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Digital Modulation

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Abstract

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- BPSK, QPSK, MSK, properties of gaussian and RRC filtering, concept of ISI
- Spread spectrum, OFDM
- TDMA, FDMA, CDMA definitions
- TDD, FDD
- Constant envelope modulation : Bluetooth, GSM
- Non-constant Envelope Modulation : EDGE, W-CDMA, 802.11b/a
- Transmitter and Receiver key measurements

About the Speaker

Chris Potter received the B.Sc. degree in Electronics in 1983 from the University of York, England and the Ph.D. degree in 1987 from the University of London, England. From 1983 to 1995, he designed a variety of microwave and RF test equipment at Marconi Instruments. From 1995 to 2002, he worked at Tality UK on RF architectures and product designs for GSM, EDGE, Bluetooth, 802.11a/b and W-CDMA. He is presently a consultant with Cambridge RF Ltd. in Cambridge UK. Since 1998, he has been involved with linear PA designs for terminals and cellular infrastructure. His main research interests are in the field of adaptive linearisation. He is also active in RF system designs for terminals and cellular infrastructure, and tools for automation of the RF design process.

Digital Modulation

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IEE Microwave Measurements
Vacation School

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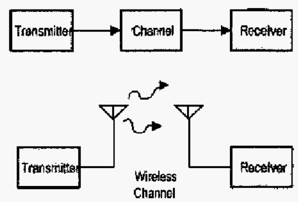

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These days its easy to get lost in a sea of Hi-Tech hype whenever one considers any aspect of information technology and telecommunications. Special buzzwords such as 'broadband', 'information superhighway', 'cable', '3G', 'Wi-Fi', and 'E-commerce' have provided convenient labels for marketing departments, investors, and journalists. However they do not provide much insight into the underlying theory and innovation that have made this technology possible.

Put very simply, a communication system transfers information from a source (transmitter) to a destination (receiver) across a channel. This forms a top level hierarchy for the standard communications system model. Historically much of the information content has contained speech and text, and more recently, data, and integrated multimedia.

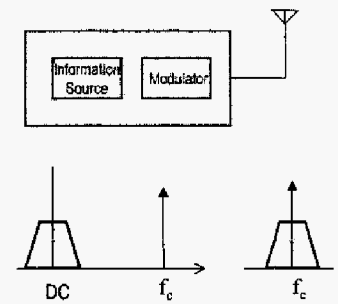
The transmitter processes the raw information content into a format suitable for transmission across the channel. The channel can contain any one of a number of transmission media including: twisted pair line, coaxial cable, optic fibre, and free space. The channel usually has some deleterious effect on the transmitted signal and it is the job of the receiver to restore the received signal to same likeness of the transmitted signal, and to provide an error free recovery of the its information content.

In this course we are predominantly concerned with the fundamentals and practical aspects of digital communications over a wireless channel but much of the content in this first session is also relevant for other types of digital transmission systems.

Information transfer consumes bandwidth and power and a consideration of both is fundamental to the design of wireless communications systems.

What is Modulation Exactly?

- Modulation
  - *One or more parameters of the transmitted signal are varied in proportion to changes in the information signal*
  - *In wireless communications information is transferred to a CW carrier signal.*
    - » Amplitude, Frequency and Phase modulation are used either separately or in conjunction with one another
    - » Information occupies bandwidth in the frequency spectrum about the carrier signal.
    - » The mean frequency of the carrier signal is much higher than the bandwidth of the information signal



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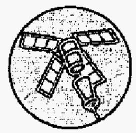
Modulation describes the process of varying one or more parameters of a signal with respect to changes in another signal. In wireless communications the information bearing signal is modulated onto a continuous wave (CW) carrier signal. A variety of modulation techniques involve modulating the amplitude, frequency, and phase of the carrier signal, either separately in combination with one another.

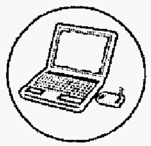
The information, or modulating signal is essentially a "lowpass", or baseband signal. A modulated carrier signal is a "bandpass" signal. The frequency domain diagram shows that the modulation process has shifted the positive and negative frequencies of the baseband signal spectrum up in frequency to be centred about the frequency of the carrier signal. Generally the carrier frequency is much greater than the bandwidth of the modulating signal.

The choice of carrier frequency has a significant influence on antenna size and gain, as well as the propagation performance of a signal through a given wireless channel.

It's a Digital World

- **The Digital Age**
  - *Major communications standards*
    - » GSM, 802.11, Bluetooth, IS-95, WCDMA, the list goes on.
    - » Notable exceptions are AM/FM broadcast radio
- **Why Digital?**
  - *Spectral efficiency*
    - » More users/greater information capacity per Hz
  - *Commercial/Consumer support*
    - » Integrated voice/data/multimedia
  - *Security/privacy*
    - » Resistance to eavesdropping
  - *Advances in VLSI technology*
    - » Portable, low cost and low power





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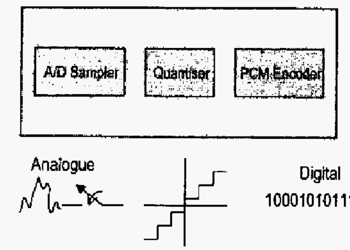
The growth of wireless communications has shown an exponential characteristic when plotted against time. Within the last twenty years dramatic improvements in circuit fabrication and very large scale integration (VLSI) techniques have made radio transceivers smaller, less expensive, easier to manufacture, and more reliable.

Digital methods of modulation, coding and network switching have allowed the deployment of large scale systems offering a range of voice, data, and multimedia services. Digital systems can operate with lower transmitter power and are more resilient channel effects such as fading. They are also able to support a higher capacity of information transfer. There have been many drivers for this technological change. Amongst these are the strong consumer and commercial demand for digital communications systems using wireless technology.

Digital Representation (Ones and Zeros)

- **Analogue signals to digital bits**
  - *Sampling, quantisation, source encoding*
  - *Uniform quantisation*
    - » Optimum SNR if all amplitudes are equally likely
  - *Nonuniform quantisation*
    - » Voice applications
      - Exploits amplitude statistics of voice ie. Peak amplitudes less likely
      - Sampled voice signal has higher SNR
  - *Source coding*
    - » PCM - Common Technique
    - » Several low rate voice coders



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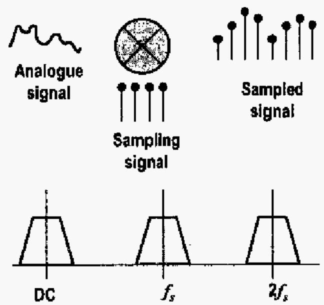
All information must be in a digital form for transmission within a digital communications system. The binary format is most common with each 'bit' having one of two possible states: commonly referred to as 'HI' or 'LO', '1' or '0' etc. More states can be accommodated using more than one bit, eg. L bits to represent 2^L states.

Information in analogue form is sampled and quantised using an analogue to digital converter (ADC) so that its amplitude at each sampling instant, the sample value, is mapped to one of 2^L predefined amplitude levels, each represented by L bits using an analogue to digital converter. This process is called Pulse Code Modulation (PCM), and is one of a number of possible source coding techniques. The objective of source coding is to provide an efficient digital representation of the analogue information. Examples include low rate speech encoders used in several mobile communications standards such as the Global System for Mobiles (GSM). Compare the 13kbit/sec GSM voice codec with the 64kbit/sec PCM codec used in the plain old telephone systems (POTS).

The quantiser characteristic is either uniform (quantisation levels equally spaced) or nonuniform (unequal spacing). Nonuniform quantisers are typically used for voice applications where more quantised levels are available for lower amplitudes, enhancing the signal-to noise (SNR) ratio of the sampled signal.

The Sampling Process

- Signals occupy bandwidth
- Time, frequency representation
  - *Multiplication in time*
  - *Convolution in frequency*
- Alias components
  - *repeat at multiples of f_s*



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To further understand the sampling process it is useful to consider both the time and frequency domain representations. The upper diagram shows the signal x(t) sampled every T_s seconds to give the sampled signal $x_s(t)=x(nT_s)$. The mathematical model consists of a multiplication operation between x(t) and a sampling signal s(t), represented by a sequence of impulse functions spaced every T_s apart:

$$x_s(t) = x(t)s(t) = x(nT_s) \qquad s(t) = \sum_n \delta(t - nT_s)$$
$$= \sum_n \alpha_n \delta(t - nT_s)$$

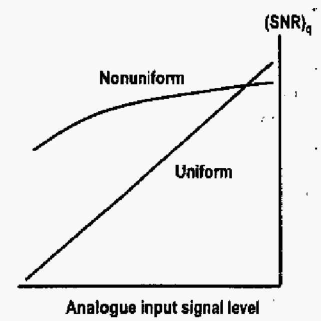
where α_n is the amplitude of the n-th sample. In the frequency domain these signals are represented by their power spectral density functions $W_x(f)$, $W_s(f)$, and $W^{xs}(f)$ and the multiplication operation is replaced by a convolution operation:

$$W_s = f_s \sum_m \delta(f - mf_s) \qquad f_s = \frac{1}{T_s} \qquad W_{xs}(f) = W_x(f) * W_s(f)$$
$$= f_s \sum_m \int_{-\infty}^{\infty} W_x(\lambda)\delta(f - \lambda - mf_s)d\lambda$$
$$= f_s \sum_m W_x(f - mf_s)$$

The sampling process has produced copies or aliases of the original signal spectrum occurring at multiples of the sampling rate f_s.

How many bits do I need?

- More bits
  - *More resolution*
  - *Decreases quantisation noise*
  - *Increases dynamic range*
  - *Higher $(SNR)_q$*
- Uniform $(SNR)_q$
  - *Linear (dB) with amplitude*
- Nonuniform $(SNR)_q$
  - *Increased for smaller amplitudes*
- Quantiser SNR



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For each sample, the amplitude of the information signal is quantised to one of 2^L amplitude levels represented by L bits. A natural question to ask is: How many bits are enough in each sample to convey the information in a digital form?

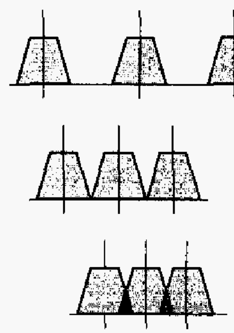
An analogue signal can be imagined as being quantised to an infinite number of levels. We could just use an arbitrarily high number of levels but this would lead to inefficiencies in the design of the digital system and the information transfer itself. In virtually all communications systems the receiver acts to recovers the original information signal from one that has a noise component. That is, the received signal has a finite SNR. The quantisation process also introduces a noise component onto a signal due to the uncertainty between the actual signal amplitude and the quantised amplitude of the sampled signal. The signal to quantisation noise ration $(SNR)_q$ depends on the type of quantisation employed and the amplitude statistics of the signal that is sampled. For the simple and often quoted example of a sinusoid with a peak to peak amplitude equal to the full scale amplitude of a uniform quantiser, $(SNR)_q = 6.02L + 1.76$ dB.

Typically, enough bits are chosen so that quantisation noise is a negligible component of the overall SNR. However if you choose too many bits you can end up making a very precise sample of the signal's noise component!

However a good reason for choosing more bits is to allow enough headroom to sample the peaks of the signal. The amplitude statistics of the signal are usually well understood and the required headroom can be designed.

How fast do I sample?

- The Nyquist Rate
- Lower bound on sampling rate
  - *Sample at greater than twice the occupied bandwidth (Hz) to avoid aliassing*
  - *In practice*
    - » anti-alias filters limit the signal bandwidth
    - » The oversampling factor >2 eases filter requirements
  - *Theorem extends to bandpass signals as well (modulated IF/RF carriers)*



Sampling at greater than the Nyquist rate.

Sampling at the Nyquist rate. (Better check those signal filters!)

Aliasing due to sampling at less than the Nyquist rate

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Another obvious question concerning the sampling of a signal is: How often do I need to sample the signal to ensure an accurate digital representation? The famous sampling theorem, states: a signal of bandwidth B may be completely determined from its samples taken 2B apart. The sample rate f_s=2B is a lower bound, and is often referred to as the Nyquist rate.

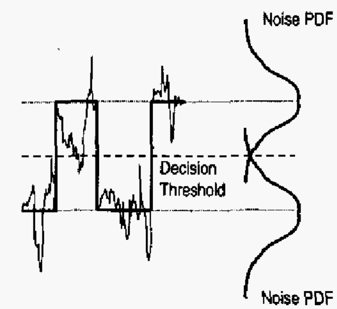
In practice it is often necessary to sample at higher rates due to the difficulties in ensuring the sampler's input signal is truly bandlimited. The frequency domain diagrams show the effect of sampling a signal at a) f_s>>2B, b) f_s approx = 2B, c) f_s<2B. For the case of c) the signal spectrum is distorted by its nearest alias component.

The sampling theorem is also extended to the sampling of bandpass signals. The Nyquist frequency must be greater than twice the signal bandwidth (Not the carrier frequency!). Additionally the signal's occupied bandwidth must not straddle an integer multiple of half the sampling rate.

