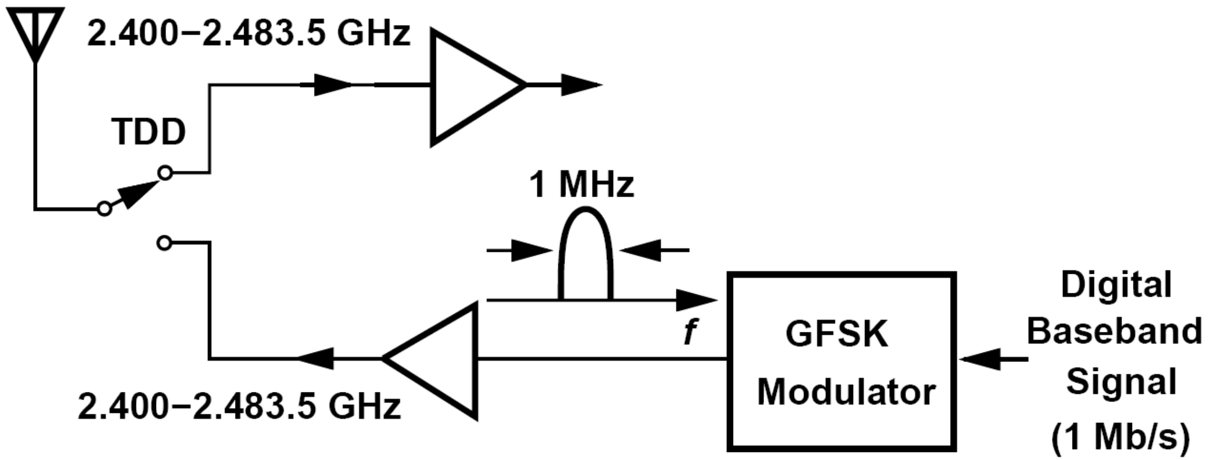
- TX output power
  - Minimum and maximum power of the transmitter defined by the standard (peak and average)
- TX EVM and spectral mask
  - Standard specifies the signal quality that must be respected
  - EVM specifies the deviation from ideal constellation points
  - Spectral mask determines limits to emission in adjacent bands (maximum PSD levels and adjacent channel leakage ratio - ACLR)
- RX sensitivity
  - Minimum detectable signal level defined by the standard in terms of BER
- RX input level range
  - Receiver must be able to handle certain signal range with acceptable noise and distortion

Wireless Standards: Specifications



- RX tolerance to blockers and dynamic range
  - The standard specifies the largest interferer that the RX must be able to tolerate while receiving a weak desired signal
  - Dynamic range of the receiver implicitly determined