Simple Communications Example

- A baseband system
  - *TX quantises information and transmits a serial bitstream*
  - *Channel adds random noise with a Gaussian PDF*
  - *The RX decides the value of each "bit"*
  - *Random noise varies the received signal's amplitude*
    - » Can cause the RX to make an incorrect decision - a bit error
    - » Can derive the bit error rate (BER) as a function of SNR
    - » BER is a key performance indicator for a digital communications system

Signal plus noise



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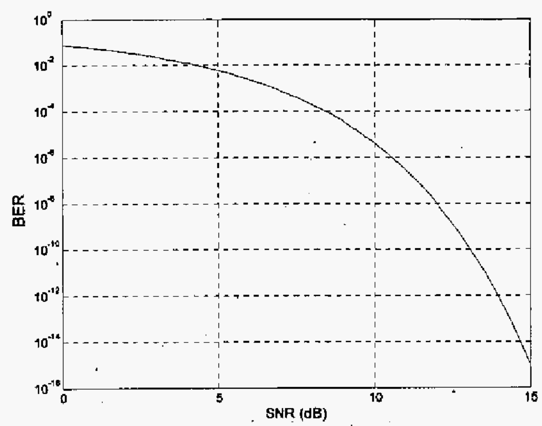
This simple example highlights the effect of additive noise on the performance of a digital communications system. The receiver decodes the binary level signal using a threshold comparison. The diagram shows the two signal states, and the decision threshold together with the superimposed noise amplitude probability density function (PDF), which are Gaussian in this case.

The tails of the PDF extend across the threshold indicating that there is a finite probability that the noise will be significant enough to cause the receiver to make a wrong decision and to cause a bit error.

A common metric of performance for digital communications is the bit error rate (BER). It gives the frequency of bits in error with respect to the number of bits transmitted and correctly decoded. Assuming that each decoded bit is an independent operation then the BER is also equal to the probability that a single bit will be in error. For this example and for many other examples the BER function as a function of SNR can be derived.

The BER Function



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The BER function for the previous example is easily derived. Referring to the threshold diagram we can see that the probability of a single bit error, P_e, is equal to the area past the threshold under each PDF.

$$P_e = \Pr(a_k = 1 \cup x_k = 0) + \Pr(a_k = 0 \cup x_k = 1)$$
$$= \Pr(x_k = 0)\Pr(a_k = 1 | x_k = 0) + \Pr(x_k = 1)\Pr(a_k = 0 | x_k = 1)$$
$$= \frac{0.5}{\sigma\sqrt{2\pi}} \int_{-\infty}^{0} \exp\left[\frac{-(y-A)^2}{\sigma^2}\right] dy + \frac{0.5}{\sigma\sqrt{2\pi}} \int_{0}^{\infty} \exp\left[\frac{-(y+A)^2}{\sigma^2}\right] dy$$
$$= \frac{1}{\sigma\sqrt{2\pi}} \int_{0}^{\infty} \exp\left[\frac{-(y+A)^2}{\sigma^2}\right] dy$$

$$t = \frac{y+A}{\sigma} \Rightarrow dt = \frac{dy}{\sigma} \quad LL: y = 0 \Rightarrow t = \frac{A}{\sigma}$$

After a change of variables

$$P_e = \frac{1}{\sqrt{2\pi}} \int_{A/\sigma}^{\infty} \exp\left(\frac{t^2}{2}\right) dt = Q\left(\frac{A}{\sigma}\right)$$

where the Q(x) function is defined by

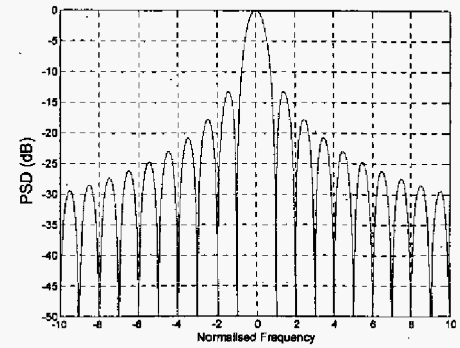
$$Q(x) = \frac{1}{\sqrt{2\pi}} \int_{x}^{\infty} \exp\left(\frac{\gamma^2}{2}\right) d\gamma$$

Note that the BER is a function of the signal peak to noise standard deviation ratio. The graph shows a decreasing BER as this ratio increases. This is what you would expect, as the noise is increased relative to the signal level, it is more likely that a bit error will occur.

Note that this example assumes the signal is transmitted through a distortion-free channel and is decoded by an ideal receiver. This is hardly ever the case in practice though these functions serve as a useful benchmark to the performance of a real system.

Ones and zeros are square waves aren't they?

- **This is a misconception**
  - *Oscilloscope observations*
    - » Digital HW signals - clocks, PRBS
- **Sampled DSP operations**
  - *Mathematically represented by delta functions*
- **What if we transmitted a Random squarewave bitstream?**
  - *Bit wide pulses - No ISI, BUT*
  - *Very wide occupied bandwidth*
    - » PSD is a sinc^2 function
    - » Multiple sidelobes - Nulls occur at non-zero multiples of the bit rate



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If you observe a binary logic level level signal such on a logic analyser or oscilloscope it appears as a squarewave signal. This can often lead to the misconception that, because we transmit digital information, we are actually transmitting squarewaves.

Recall that a sampled digital signal is represented mathematically by an amplitude weighted train of impulses and that it is the amplitudes that are encoded into 'bits'. In the previous BER example binary information was sent as non-return to zero (NRZ) squarewaves. However this is rarely, if ever, done in practice, mainly for reasons of bandwidth efficiency. The power spectrum W(f) of a random NRZ binary sequence shown in the diagram is given by:

$$W(f) = 2(T_b A)^2 \operatorname{sinc}^2(T_b f) \qquad \operatorname{sinc} x = \frac{\sin \pi x}{\pi x}$$

The spectrum consists of a main lobe with a number of minor sidelobes extending the occupied bandwidth towards infinity. This is not a good spectral characteristic if a number of different users are to be supported.

Information is transmitted in practice using shaped pulses. Each bit of a NRZ binary signal can be considered as a pulse, but its shape is a squarewave. Bandwidth concerns aside, a squarewave pulse has the desirable property of being truncated in the time domain to the period of the information rate. This means that each received pulse does not contain any residual components from previously received pulses. This effect, often referred to as intersymbol interference (ISI), can have a major impact on BER performance. Unfortunately any attempt to shape the pulse to lower its occupied bandwidth will increase its length in the time domain.

Pulse Shaping - Bandwidth and ISI

- In practice we transmit filtered pulses - Matched filtering
  - *Lower occupied bandwidth BUT*
    - » Pulse lasts over several "bit" periods
    - » Timing errors lead to ISI
  - *Pulse shaping trades off bandwidth and sensitivity to ISI*
  - *Several classes of pulse shapes conform to preferred set of characteristics*
  - *Raised Cosine (RC) is the most common*
    - » Usually split into two root raised cosine filters in the transmitter and receiver
    - » The channel also adds a contribution to the received pulse shape
    - » The receiver often uses an equaliser to restore the pulse prior to demodulation

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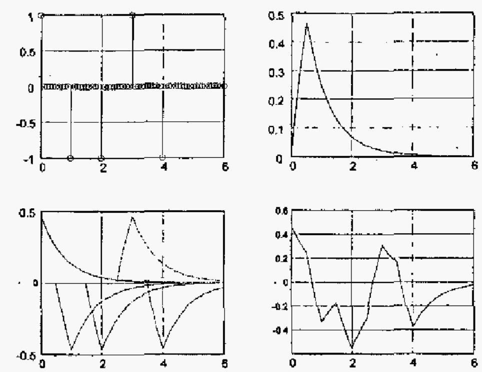
The strategy for determining a suitable pulse shape usually involves trading off the conflicting requirements of a low occupied bandwidth and a low level of ISI. Each pulse is formed by convolving the digital samples with a certain type of filter. Because the samples are impulse functions the resulting pulse shape is the same as the impulse response of the filter.

The Raised Cosine is the most commonly used class of filter.

Simple pulse filtering example

1st order lowpass of DAC output and ISI contribution



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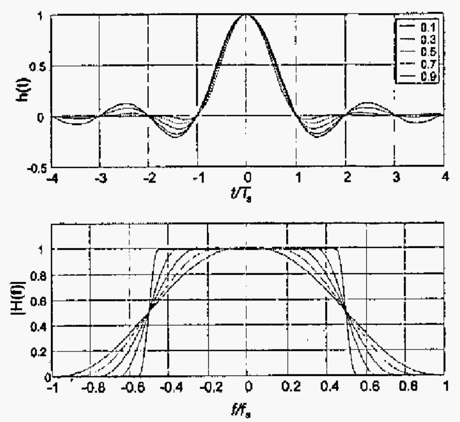
This example highlights the importance of the choice of pulse shaping filter. The top left diagram shows the digital impulses. They are passed through a digital to analogue converter (DAC) which applies a zero order hold (ZOH). At the DAC output is a 1st order lowpass filter (LPF) with a 3dB frequency much less than the sampling rate. The top right diagram is the pulse shape resulting from the convolution in time of the DAC ZOH with the LPF.

The bottom left diagram shows each individual pulse shape, and the overall signal is shown in the bottom right diagram. This shows a significant degradation due to ISI and a moderate amount of noise could force the receiver into making a bit error.

Raised Cosine Example



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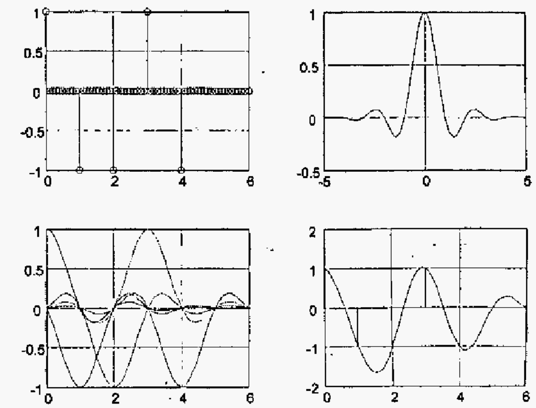
The magnitude response, |H(f)|, and impulse response, h(t), of a raised cosine filter are given by

$$|H(f)| = \begin{cases} 1 & |f| \le \dfrac{1-\alpha}{2T_b} \\ \cos^2\left[\dfrac{\pi T_b}{2\alpha}\left(f - \dfrac{1-\alpha}{2T_b}\right)\right] & \dfrac{1-\alpha}{2T_b} < |f| \le \dfrac{1+\alpha}{2T_b} \\ 0 & |f| > \dfrac{1+\alpha}{2T_b} \end{cases} \qquad h(t) = \frac{\cos(\alpha\pi t)}{T_b\left\{1-\left(\dfrac{2\alpha t}{T_b}\right)^2\right\}}\operatorname{sinc}\left(\frac{t}{T_b}\right)$$

The diagrams show the time domain pulse shape and frequency response of raised cosine filters for various values of α. This parameter is often called the bandwidth expansion factor because its dominant effect on the occupied bandwidth. A desirable characteristic of the pulse shape is that it has a maximum value at the origin (the receiver decision point) and a zero value at multiples of the symbol period (no ISI). However small errors in the timing will result in a small ISI contribution. The effect of α is to reduce the height of the minor lobes and hence the sensitivity to ISI as it increases. This comes at the expense of an increase in the occupied bandwidth, which is given by $0.5(1+\alpha)/Tb$ for a baseband signal.

Raised Cosine Pulses



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The benefits of a raised cosine pulse shape are evident in this example. The lower left trace shows that there is negligible ISI contribution because the pulse shape has a zero value at each sampling point other than its own.

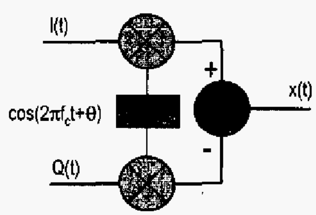
Raised cosine filters are almost exclusively implemented digitally these days, although analogue approximations were used in the past, usually with linear phase designs using Bessel and Butterworth filters.

Note that the DAC ZOH effect (|sinc| frequency domain effect) is still there but the digital implementation allows the filter to be incorporated with x/sinx equalisation.

Characterisation of a Modulated Carrier

- Equivalent forms of a modulated carrier
- Complex envelope notation

Complex BB to Real RF



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Most of the previous slides have concentrated on the baseband signal. In wireless communications this is modulated onto a carrier signal. A modulated carrier signal x(t) may be expressed as:

$$
\begin{aligned}
x(t) &= r(t)\cos(2\pi f_c t + \theta + \phi(t)) \\
&= I(t)\cos(2\pi f_c t + \theta) - Q(t)\sin(2\pi f_c t + \theta) \\
&= \mathrm{Re}\left\{c(t)e^{j2\pi f_c t + \theta}\right\} \\
c(t) &= I(t) + jQ(t) \\
&= r(t)e^{j\phi(t)}
\end{aligned}
$$

$$
r(t) = \sqrt{I(t)^2 + Q(t)^2} \qquad \phi(t) = \arg\left[\frac{Q(t)}{I(t)}\right]
$$

where f_c and q are the frequency and phase* of the CW carrier signal itself and r(t) and f(t) are amplitude and phase components contributed by the modulating signal c(t). The complex form of c(t) allows representation for modulation schemes which apply a joint amplitude and phase modulation of the carrier. The transmitter's modulator converts the baseband information signal into the modulating signal.

The diagram shows an transmitter architecture for the transfer of the baseband modulated signal onto a carrier.

* Note that a frequency modulation component may also represented by this expression using the integral relationship between frequency and phase.

Bits and Symbols

- Binary alphabet is 1 bit per symbol
- Larger symbol alphabets can be used
- Improved spectral efficiency (Bits / sec / Hz)
- Pulse shaping is done on the symbol
- ISI considerations are the same

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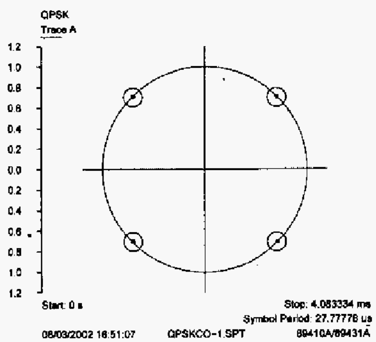
We have seen so far that Digital communication involves the transfer of information that has been converted to bits. The simple communications example used a binary level signalling scheme and with pulse shaping we have found that we can limit the bandwidth to be of the order of the bit rate.

An increase in spectral efficiency is possible by transmitting symbols from an alphabet that contains more than two states, as for the binary signal example. What is usually done is symbols are mapped from N bits to form an alphabet, or constellation of 2N symbols.

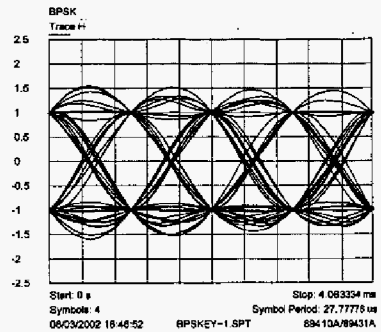
The advantage here is that the preceding discussion on pulse shaping and occupied bandwidth is still the same, except that we now talk in terms of symbols instead of bits.

Constellation and Eye Diagrams

QPSK Constellation



BPSK/QPSK eye diagram



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The constellation and eye diagrams are a representation of the modulation content of a signal. They can be used to identify the type of modulation and to assess the signal quality .

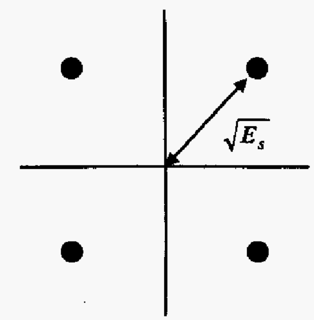
The constellation diagram is a rectangular plot of the real and imaginary components of a complex modulated signal. The most common form of the plot shows the actual decision point taken by the receiver of the symbol in Euclidean space. It is common to also show the transitions between symbols.

An eye diagram shows the real or imaginary part of a complex modulation. It consists of multiple overlays of symbol pulses over a period of 1 or 2 symbols. The receiver will usually try and sample the received symbol at the maximum eye opening.

Signal and noise power, energy per bit etc.

$$SNR = \frac{C}{N} = \frac{A^2}{2} \cdot \frac{1}{N_0 B} = \frac{A^2 T_s}{2N_0} = \frac{E_s}{N_0}$$

$$\frac{C}{N} = \frac{r_s E_s}{N_0 B} = \frac{r_b E_b}{N_0 B} = \frac{E_s}{N_0} = \frac{\eta E_b}{N_0}$$



Noise power is measured in the signal bandwidth

For linear modulation r_s=B

Raised Cosine rolloff factor does not influence noise bandwidth

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Signal to noise ratio is often represented in terms of carrier power, C, noise power, N, measured in the signal's bandwidth, B. This is not necessarily the occupied bandwidth, but more usually equal to the symbol rate, particularly for linear modulation schemes. Other common terms are the energy per bit and energy per symbol, E_b and E_s. Noise power (assuming Gaussian distributed white noise) is also represented by its spectral density, N_0.

For a signal of peak amplitude A, its power and energy per symbol are given by:

$$C = \frac{A^2}{2} \qquad E_s = \frac{AT_s}{2}$$

where Ts = 1/ B= r_s is the symbol period.

$$SNR = \frac{C}{N} = \frac{A^2}{2} \cdot \frac{1}{N_0 B} = \frac{A^2 T_s}{2N_0} = \frac{E_s}{N_0}$$

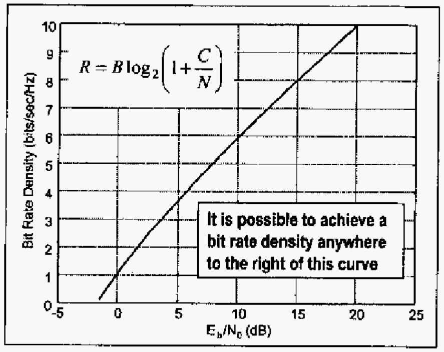
$$\frac{C}{N} = \frac{r_s E_s}{N_0 B} = \frac{r_b E_b}{N_0 B} = \frac{E_s}{N_0} = \frac{\eta E_b}{N_0}$$

C = Carrier Power
T_s = Symbol period
B = Bandwidth
A = Peak Signal Amplitude

N = Noise Power (in signal bandwidth or channel bandwidth)
r_s = Symbol rate
N_0 = Noise power spectral Density W/Hz (AWGN)
E_b/N_0 is the main way of comparing modulation schemes

How much data can I send?

- Shannon's capacity formula
  - *An upper bound for the information rate*
  - *Assumes a Gaussian noise channel*
  - *trade offs between power and bandwidth*
  - *Shannon gave us the bound on capacity but not how to achieve it!!*
  - *Example: It took us 50 years to get close to the bound for a normal telephone line modem.*



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Shannon's formula shows the fundamental bound on the information rate R (bits/sec).

$$R = B\log_2\left(1+\frac{C}{N}\right)$$

The maximum rate is a function of bandwidth and SNR. Many systems lie below this curve because they are bandwidth limited (ie. The receiver needs a high SNR) or power limited (ie. A high bandwidth is needed).

The bound is often represented in terms of the bit rate density r = R/B (bits/sec/Hz). The formula assumes the signal can be encoded such that a low BER is possible at the desired SNR. The signal encoding is often a combination of source coding, which aims to provide an efficient digital representation of the information, and channel coding, which adds redundancy to the information but gives and error detection and/or error correction capability. There are a variety of encoding schemes used in every communications system standard.

Using the bit rate density, we can derive its dependence on the required energy per bit, and this is plotted above.

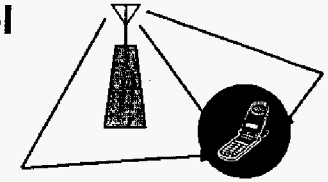
$$\frac{R}{B} \le \log_2\left(1+\frac{C}{N}\right)$$

$$\le \log_2\left(1+\frac{R}{B}\cdot\frac{E_b}{N_0}\right)$$

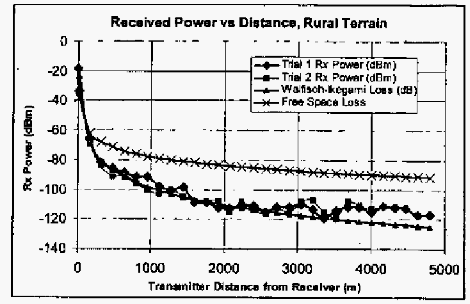
$$2^r \le 1 + r\frac{E_b}{N_0}$$

$$\frac{E_b}{N_0} \ge \frac{2^r - 1}{r}$$

Wireless Channel Characteristics



- Propagation in free space
  - *Geo-Sat and LOS links ∝ d^2*
- Basic propagation mechanisms
  - *Reflection, Diffraction, Scattering*
- Higher order path loss
  - *propagation surveys fit to models*
- Multipath fading
  - *Frequency selective and flat fading*
- Doppler shift in mobile applications



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The wireless channel, like any transmission medium acts to attenuate the signal between the transmitter and the receiver. The attenuation, or path loss, is known to vary with the square of the transmission distance for a fixed line of sight (LOS) radio link with high masts. However there are many cases where the effects of the wireless channel can be more severe, and the path loss distance exponent may be higher, or may vary according to distance and frequency.

There are three basic mechanisms for propagation:

Reflection - Propagating wave reflects off a surface with dimensions >> λ.

Diffraction - Propagating wave obstructed by a surface with sharp edges, producing secondary waves that bend around the surface

Scattering - Propagating wave encounters many small particles, such as foliage and street lamps, and the resulting reflections are diffused through space.

Analytical and empirical channel models exist to quantify the propagation effects of a particular radio link. Most of these are based on measured results from experimental trials, and the use of statistical techniques.

Multipath fading can be a problem for transmission in cluttered environments, where the radio signal can be reflected off surfaces such as the walls of buildings. Because of this the receiver can receive multiple instances of the transmitted signal via different paths, with each path undergoing a different overall phase shift and path loss . This phenomenon can cause massive variations in the received signal power. This effect can also be time and frequency variant, particularly if the radio transceivers are mobile.

Linear Modulation Schemes

- Amplitude varies linearly with modulating signal
- Bandwidth efficient
- Require linear amplifiers with poor power efficiency
- Widely used schemes
  - *M-ary PSK*
  - *M-ary QAM*

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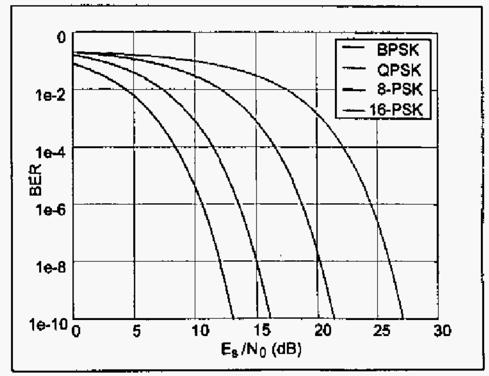
All linear modulation schemes contain some degree of AM content. They are generally very bandwidth efficient, the noise bandwidth is equal to the symbol rate when raised cosine filtering is used. However the power efficiency can be affected by the requirement for linear amplifiers to preserve the amplitude mapping of the modulation envelope.

The most widely used schemes are phase shift keying (PSK) and quadrature amplitude modulation (QAM).

M-ary PSK

- Polar Symbol constellation
  - *equally spaced constant radius symbol locations*
  - *BPSK, QPSK, 8-PSK most common*
  - *RRC, and RC pulse shaping common*
- Variants of QPSK
  - *DQPSK*
  - *pi/4-DQPSK*
  - *OQPSK*



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M-PSK complex symbol locations, i = 0,...,M-1

$$c_i = \begin{cases} \sqrt{E_s}\exp\{j[(i-1)\pi]\} & M = 2 \\ \sqrt{E_s}\exp\left\{j\left[\frac{2\pi i}{M} + \frac{\pi}{M}\right]\right\} & M > 2 \end{cases}$$

M-PSK BER expressions

$$P_e = 2Q\left[\sqrt{\frac{2E_s}{N_0}}\sin\left(\frac{\pi}{M}\right)\right]$$

There are also commonly used variants of QPSK and 8-PSK. A major advantage of these schemes is that they avoid symbol-to-symbol transitions through the origin of the constellation diagram, thus reducing the peak-to-average power of the the modulated carrier signal (lower AM content).

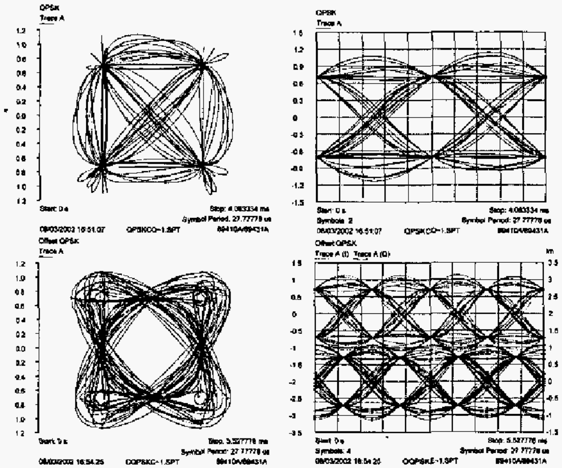
OQPSK: Imaginary component of the symbol is delayed in time by half a symbol period.

$\pi/4$ -QPSK: Two QPSK constellations are used, offset by $\pi/4$ radians. Constellation mapping alternates with successive symbols

Recent 8-PSK variant, $3\pi/8$-8PSK (GSM/EDGE): Two 8PSK constellations are used, offset by $3\pi/8$ radians. The symbol mapping alternates between the two constellations for successive symbols.

The BER results show the trade off made between power and efficiency. The higher order schemes require a greater SNR to achieve a given BER.

QPSK and OQPSK



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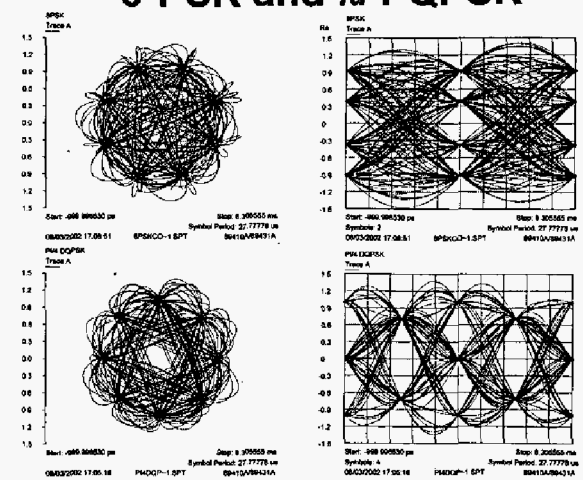
Here the differences in the two constellations are apparent. The QPSK signal can be thought of as two orthogonal BPSK signals and the I/Q eye diagrams are identical. The lower peak to average power in OQPSK is due to the lack of symbol-symbol transitions through the origin. The I/Q eye diagrams for OQPSK show time delay of the Q channel pulses.