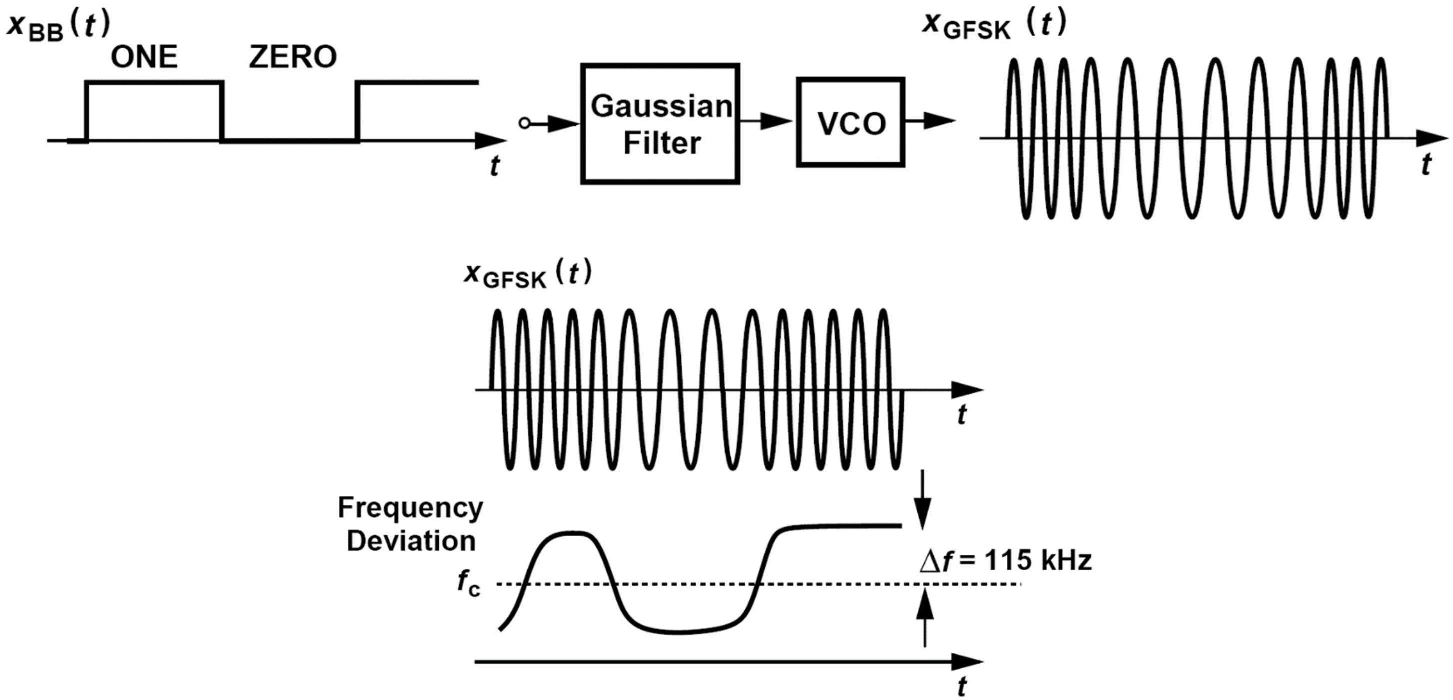
Bluetooth: Air Interface



- 2.4 GHz ISM band
- 1 Mb/s data rate in each channel, 1 MHz wide channels
  - 80 channels
  - Higher data rates supported by newer versions of the standard (BT enhanced data rate)

Bluetooth: Modulation

- Gaussian minimum shift keying (GMSK)



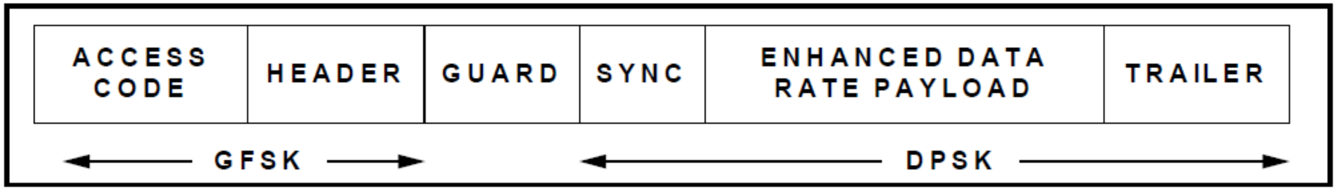
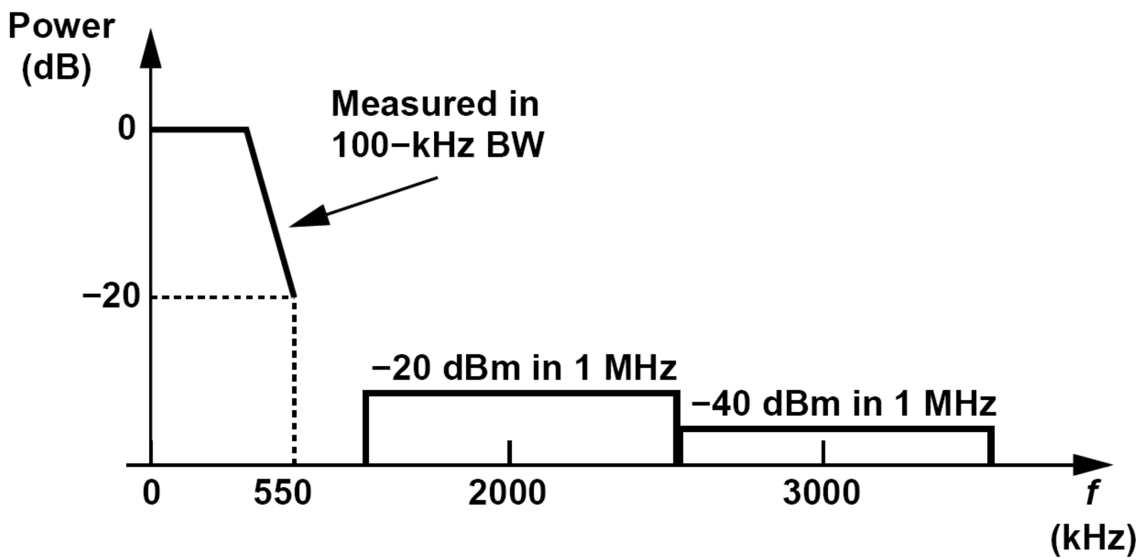


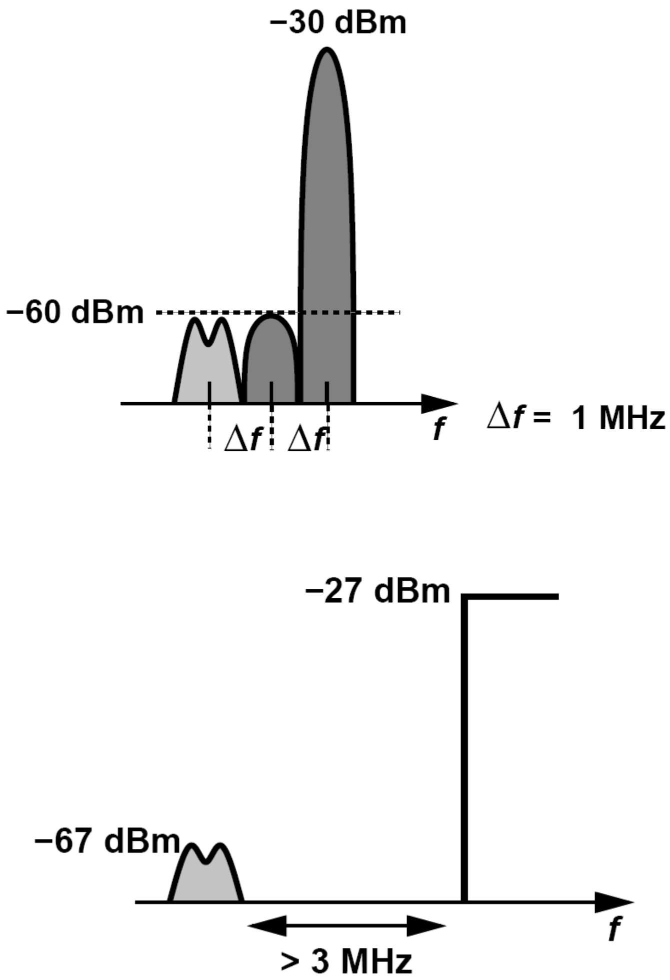
Figure 1.3: Standard Enhanced Data Rate packet format