8-PSK and $\pi/4$-QPSK



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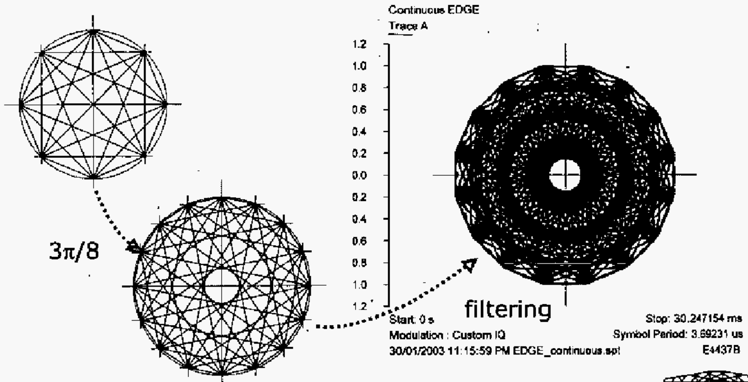
Two 8-point constellations are shown but only 8-PSK contains 8 independent symbols. The $\pi/4$-QPSK uses two 4-point constellation mappings, offset by 45°, which are alternated for each alternate symbol. This avoids symbol-symbol transitions through the origin and hence lowers the peak to average power of the modulated carrier signal.

EDGE modulation

- Enhanced Data rates for Gsm Evolution
- **EDGE uses 8-PSK with 3π/8 shift between each symbol, filtered to keep to the same spectrum as plain GSM**



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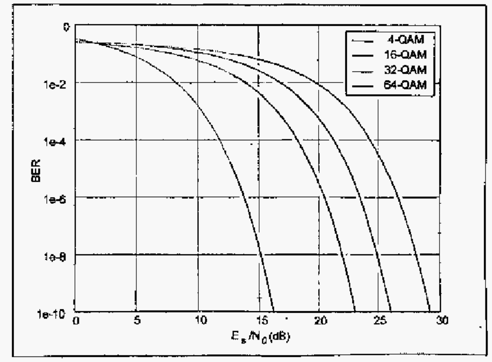
The 3π/8-8PSK modulation scheme used in EDGE follows a similar idea to π/4-QPSK modulation; a 16-point constellation is formed from two 8-point mappings, offset by 3π/8 radians. Again this results in no symbol-symbol transitions through the origin and lowers the peak to average power compared to normal 8-PSK.

M-ary QAM

- **Rectangular Symbol constellation**
  - ***4-QAM identical to QPSK***
  - ***16/32/64/128/256-QAM most common***
  - ***RRC, and RC pulse shaping common***
  - ***Spectral efficiency identical to PSK for the same bits/symbol***
  - ***QAM more power efficient - 16-QAM and 8-PSK similar BER***



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The most common M-QAM complex symbol locations fall in a rectangular array. The coordinates of each symbol may be defined as:

$$c_i = a_i + jb_i = (a_i, b_i) \qquad i = 0, \ldots, M-1$$

As an example, the locations for a square 16-QAM constellation are:

{(-3,-3)(-1,-3)(1,-3)(3,-3)(-3,-1)(-1,-1)(1,-1)(3,-1)(-3,1)(-1,1)(1,1)(3,1)(-3,3)(-1,3)(1,3)(3,3)}

Approximate BER expressions

4-QAM

$$P_e \approx 2Q\left(\sqrt{\frac{2E_s}{N_0}}\right)$$

16-QAM

$$P_e \approx \frac{3}{4}Q\left(\sqrt{\frac{E_s}{5N_0}}\right)$$

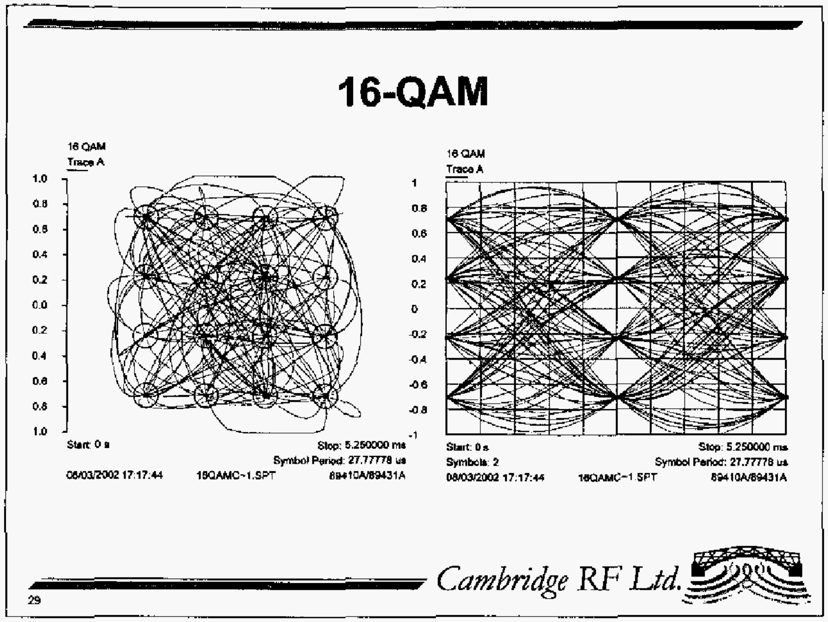
32-QAM

$$P_e \approx \frac{7}{10}Q\left(\sqrt{\frac{E_s}{10N_0}}\right)$$

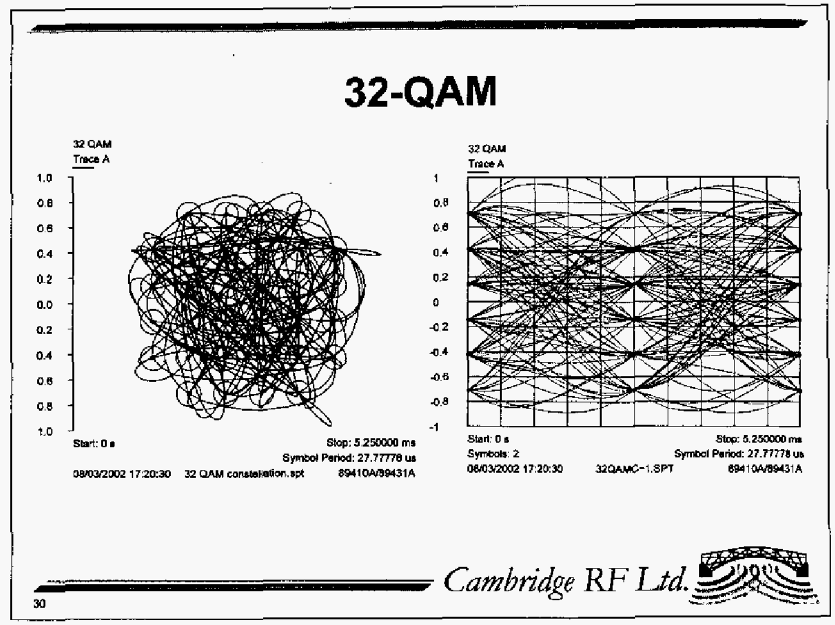
64-QAM

$$P_e \approx \frac{7}{12}Q\left(\sqrt{\frac{E_s}{21N_0}}\right)$$

The higher order schemes require high SNRs. Notice the performance is 16-QAM is very close to that of 8-PSK, but 16-QAM is more spectrally efficient.

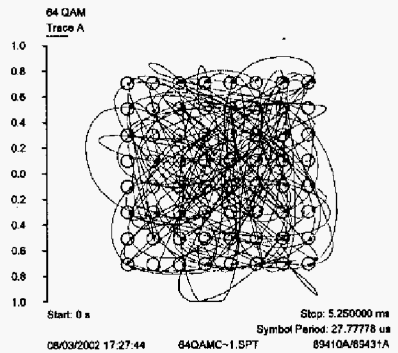


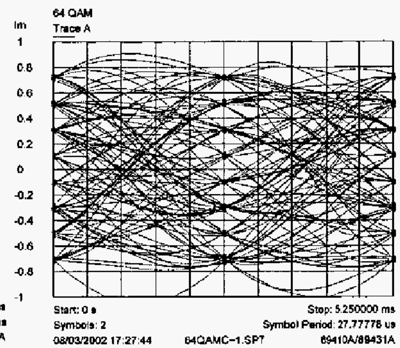
16-QAM 4x4 constellation and eye diagram. Both the I and Q channels consist of filtered 4-level amplitude pulses.



32-QAM (5bits/symbol) constellation (6x6 with one symbol removed from each corner) and eye diagram. Both the I and Q channels consist of filtered 6-level amplitude pulses.

64-QAM





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64-QAM (6bits/symbol) 8x8 constellation and eye diagrams. Both the I and Q channels consist of filtered 8-level amplitude pulses.

Non-Linear Modulation Schemes

- Constant carrier amplitude
- Generally occupy a larger bandwidth
  - *However certain pulse shapes can give low out of band spectral power*
- Can use nonlinear power-efficient amplifiers
- Widely used schemes
  - *2-level, 4-level FSK*
  - *MSK*
  - *GMSK*

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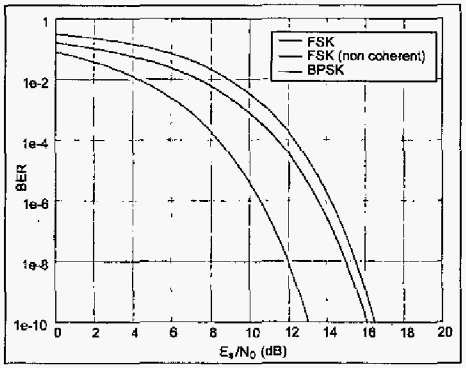
Nonlinear modulation schemes maintain a constant modulation envelope regardless of the modulating signal. Because of this nonlinear amplifiers may be used to increase the overall power efficiency of the system.

They can be less bandwidth efficient than linear schemes, but are often used in conjunction with modulation filters to lower the out of band spectral power. Two most common schemes fall under the umbrella of continuous phase modulation schemes: frequency shift keying (FSK), a 'digital' form of frequency modulation, and minimum shift keying (MSK), which can be thought of as either a phase modulation or as a special case of FSK.

FSK

- Each FSK symbols is mapped to a frequency above/below the carrier
- Simple transceiver architectures
- 2-level and 4-level FSK most widely used
  - *2-FSK unfiltered pulse used in POCSAG pagers/radio telemetry*
  - *4-FSK RC filtered pulse in ERMES/FLEX higher speed paging*
  - *GFSK Gaussian filtered pulse used in Bluetooth*



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Frequency shift keying is a "digital" frequency modulation scheme. The simplest form of FSK is for 1bit/symbol, or 2-level FSK. The carrier is set to one of two frequencies above or below the the nominal carrier frequency depending on the symbol state.

Because the modulated carrier frequency is above or below its nominal centre frequency, the I/Q baseband signals either lead or lag each other in phase. Many pocket pagers for the POCSAG paging standard use a simple form of unfiltered 2-level FSK whereby a direct conversion receiver and D-Flip Flop can be used as a simple demodulator, instead of an FM discriminator

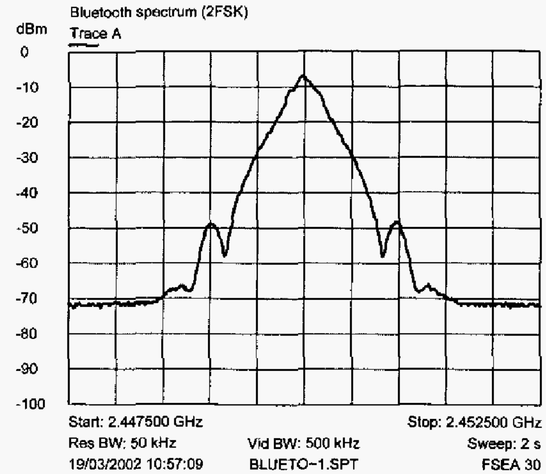
The unfiltered FSK power spectrum is the sum of each spectrum associated with a particular symbol, centred at its offset frequency.

Greater spectral efficiency is observed in 4-level FSK, using 2bits/symbol and pulse shaping. This modulation scheme is used in mobile data standards such as RD-LAP and advanced paging standards such as ERMES and FLEX

The BER results show that FSK does not perform as well as BPSK. However it has the advantage of a simpler transceiver architecture. BPSK requires a coherent demodulation, that is, the demodulator has to align to the carrier phase of the received signal. It is possible to demodulate FSK noncoherently with only a small degradation in the BER performance as shown in the graph.

Filtered FSK

- Bluetooth GFSK
  - *"Triangular" Power Spectral Density (PSD)*
  - *140kHz deviation, 1Msym/s, 1MHz channels*
- 4-FSK is similar
  - *RC filtering*



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Whilst all the M-PSK and M-QAM power spectra have the familiar "sinc" shaped main lobe, power spectrum of nonlinear modulations schemes employing the use of a modulation filter can decay more rapidly at offsets from the carrier. The example shows a Gaussian filtered FSK (GFSK) signal as used in Bluetooth. Note the "triangular" shape of the power spectrum, which is also found in other filtered nonlinear schemes such as raised cosine filtered 4-level FSK.

MSK (Minimum Shift Keying)

- MSK = FSK at the minimum possible frequency deviation
  - *Frequency deviation = half the symbol rate*
  - *Can also be thought of as phase modulation - ±90° each symbol*
  - *Equivalent to OQPSK with a "half cosine" pulse shape*
- Most commonly used with a Gaussian filter
  - *Lowers the out of band spectral power*
  - *Used in GSM*

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Minimum shift keying (MSK) is a 1bit/symbol modulation scheme where the frequency shift between symbols is equal to half the symbol rate. This also means that the total carrier phase shift over a symbol period is ±90° depending on the symbol state.