Bluetooth: Spectral Mask



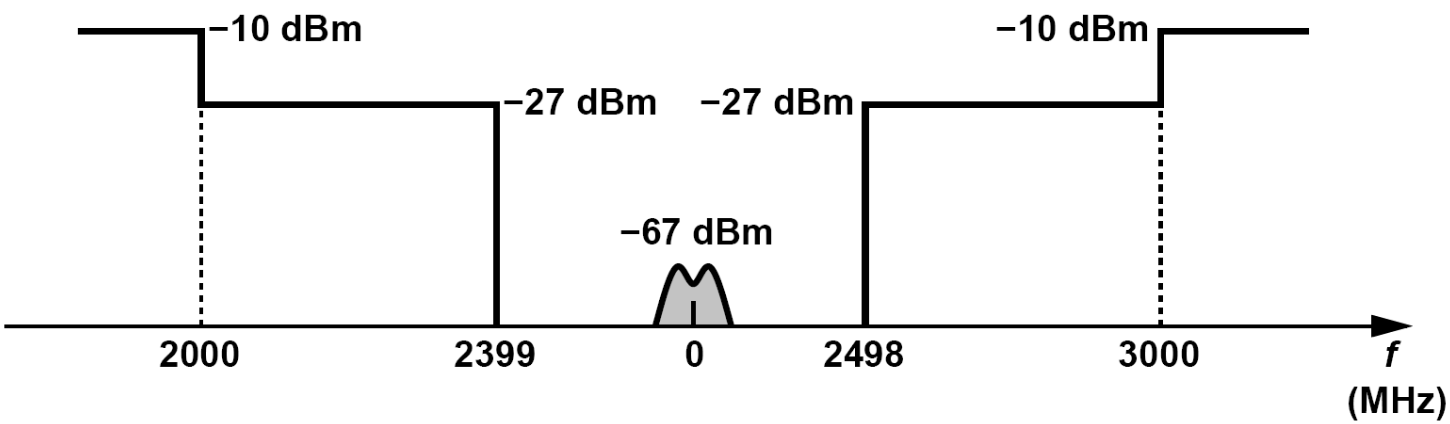
- Different classes of devices supported
- Output power can go up to 20 dBm
- Most commercial devices support output power of 0 dBm
- Receiver must be able to handle input signals up to -20 dBm

Bluetooth: Blockers



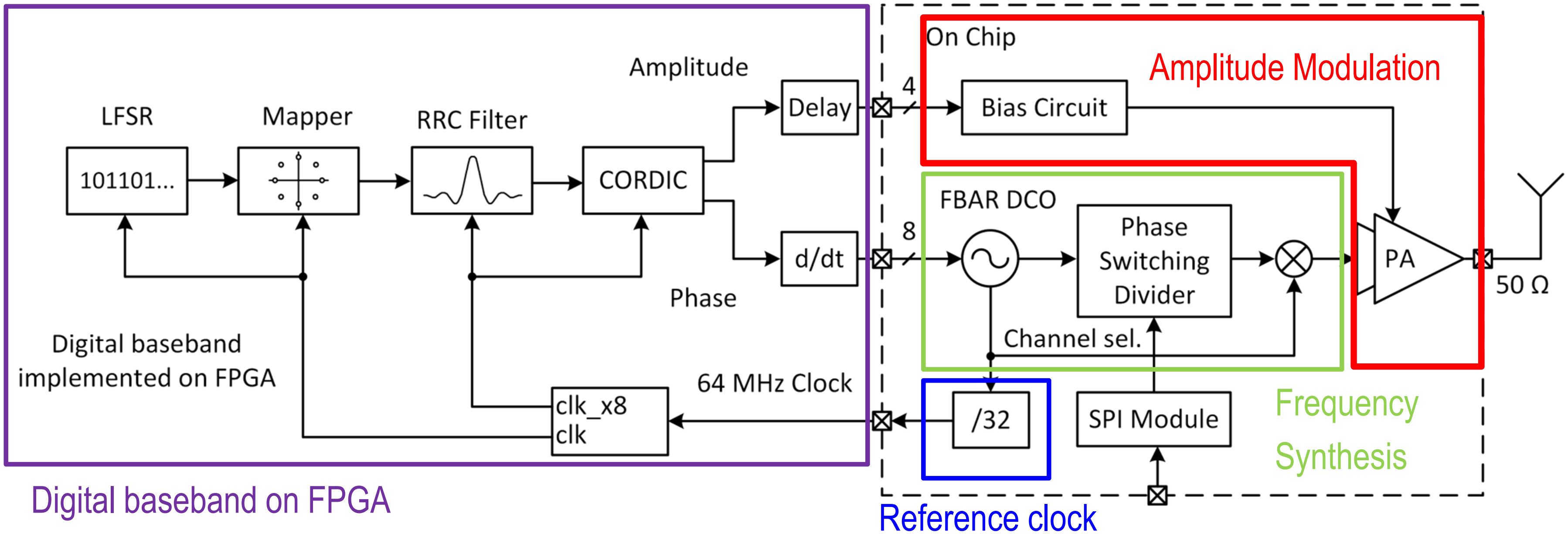
- Reference sensitivity of −70 dBm
  - Most commercial devices in the −90 dBm range
- Blocking test for adjacent and alternate channels:
  - Desired signal 10 dB higher than reference sensitivity. Adjacent channel with equal power modulated. Alternate adjacent channel with −30 dBm modulated.
- Blocking test for third or higher adjacent channel:
  - Desired signal 3 dB above sensitivity level, modulated blocker in third or higher adjacent channel with power −27 dBm

Bluetooth: Blockers



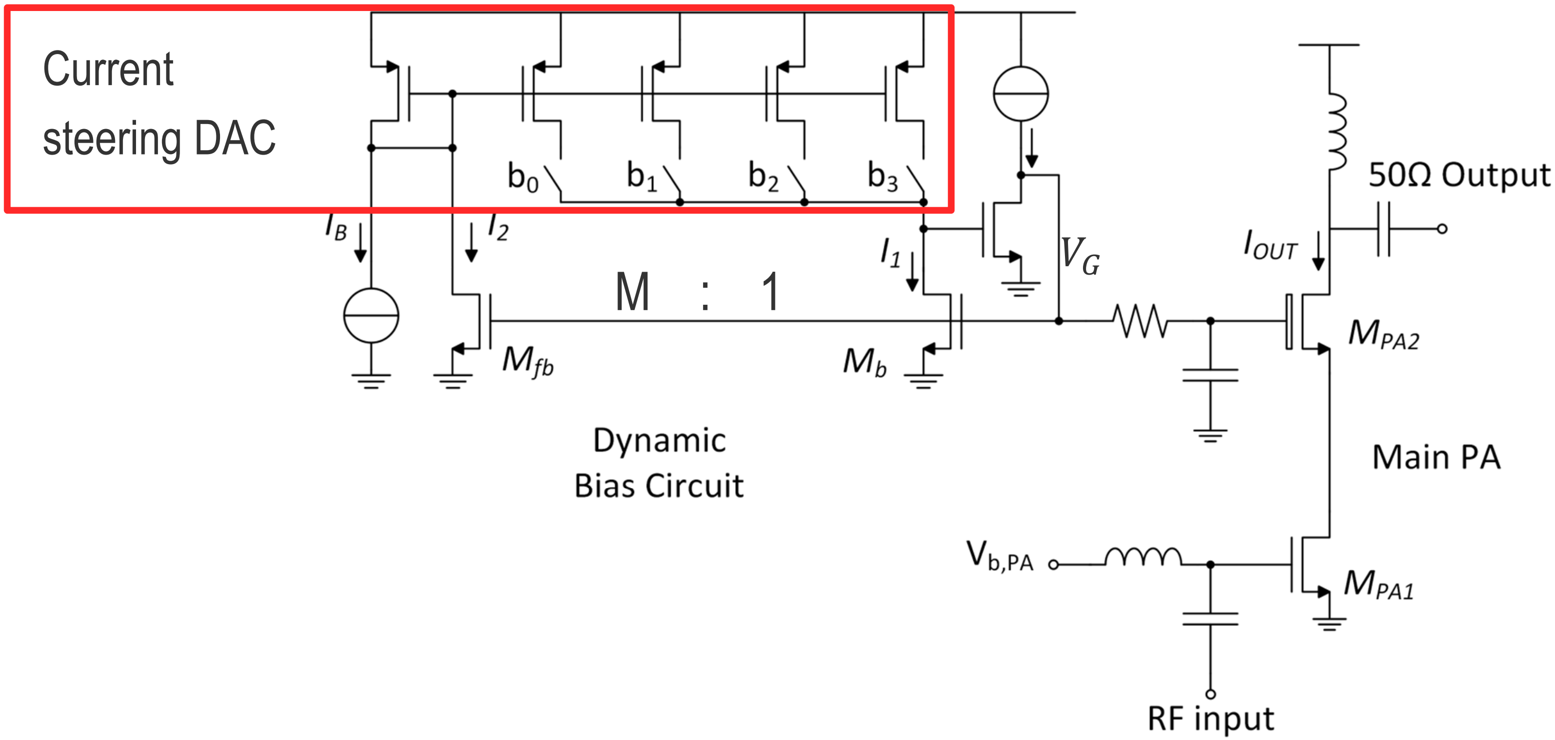
- Reference sensitivity −70 dBm
- Out of band blockers:
  - Desired signal at −67 dBm, tone level of −27 dBm or −10 dBm must be tolerated according to the tone frequency range