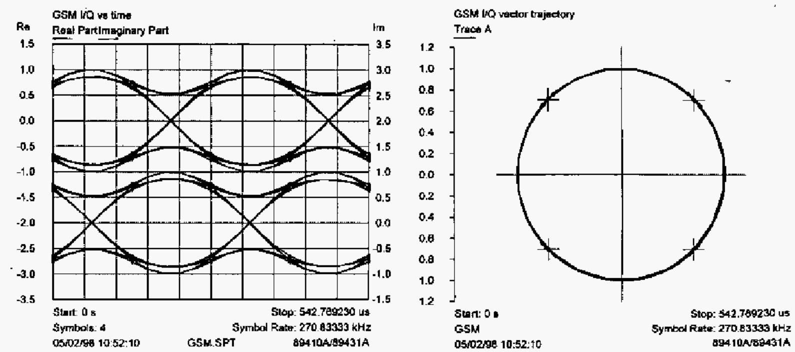
Because the modulation scheme is still based in 2-level FSK the I/Q baseband signals lead or lag each other in phase, for MSK the phase difference is 90° and resembles OQPSK with a sinewave pulse shape.

A commonly used variant is Gaussian Minimum Shift keying (GMSK) where the symbols are shaped by a Gaussian filter prior to modulation onto the carrier. This has the effect of lowering the out of band spectral energy of the modulated signal. The filter bandwidth is specified by its BT number, which is the product of the filter's 3dB frequency and the symbol period. A lower number increases bandwidth efficiency, but at the expense of ISI. GMSK modulation is used in cellular standards such as GSM and wireless data standards such as Mobitex. The BER for GMSK is given by:

$$P_e = Q\left(\sqrt{\frac{2\alpha E_b}{N_0}}\right)$$

where $\alpha \approx \begin{cases} 0.68 & BT = 0.25 \\ 0.85 & BT = \infty \quad (MSK) \end{cases}$

GMSK modulation



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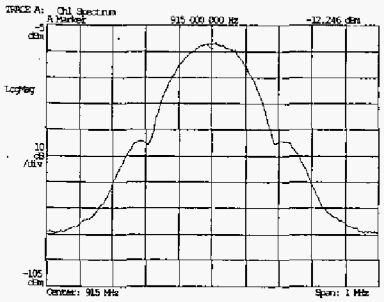
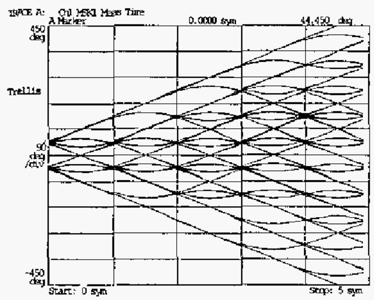
The I/Q channel eye diagrams are similar to OQPSK. Note the smooth phase changes in the I/Q plane. The circular shape indicates a constant carrier envelope.

GSM Modulation





GSM modulation accuracy is measured by Global Phase Error = rms deviation of actual phase trajectory from ideal phase trajectory

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In polar coordinates, it is useful to display the phase as an eye diagram of unwrapped phase vs time (known as a Trellis diagram). As we shall see, transmitter architectures began with IQ modulation, and later moved to directly modulating a VCO with the phase of the modulation.

OFDM

- **Military origins with HF parallel tone radios**
- **Resistance to multipath**
- **Spectral efficiency**
  - ***Using high order QAM carriers***
- **Requires linear amplifiers**
- **Requires tight frequency control**

- **Widely used in DVB and WLAN**

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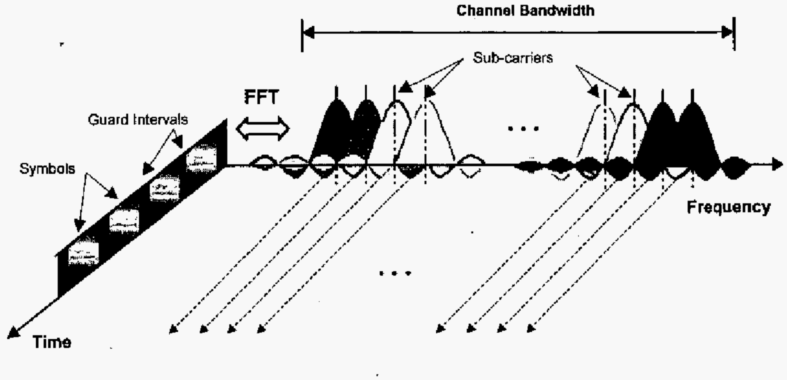
Also referred to as multi-carrier or discrete multi-tone modulation, orthogonal frequency division multiplexing (OFDM) is an extension of M-ary FSK, where a number of parallel modulated carriers, orthogonal to each other used to create a transmission signal that is power efficient rather than bandwidth efficient. The baseband signal is first modulated, in many cases with a high order QAM scheme. A block of N complex symbols are then taken. Each symbol in the block is multiplied by a subcarrier signal, and the frequency spacing of each carrier is equal to 1/(NTs), which ensures the orthogonality of the subcarriers and a smooth phase transition between symbols. The benefits of OFDM are high spectral efficiency, resiliency to RF interference, and lower multi-path distortion. An OFDM system designed to work against frequency selective fading has the data bits coded across the OFDM subbands in order to achieve an acceptable BER performance. Disadvantages of OFDM include the need for linear amplifiers due to a high peak to average power ratio, and a sensitivity to frequency offsets.

Although developed many years ago for military applications, OFDM is widely used used in digital broadcast such as digital audio and digital TV, and in WLAN standards such as 802.11a. OFDM is also being considered for future broadband applications such as wireless ATM and 4G standards.

OFDM Concept



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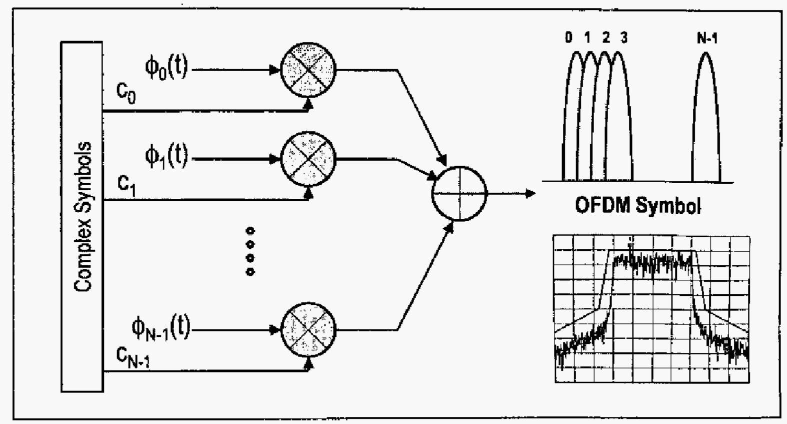
The technique of Orthogonal Frequency Division Multiplexing (OFDM) is based on the older technique of Frequency Division Multiplexing (FDM). In FDM, different streams of information are mapped onto separate parallel frequency channels. Each FDM channel is separated from the others by a frequency guard band to reduce interference between adjacent channels.

The OFDM technique differs from traditional FDM in the following interrelated ways:

1. very many multiple carriers (called sub-carriers) carry the information stream,
2. the sub-carriers are created in such a way as to be orthogonal to each other, and
3. the symbols change at a slow rate but have a high number of bits per symbol, such that multipath and channel fading effects have a small effect. A guard time may be added to each symbol to combat the channel delay spread.

OFDM Transmitter Signal Generation



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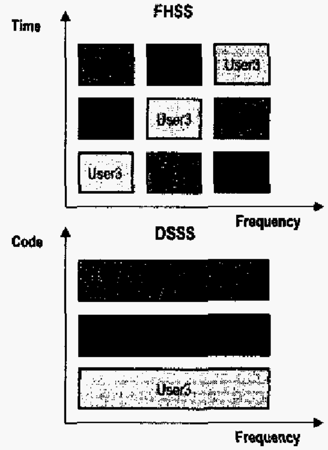
The diagram shows the conceptual generation of an OFDM signal. Each $\phi(t)$ is an orthogonal subcarrier, and if taken in its complex exponential form, OFDM generation starts to resemble the structure of the discrete Fourier transform. The use of the fast Fourier transform FFT and its inverse, the IFFT allow for computationally efficient signal processing of OFDM signals.

Spread Spectrum

- Associated with CDMA
- Military origins
- *Direct Sequence* spread spectrum
- *Frequency Hopping* spread spectrum
- Multipath resistance
- Process gain



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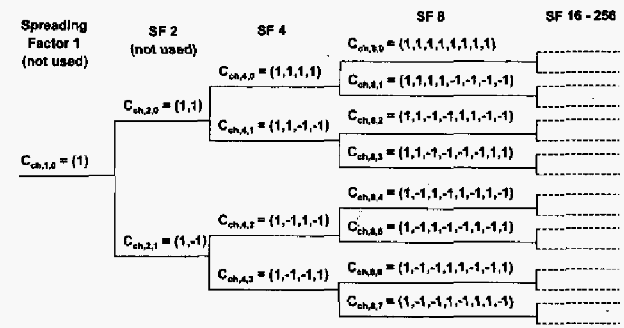
Spread spectrum is a technique in which the transmitted signal bandwidth is orders of magnitude greater than the symbol rate. A spread spectrum signal is controlled by a spreading signal, usually a pseudorandom (PN) sequence which can be reproduced at the receiver to despread and restore the original information signal. There are two main types of spread spectrum: Direct sequence (DS) and Frequency Hopping (FH)

In FHSS a sequence of modulated carrier bursts are transmitted on a pseudorandom set of carrier frequency channels called a hopset. Each channel contains enough bandwidth for the modulated burst, and is called the instantaneous bandwidth.

Resistance to multipath fading and interference is a major advantage of spread spectrum. If the signal's bandwidth is much larger than the frequency selectivity of the fade, only a portion of the signal's energy is lost. A measure of a spread spectrum signal's resistance to fading and interference is its processing gain. The higher the gain the better. For a DSSS signal this is the ratio of the spread bandwidth to the signal bandwidth. For a FHSS signal this is the ratio of the total hopping bandwidth to the instantaneous signal bandwidth.

CDMA - How it Works (i)

- Each data bit to be transmitted actually causes a whole sequence of bits to be transmitted instead
  - *OVSF (Orthogonal Variable Spreading Factor) or Walsh Codes*



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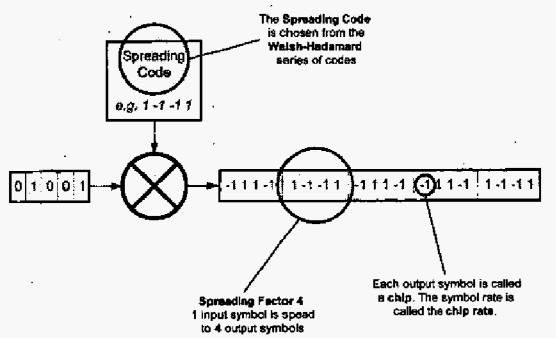
In DSSS the information signal is multiplied by the PN spreading sequence (an exclusive-OR operation in the digital domain). Each pulse of the PN sequence is called a chip and the final spread signal is expressed in terms of its chip rate. In most cases this signal is then BPSK-, or QPSK-modulated onto a carrier for wireless transmission.

Chip Rate and Spreading Factor



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CDMA - How it Works (ii)

- Walsh Codes – Have the property that several codes can be superimposed and then individually recovered by a correlating receiver
- Transmit a given code or its inverse to transmit a "0" or a "1"
- Transmit a longer code (higher Spreading Factor) when the unspread data rate is low, this gives more immunity against noise (coding gain).
  - *Voice uses high SF, internet downloads and videoconferencing uses smaller SF*
- The bit rate of the spread data is always the same (e.g. 3.84 MHz for 3GPP) – the "chipping rate"

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CDMA - How it Works (iii)

- In the uplink, most of the superposition of W-CDMA modulated data occurs in the air
  - *This requires a modulation format that can be superposed – BPSK or QPSK*
  - *All handsets in the cell are on the same frequency*
  - *Issues with getting the arrival power levels to be the same*
- In the downlink, it is assembled in the Base-Station's baseband processor
  - *Combination occurs with other Base-Stations on the same frequency – implies difficult signals to amplify and transmit*

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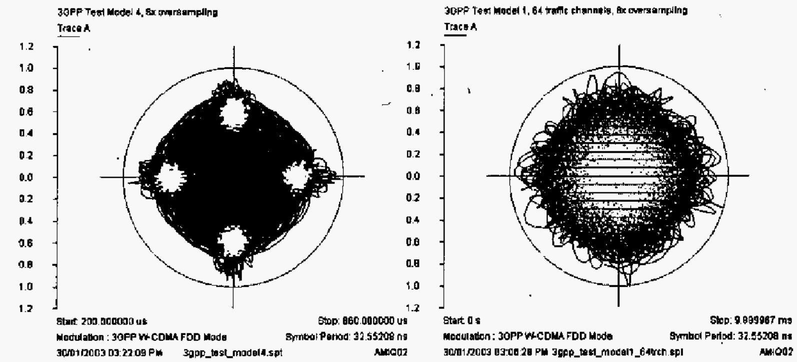
3GPP (Node-B) IQ constellation



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Multiple Access Techniques

- Simplex, Duplex modes
  - *TDD, FDD*
- All intended to support an increased number of users
  - *Originally 1 user occupies the frequency band for the duration of information transmission*
  - *FDMA, TDMA, CDMA, SDMA*
  - *All based around concept of orthogonality between users*
  - *Interference from other users*
    - » Adjacent channel interference (ACI), co-channel interference (CCI), multiple access interference (MAI)
- Combined techniques
  - *e.g. GSM with sectored cells FDMA/TDMA/SDMA*

Most communications systems allow a large number of users to communicate either simultaneously or via a shared provision of resources, using multiple access techniques

Communications over a single channel: Simplex = one way communications, Duplex = two way communications

TDD: time division duplex, two way communication on the same frequency separated by time intervals

FDD: frequency division duplex, two way communication at the same time but different frequency.