Transmitter Example



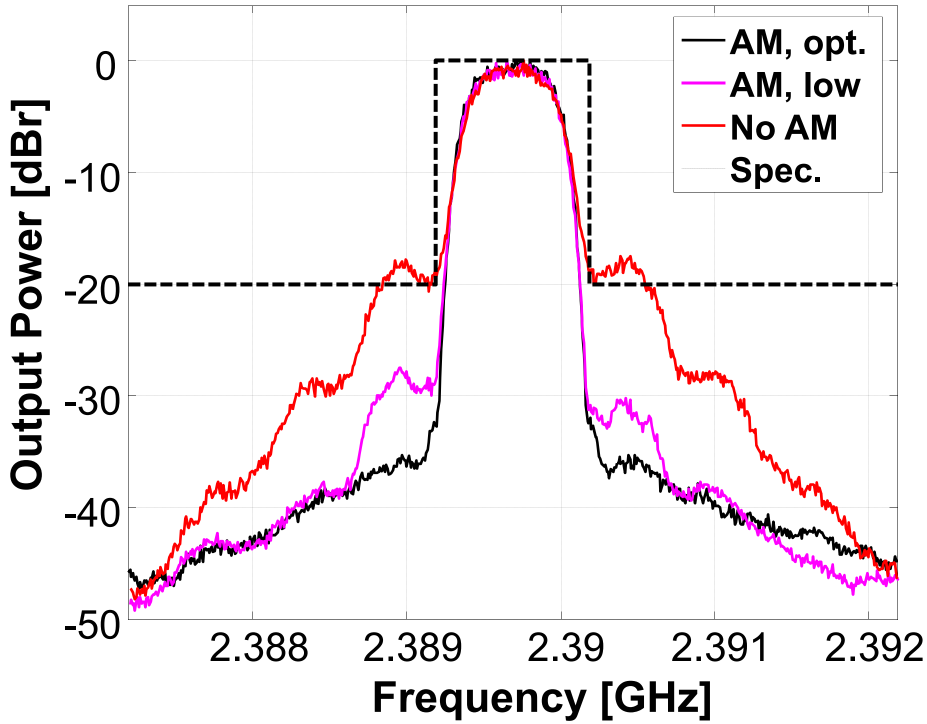
- Transmitter targeting BT and IEEE 802.15.6
- Digital baseband implemented on an FPGA
  - For testing only
  - Spectral mask and EVM requirements
  - No packet formation, continuous stream used

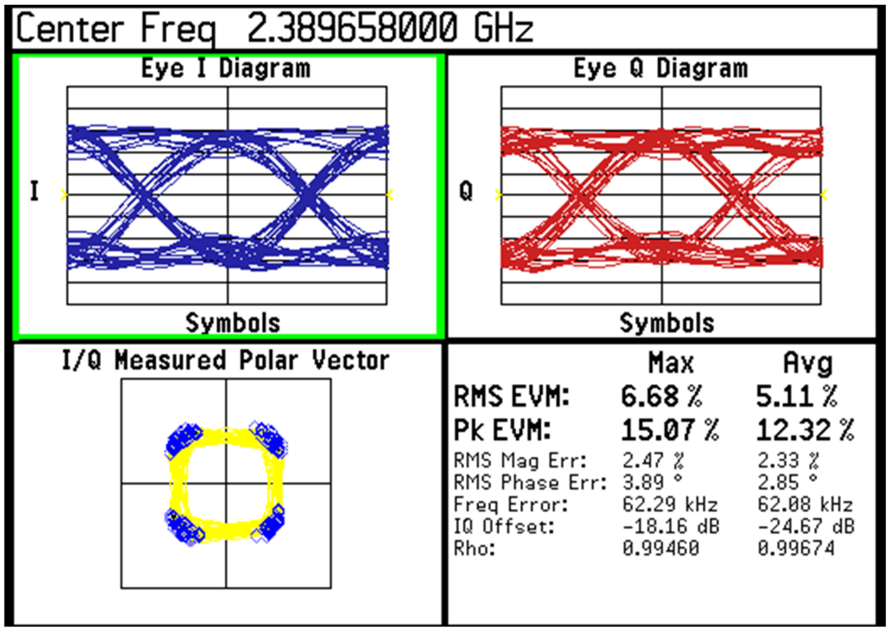
Transmitter Example: Digital-Analog Interface



- Polar architecture – typical in LP transmitters
- Amplitude modulation implemented directly in the power amplifier
- Phase modulated directly at the DCO (digitally controlled oscillator)