All multiple access methods use the concept that each support user can be made orthogonal to any other. This can be done by separating them in frequency, time, assigning them a unique code, exploiting spatial separation, or some combination of these.

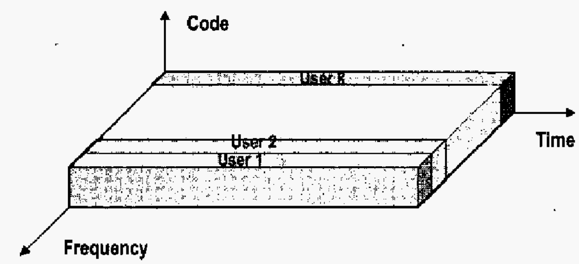
These ideas form the basis of:

FDMA: Frequency division multiple access TDMA: Time division multiple access

CDMA: Code Division multiple access SDMA: Space division multiple access

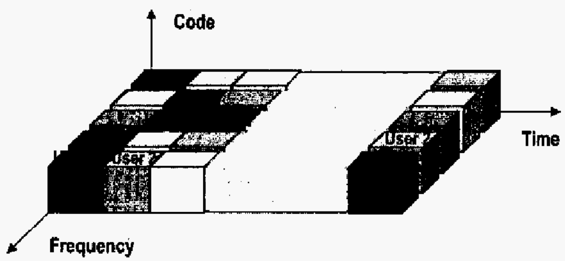
All multiple access techniques suffer a performance degradation brought on by the interference of other users. A user may interfere with other FDMA users if an appreciable part of their signal lies in another frequency channel. This is adjacent channel interference (ACI). TDMA users on the same frequency (such as neighbouring cellular users) may contribute a finite level of co-channel interference (CCI). For CDMA users, finite crosscorrelation of codes and the power of each user in addition to background noise levels contribute to multiple access interference (MAI).

FDMA



Each user is assigned a frequency channel. Users may communicate at the same time.

TDMA



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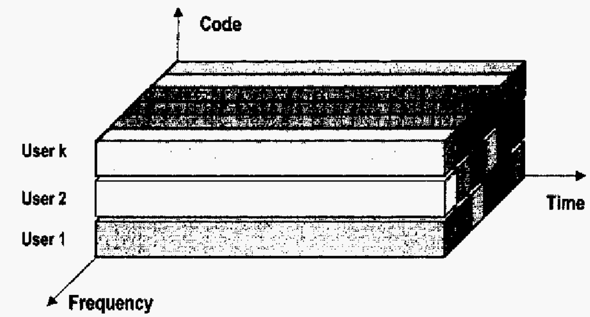
Users can communicate on the same frequency but are assigned different slots in time on a cyclical basis. Usually combined with some frequency hopping to mitigate against multipath fading.

CDMA



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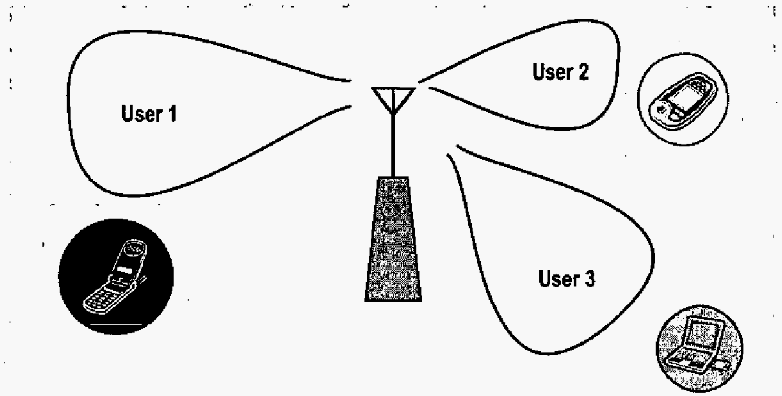
Users can communicate on the same frequency and at the same time but are assigned a unique code with low crosscorrelation properties.

SDMA



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The radiated energy of a radio basestation and/or user terminal is controlled so that a user may communicate in any multiple access mode but with a much reduced interference level from other users. Sectored antennas are widely used in cellular systems and the area of adaptive or "smart" antennas is quite an active area of research and development

Modulation measurements (Transmitter)

- Power
- Two-tone Intermodulation
- Adjacent Channel Power (ACP)
  - *GSM and W-CDMA measurement examples*
- EVM
  - *RMS, peak, 95th percentile*
  - *Measurement uncertainty properties of small EVMs*

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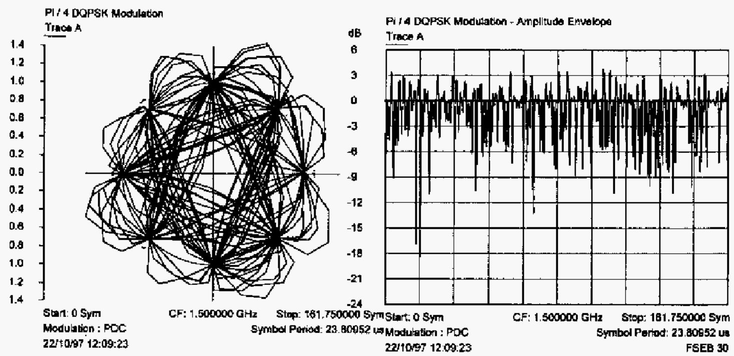
This section looks at ways of characterising the amount of distortion in a modulated RF signal, and also at the measurements made on components to determine the effect of their nonlinearity on modulation quality.

We will see that amplitude compression characteristics (i.e. AM-AM) and change of phase with level (i.e. AM-PM) are the source of nearly all of the modulation quality impairments we are likely to measure. Therefore, we will also deal with diagnostic techniques to quantify and minimise the source of the problems.

Modulation Characteristics



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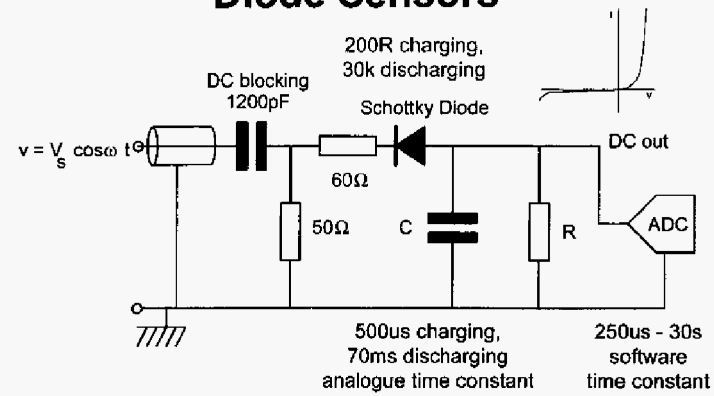
When measuring the power of a transmitter with modulation present, it is necessary to be aware of the amplitude variations associated with that modulation. Shown above is a typical example using Pi/4 DQPSK modulation (used in NADC and TETRA, etc). Although the amplitude is constant at each instant when the symbols are valid, the trajectory between them involves significant amplitude changes.

So what happens when power sensors are presented with this sort of modulation?

Transient Behaviour of Diode Sensors



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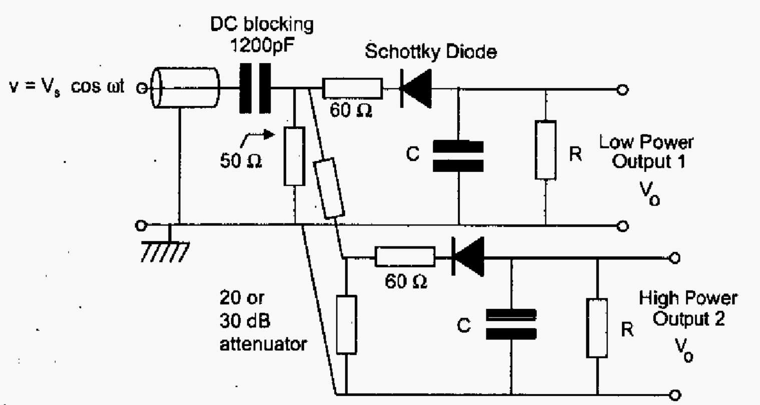
It is the nature of rectification which causes the diode to charge up the DC output circuit quickly when following an increase in power, and then to play little part when the RF level falls. Below –30dBm, the equivalent resistance of the diode is essentially constant. However, when measuring levels above +10 dBm, the equivalent resistance of the diode might be as low as 200R when charging and then about 30k when discharging.

This means that it responds to the peaks in the modulation and then remembers the peak value when the next trough occurs. So a diode-based sensor tends to measure a value bigger than it should.

The degree to which this happens depends on the general level of the RF, the ratio between peaks and troughs, and the modulation rate in comparison with the detector response time. When operating well beyond the square law region (above +10 dBm), the detector is behaving as a peak detector, whereas at small signal (less than -30 dBm), the detector operates nearly as well as the thermal types. In practice, measuring digital modulation such as Pi/4 DQPSK at high signal level with a diode detector will result in 1 to 2 dB error.

Wider Dynamic Range Diode Detection



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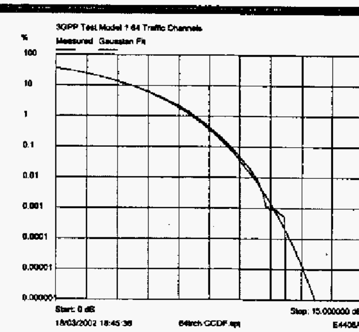
Recent detector types include a composite detector – attenuator – detector arrangement. Here, we choose the most appropriate detector for the power level being measured

Why does linearity matter?

- WCDMA modulation requires large peak to mean ratio
  - *Implies that non-linearised PAs require extremely high backoff*
  - *11.2 dB Peak - Mean implies 264 W peak for 20W average*
- Efficiency and thermal management implications



Measured

Measurement Parameter	Value
Average input power	-3.04 dBm
Probability at average input power level	36.25 %
Power level that has 10% of the power	3.66 dB
Power level that has 1% of the power	6.77 dB
Power level that has 0.1% of the power	8.61 dB
Power level that has 0.01% of the power	9.77 dB
Power level that has 0.001% of the power	10.42 dB
Power level that has 0.0001% of the power	11.20 dB
Peak power	11.20 dB
Count	3986592.00

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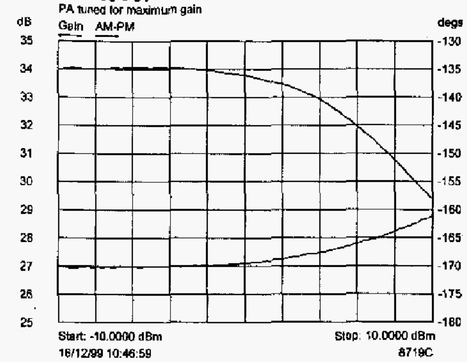
The simplest solution to overcoming large-signal effects is to operate the circuit at a low enough level to ensure linear operation. This works well in most circumstances, provided that noise does not become an issue. However, the price paid is in the component cost and current consumption of an over-specified signal headroom.

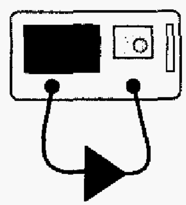
In power amplifier design, this become particularly acute, especially given the statistics of some modulation formats. For example, a W-CDMA base-station typically has amplitude excursions ranging from 70dB below to > 10dB above the mean level. To accommodate this dynamic range without clipping implies that the PA will always be operated with at least 10dB backoff, with correspondingly poor efficiency. Even then, achieving linear performance at the peaks requires careful design and of course a suitably large output device.

Next, we consider the tests used to characterise linearity in the presence of modulated signals.

AM-AM and AM-PM nonlinearities

- **Typical PA gain compression and phase change with input level**





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The key factor is that for components such as amplifiers, mixers, and ferrite isolators, the transfer function is not constant as the level of the signal is varied. Variations in transfer function with frequency are a small-signal issue, but even with a flat frequency response, we also need to consider level-dependency.

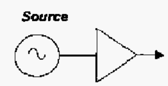
Consider a simple amplifier stage with an unmodulated sinewave RF signal at its input. While keeping the level constant, the phase of the sinewave is advanced by 90 degrees. Regardless of the linearity of the amplifier, the phase change will be precisely transferred to its output. The amplifier is unaware of, and unaffected by the absolute phase of the carrier, so we don't see any error incurred by phase modulation at the input.

Next, let us vary the amplitude of the input signal, while keeping the phase constant. This can be achieved using a network analyzer set to perform a power sweep (see graph above). There are two effects that can be observed. The first is gain compression, where the amplifier output has an upper limit to its capability, and gradually more and more of the wave form is clipped as the level is increased. So amplitude modulation of the carrier passing through the amplifier can be thought of as creating an additional AM component at the output due to the compression characteristic. This is known as AM to AM conversion.

The second effect is that the phase of the output relative to the input changes as the level is raised. This effect is known as AM to PM conversion.