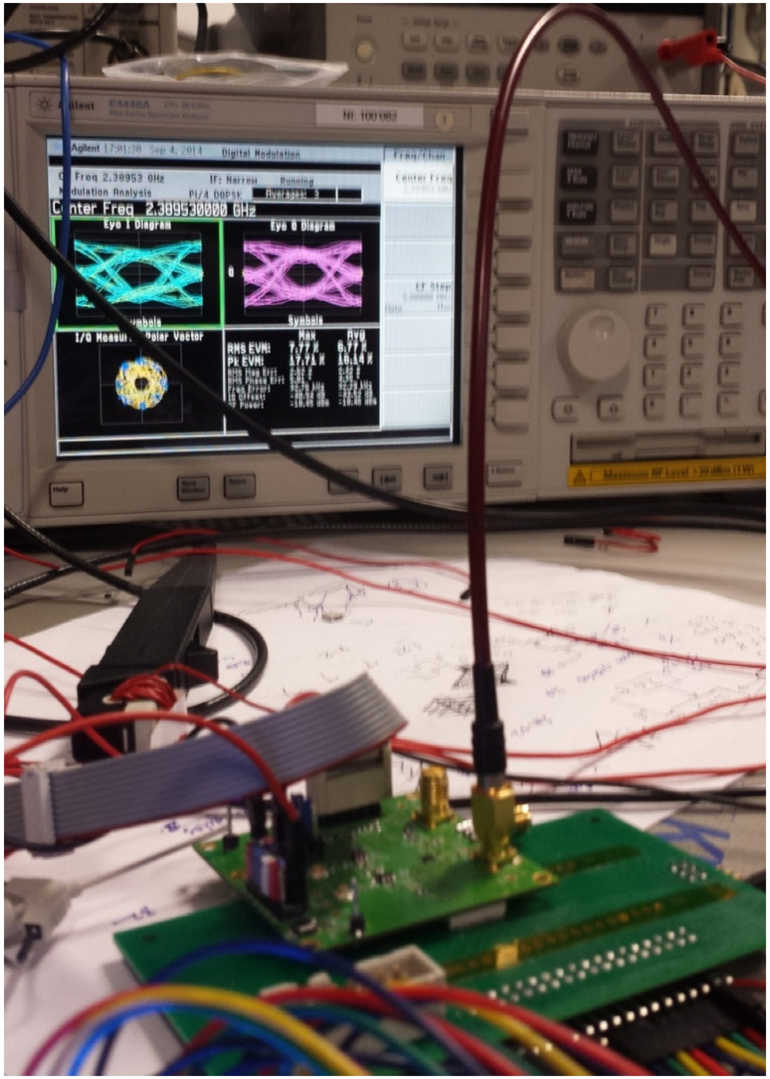
Transmitter Example

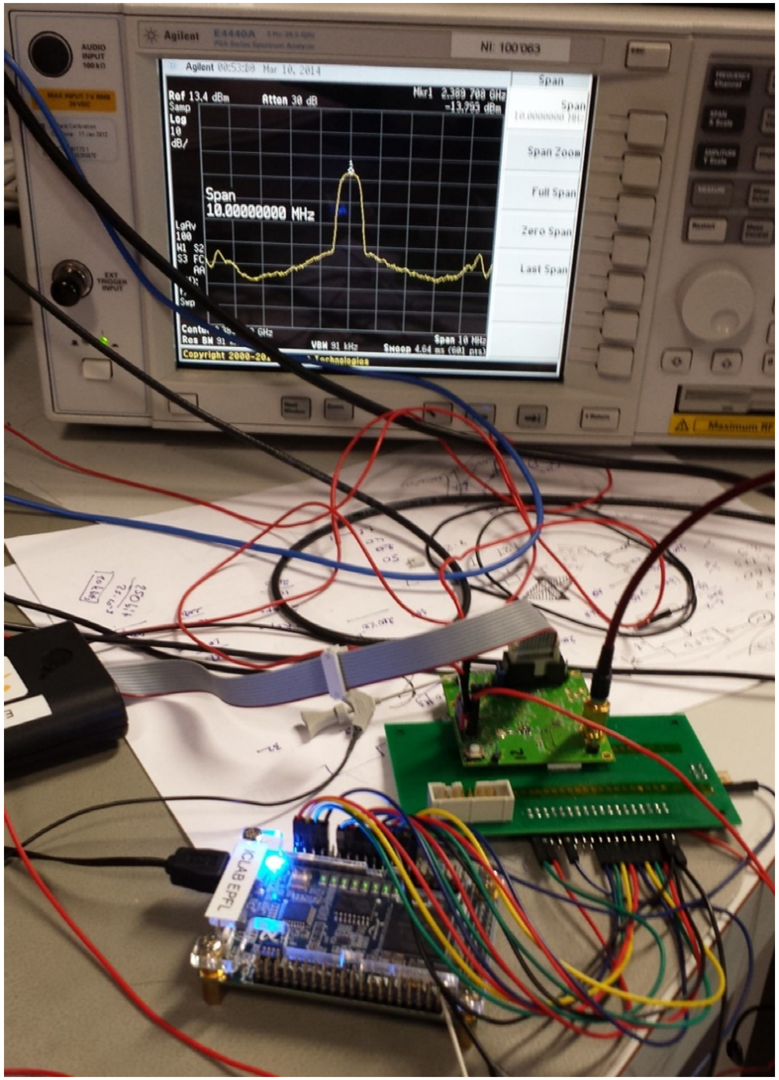




- Spectral mask and EVM measurement

Transmitter Example





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Most of this Chapter is based on Chapter 3 of Reference [1]

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