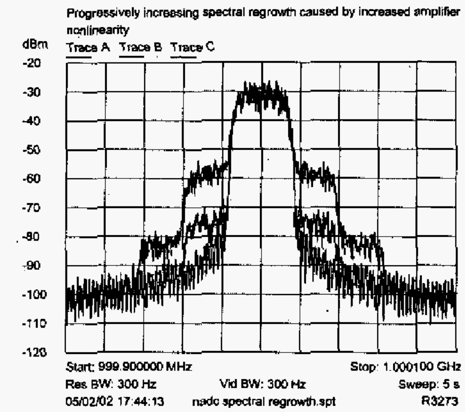
In conclusion, amplitude variations in a modulated carrier are going to create both amplitude and phase errors. Conversely a constant-amplitude modulation technique will not be subjected to these problems.

Spectral Regrowth



$$V(t) = \sum_{n=1} a_n \sin(2\pi f_n t)$$

- **Modulated signal is a represented by a sum of many sinewaves**
- **Non-linear amplification causes intermodulation from these components**



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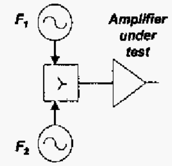
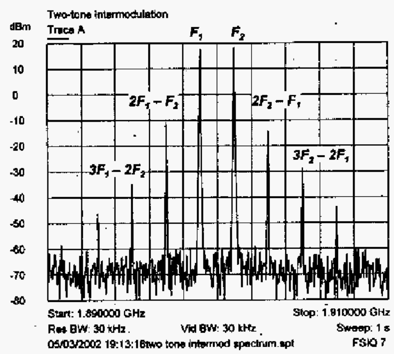
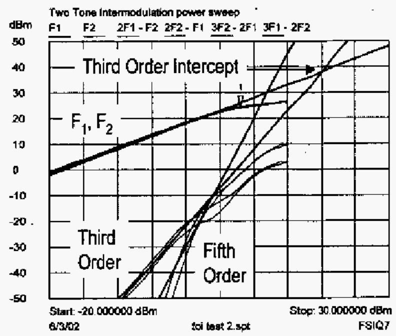
When a modulated carrier is passed through a non-linear amplifier, the effect of the non-linearity is apparent on the frequency spectrum of the modulation. Above, is an example showing the effect of differing amounts of distortion applied to an NADC modulated carrier (North American Digital Cellular, utilising Pi/4 DQPSK).

Increasing distortion gives rise to characteristic shoulders in the output spectrum. These shoulders are referred to as spectral regrowth. Spectral regrowth arises from mixing products of the constituent spectral components in the signal. Any modulated signal V(t) can be decomposed into an equivalent sum of sine waves with properly weighted coefficients. An unwanted product at a frequency (2fn – fm) results from mixing of the components fn and fm, and each of these possibilities within the modulation cause the first set of shoulders either side of the wanted signal. The next pair of shoulders arise from mixing of the form (3fn – 2fm). These are referred to as 3rd order and 5th order products respectively, from the form of the mixing equations.

Two-tone Intermodulation







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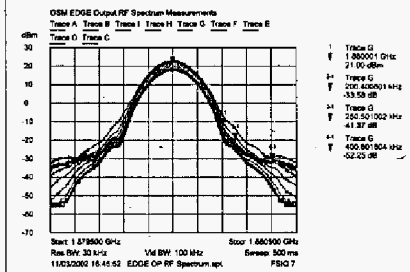
A simplified form of spectral regrowth can be observed when two CW sources are combined at the input of an amplifier. This measurement is used as a measure of amplifier or receiver linearity. The frequencies of each source are such that they are inside the passband of the device under test. The non-linearities in the amplifier will produce sum and difference products of the form (Nf1 – Mf2). The components (2f1 - f2) and (f1 - 2f2) are known as the third order products, and the components (3f1 - 2f2) and (2f1 - 3f2) are known as the fifth order products. Higher order products also exist but are generally negligible in comparison.

The test is performed by sweeping the levels of the f1 and f2 signals, and observing the levels of the third and fifth order components with a spectrum analyzer. By plotting the levels of each of the components as a function of input power level, it can be seen that the third order products rise at a rate of 3dB / dB and the fifth order products rise at a rate of 5dB / dB. The intercept point of the line through the third order products and the line through the un-mixed signals is known as the Third Order Intercept point (TOI).

When making this measurement, beware of unwanted mechanisms causing the intermodulation products being measured. These include signals from one source leaking through the combiner and reaching the output stages of the other source. Not only can the output amplifiers create intermodulation, but also the sum of the two signals has an AM component equal to the difference frequency. This can fall within the bandwidth of the source level control circuit, so it tries to react by applying an inverse AM component to the signal generator output.

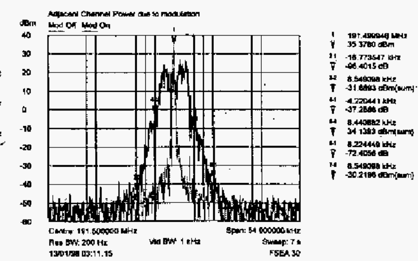
In practice, the isolation between the sources needs to be increased through the use of attenuators or isolators, to a comparable or lower level than the amount of intermodulation that is to be measured.

Adjacent Channel Power



Trace A

Measurement Parameter	Rel Lower	Abs Lower	Rel Upper	Abs Upper
Mod'n 0 Hz Offset	0 dB	-0.16 dBm	0 dB	-0.16 dBm
Mod'n 200.000 kHz Offset	-30.99 dB	-31.16 dBm	-31.06 dB	-31.26 dBm
Mod'n 250.000 kHz Offset	-39.59 dB	-39.77 dBm	-38.10 dB	-38.28 dBm
Mod'n 400.000 kHz Offset	-56.03 dB	-56.21 dBm	-54.25 dB	-54.44 dBm
Mod'n 600.000 kHz Offset	-62.70 dB	-62.89 dBm	-62.24 dB	-62.43 dBm
Mod'n 1.200 MHz Offset	-63.34 dB	-63.52 dBm	-61.38 dB	-61.57 dBm
Mod'n 1.800 MHz Offset	-79.26 dB	-79.43 dBm	-78.87 dB	-79.05 dBm



- Interested in leakage into other users' channels

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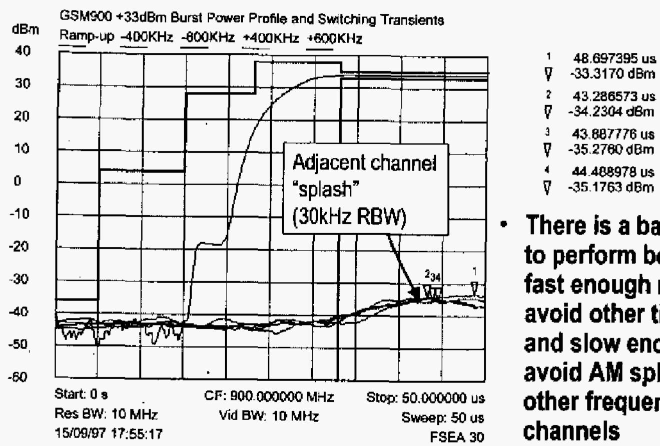
In a communication system having different users on different frequency channels, spectral regrowth becomes a problem when the transmitter belonging to one user starts to leak into another's frequency channel. This is especially important if the offending transmitter is interfering with a distant user whose signal is more prone to interference.

Adjacent channel interference from a transmitter is specified in one of two ways. The GSM system specifies a certain type of filter to be used (5-pole synchronously-tuned filter, 30kHz resolution bandwidth), and gives limits for the measurement obtained at various offsets. The graph above left shows an example of this.

Most systems specify integrated power in the adjacent channel (and other channels). This effectively permits the measurement filter shape to be mathematically constructed from a number of measurements with a much narrower filter. The measurements can be made either with a spectrum analyzer or with a vector signal analyzer. The results are then calculated by appropriate processing of the frequency or time data. For the spectrum analyzer, this involves converting each data point to a measurement in Watts/Hz, summing across the desired band, and converting back to dBm.

Power-Time Response



- **There is a balancing act to perform between a fast enough ramp to avoid other time slots, and slow enough to avoid AM splash in other frequency channels**

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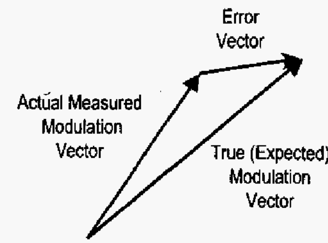
Transmitters switch on and off, and the transmit level is frequently adjusted to be the minimum that achieves the desired operating performance. Consequently, there is a need to characterise the setting accuracy of the desired power levels, and also to measure the characteristics of the power ramping profile.

The main issue with transiently changing levels is the generation of AM sidebands due to the level change. This will affect users in adjacent frequency channels. There will be a specification limit on permissible levels due to switching transients for a given standard. A compromise has to be made between stepping rapidly enough to avoid interfering with users in preceding of following time slots, and stepping slowly enough to avoid interfering with users on other frequency channels. The profile of the ramp is designed carefully to minimise both of these issues, and it has to be measured in both time and frequency domains to ensure compliance.

The above example is of a GSM handset transmit slot. Only the rising edge of the burst is shown, but a similar measurement has to be made on the falling edge. The measurement is performed with a spectrum analyzer set to the correct frequency channel, and zero span. Effectively it measures power vs time in this mode, and the sweep rate and measurement bandwidth are set high enough to capture all of the burst profile. Too low a resolution bandwidth would cause the risetime to be overestimated. The instrument mode is modified slightly to measure the adjacent channel switching transients ("splash"). The analyzer is tuned to the appropriate offset and the resolution bandwidth is lowered so that the on-frequency component is rejected. The GSM spec also calls for the spectrum analyzer to be set to max hold for this measurement.

Error Vector Magnitude (EVM)



- *EVM* – **the magnitude of the error vector between the vector representing the actual transmitted signal and the vector representing the error-free modulated signal**

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EVM is used as a figure of merit for all types of modulation, but is especially useful for those involving both amplitude and phase (or frequency) components.

If the measured modulation vector is drawn on an I-Q diagram, and the expected or desired vector is also drawn, the "length" between the ends of each vector is an indication of the modulation accuracy. Zero EVM implies a perfect modulation, and specifications in the region 5% to 20% are common.

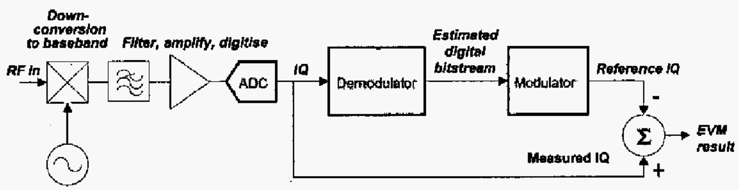
Although EVM can be continuously evaluated, it is normally only reported at the sampling instances in time where a symbol decision is made. Then the EVM values for each symbol are summarised by averaging or reporting of peaks.

For slot-based communications standards, the EVM is calculated for all the symbols in each slot, and summary results for each slot are produced.

The error vector for a real measured signal is influenced by a number of effects, many of which are tracked out by the receiver. Consequently, these are also removed from the measurement by a Vector Signal Analyzer. These effects include optimisation of symbol timing, amplitude scaling, amplitude droop across the slot and frequency offset.

EVM Measurement

- **A Vector Signal Analyzer demodulates and re-modulates the measured signal to construct a reference signal**
  - ***The difference between measured and reference traces at symbol decision times is the EVM result***
  - ***For GSM and other TDMA systems, EVM is computed only in the useful part of the burst.***
  - ***Symbol timing and amplitude droop is chosen such that result is minimised***
  - ***The following corrections are optimised to minimise the EVM result :***
    - **Symbol timing offset, frequency error, amplitude scaling, amplitude droop (across a burst)**



A Vector Signal Analyzer (VSA) is used to measure EVM. It downconverts the RF signal to a baseband IQ representation and then digitises it. It then finds the difference between the measured and reference waveforms using IQ representation. First, it has to create the reference waveform, and for this, it uses knowledge of the modulation format and the symbol rate to demodulate the measurement and recover the digital data being transmitted. Having established the digital data, it is possible then to re-modulate it and create a numerically ideal version of the measured signal.

EVM does not include certain error components, including long-term symbol rate errors and scaling between the measured and reference waveforms. The VSA has to perform resampling of the digitised data, and find an optimum symbol rate and timing phase such that the EVM result is minimised. It also used a straight-line fit to model the general amplitude level and slope of the data. These are attributes normally expected of a demodulator in a real receiver.

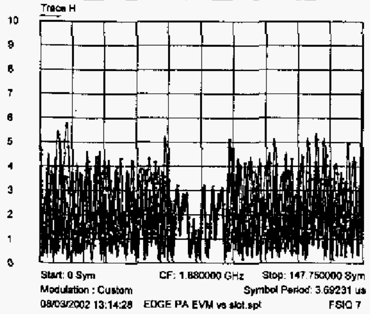
Types of EVM Measurement

- RMS EVM – ***is a scalar result computed once per burst, computed over 142 symbols. Averaged over a minimum of 200 bursts.***
- Max RMS EVM – ***is the maximum RMS EVM result obtained since measurement restart.***
- Peak EVM – ***the largest EVM within each burst is returned for this result. Averaging is done for this value over the number of bursts specified.***
- Max Peak EVM – ***is the maximum EVM that has occurred since measurement restart. This value is the worst-case outlier found over all EVM measurements.***
- 95th%tile EVM – ***is the EVM value below which 95 percent of all EVM values exist.***

Most Vector Signal Analyzers produce a number of summary EVM results from the individual values calculated at each symbol. Above are the definitions of the types of EVM result displayed for GSM EDGE.

GSM-EDGE EVM measurements

EDGE handset power amplifier EVM vs symbol in one slot



Trace A

Measurement Parameter	Value	Symbol No
Error Vector Magnitude (RMS)	2.11 %	
Error Vector Magnitude (Peak)	4.93 %	130
Magnitude Error (RMS)	1.19 %	
Magnitude Error (Peak)	-3.11 %	130
Phase Error (RMS)	1.30 degs	
Phase Error (Peak)	-10.13 degs	123
Frequency Error	27.31 Hz	
Peak Frequency Error	27.31 Hz	
Amplitude Droop per Symbol	8.11 dB	
IQ Offset	0.35 %	
IQ Imbalance	0.22 %	
Rho Factor	1.00	

- **Typical measurements provided by Vector Signal Analyzer**

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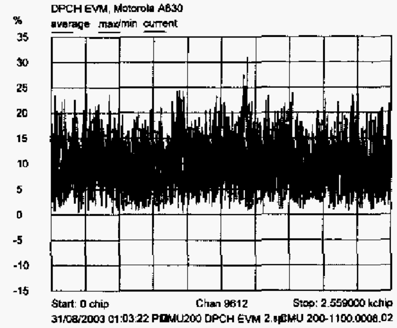
Modulation standards such as EDGE, W-CDMA, NADC require modulation accuracy to be measured in terms of EVM, since amplitude errors are just as important as phase errors. A vector signal analyzer will also produce measurement statistics indicating the magnitude and phase components or the error, and other metrics such as IQ imbalance and offset.

W-CDMA EVM

- **Handset Transmit Output**

DPCH EVM, Motorola A830



average

Measurement Parameter	Current	Average	MinMax
Peak EVM	20.68 %	22.07 %	26.76 %
RMS EVM	8.59 %	8.82 %	9.27 %
Peak Magnitude Error	-16.35 %	16.04 %	-16.84 %
RMS Magnitude Error	5.72 %	5.80 %	5.87 %
Peak Phase Error	-11.11 degs	12.11 degs	13.84 degs
RMS Phase Error	3.28 degs	3.75 degs	4.20 degs
IQ Origin Offset	-38.82 dB	-36.65 dB	-37.77 dB
Frequency Error	11.44 Hz	17.75 Hz	37.99 Hz
Peak Code Domain Error	-29.52 dB	-27.88 dB	-26.83 dB
UE Current Power	-32.03 dBm		
Out of Tolerance Power	0 %		
Slot Number	0		

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EVM is also measured on the HPSK output from 3GPP phones in a similar manner. Regeneration of the reference modulation is a more involved process, since the modulation contains data and control components which can be at different power levels.

EVM Specifications in Practice

- EDGE (Enhanced Data rates for GSM Evolution) modulation specification for Base-stations
  - *7% RMS EVM and 22% peak EVM, averaged over 200 transmission bursts.*
- In practice, a design target needs some production and drift/ageing margin
  - *eg 5% RMS EVM and 15% peak EVM*
- The key questions are:
  - *How do measurement uncertainties affect the target? Do they significantly tighten the target spec? (Typical test equipment has a data sheet accuracy of 1%.)*
  - *How do the EVM contributions of each stage of the transmitter add up? If there are 5 stages, do they all have to be 1% EVM each?*

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As with any other type of RF measurement, the accuracy of the test equipment needs to be taken into account when interpreting readings. Usually, a measurement uncertainty figure means the reading could be high or could be low, by the same maximum amount. However, EVM is a single-sided function – negative values cannot exist.

The issue of combining EVM values arises when cascading stages in a transmitter line-up as well as when accounting for EVM errors.

EVM Combination Mechanisms

- Systematic non-linearities can be cascaded and superposed such that sometimes worst-case addition occurs
  - *i.e. two stages that each have 1% peak EVM could cause 2% peak EVM if the non-linearities were suitably aligned.*
- Uncorrelated errors such as group delay distortion, phase noise, floor noise, and DAC quantisation can be treated in the same way as summing noise powers
  - *i.e. they add in a Root-Sum-Squares (RSS) fashion. E.g. two contributions, each 1%, would sum to give 1.414%. Importantly, this means that adding a small random error mechanism cannot reduce the overall EVM level.*
- Test equipment has a predominantly random measurement error, so it always increases the reading relative to the true value.
- Consequently it is safe to use the indicated EVM readings of a vector signal analyzer without adding margin for measurement uncertainty.

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Analysis shows that systematic non-linearities such as amplifier compression and AM-PM can be cascaded linearly, i.e. two stages that each have 1% peak EVM could cause 2% peak EVM if the non-linearities were suitably aligned. However, uncorrelated errors such as group delay distortion, phase noise, floor noise, and DAC quantisation can be treated in the same way as summing noise powers, i.e. they add in a Root-Sum-Squares (RSS) fashion. For example, two contributions, each 1%, would sum to give 1.414%.

In the case of test equipment measurement uncertainty, the equipment has a very small systematic component, since the measurement chain operates at a signal level with sufficient input attenuation to ensure it is linear. Since the measurement error is predominantly random, it always increases the reading relative to the true value. Consequently it is safe to use the indicated EVM readings of a vector signal analyzer without adding margin for measurement uncertainty.

EVM Uncertainty Guidelines

- Random errors always worsen the EVM
- Systematic errors might either add to, or subtract from, the contribution from the DUT.
  - *Test equipment characteristics are almost entirely random in nature, it is safe to take the indicated reading as being the true value without need to add any further measurement uncertainty.*
- Cascaded stages in an RF design have to be analyzed to determine the overall effect.
  - *worst-case error is unlikely, due to the non-linear relationship between amplitude compression and EVM, and also due to the potential for systematic errors to cancel out.*
- Validated for a variety of different modulation formats
- It is not permissible to linearly subtract an EVM measurement uncertainty from a measurement result in order to publish a smaller "compensated" result.

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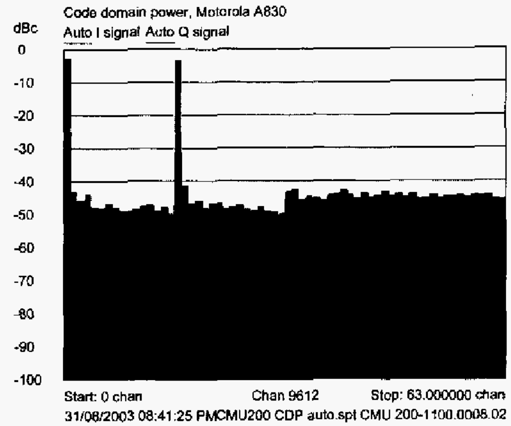
W-CDMA Code Domain Power



Auto I signal

Measurement Parameter	Value
EVM (RMS) Current	8.52 %
I/Q Origin Offset	-39.18 dB
Frequency Error	-42.11 Hz
UE Current Power	-32.74 dBm
Slot Number	0

Auto Q signal

Measurement Parameter	Value
EVM (RMS) Current	8.15 %
I/Q Origin Offset	-38.87 dB
Frequency Error	-71.87 Hz
UE Current Power	-32.77 dBm
Slot Number	0

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A Code Domain Power plot shows the power present for each available spreading code (typical plot above). The W-CDMA signal is first de-scrambled and then attempted to be de-spread using all possible spreading codes. This gives the code-domain equivalent of a frequency spectrum measurement, showing which of the available channels are in use.

Peak Code Domain Error (PCDE)

- EVM measures modulation accuracy in composite RF signal at QPSK
- Need to analyse accuracy of each code channel
  - *The Peak Code Domain Error is computed by projecting the power of the error vector onto the code domain at a specific spreading factor. The Code Domain Error for every code in the domain is defined as the ratio of the mean power of the projection onto that code, to the mean power of the composite reference waveform, expressed in dB. The Peak Code Domain Error is defined as the maximum value for the Code Domain Error for all codes.*
  - *The measurement interval is one power control group (timeslot).*

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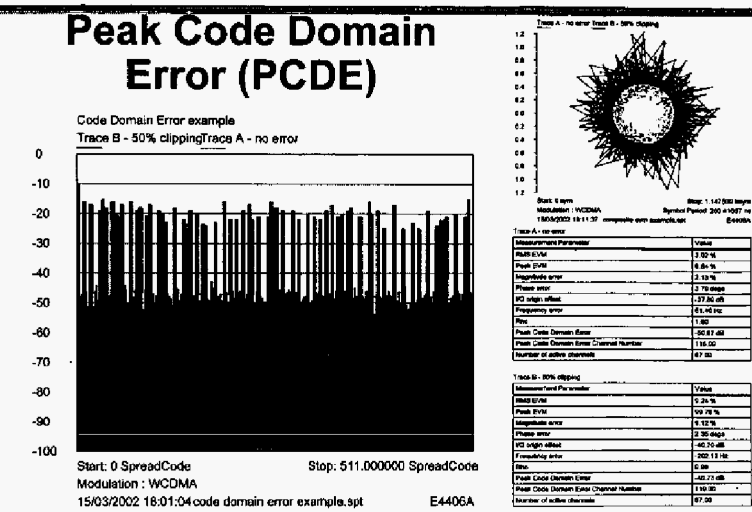
Peak Code Domain Error (PCDE)



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While Composite EVM gives an overall result in a W-CDMA system, the IQ representation of the transmit signal or even of the EVM vector is difficult to use for diagnostic purposes. Therefore, code domain analysis of the modulation error is used to gain a better understanding. This reveals the modulation quality available to each individual user. More importantly, code-domain errors cause leakage into other users' channels, reducing their signal to interference ratio.

Code Domain Error is calculated from the composite EVM measurement samples, i.e. the difference vector between the measured and reference IQ data is processed using code domain analysis. The error vector data is de-scrambled and then de-spread using each of the available spreading codes. The spreading code having the greatest error is reported as the peak code domain error. The result is ratioed to the mean power of the original signal and expressed in dB. The Peak Code Domain Error spec limit for a base-station is -33dB at spreading factor 256.

Modulation Measurements (Receiver)

- BER
  - *Bit errors, block errors, frame erasure, sync errors.*
  - *Typical measurement system, including loopback mode*
- Sensitivity
  - *Definitions, e.g. 1E-3 BER point*
- Selectivity Measurements
- Spurious Response Measurements
  - *Measurement techniques – sweep of test interferer, measuring analogue IF / IQ / RSSI level*

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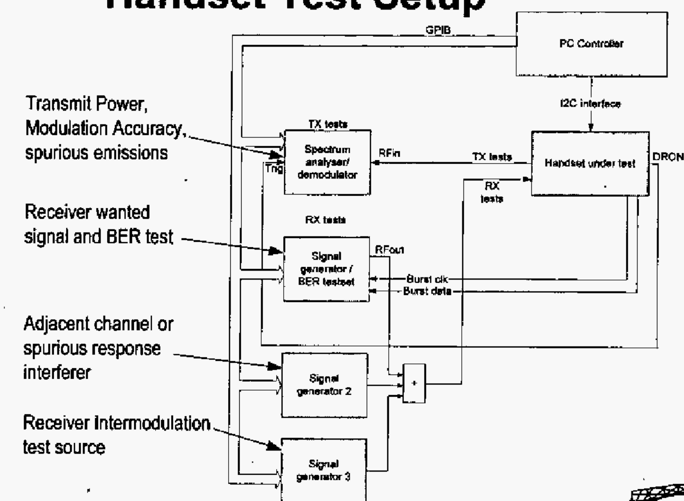
Handset Test Setup



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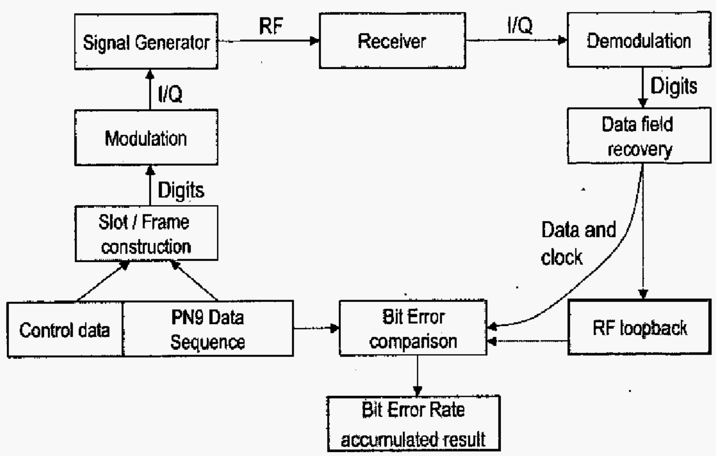
In practice, testing of a handset for receiver or transmitter parameters involves having an RF connection in place of the antenna, a means of configuring it for the various tests, and access to the received data. Above is a typical configuration for R&D testing, where the received data and associated clock are available on test pins of the baseband chip. Sometimes it is necessary to re-transmit the received data when access to the baseband is not possible. The handset is controlled from a PC by means of an I2C or serial interface. This configures the handset into a test mode, where the normal protocol stack is bypassed. It is then possible to control the registers of the baseband chip to force transmission and reception on particular slots or at particular levels.

Around the handset under test, a combination of sources are required to provide the wanted and interferer signal components. These are combined such that the sources have sufficient isolation between themselves to avoid interaction in the sources. 40dB isolation from one source to any other is a typical practical requirement.

Bit Error Rate testing is commonly used to check the integrity of a complete RF transceiver link, but is of particular interest to RF engineers when integrating the Receiver RF section with the baseband section of a system. The composite performance of receiver and baseband can be quantified by BER testing, and it is used to check performance in a number of adverse condition scenarios.

Bit Error Rate Testing



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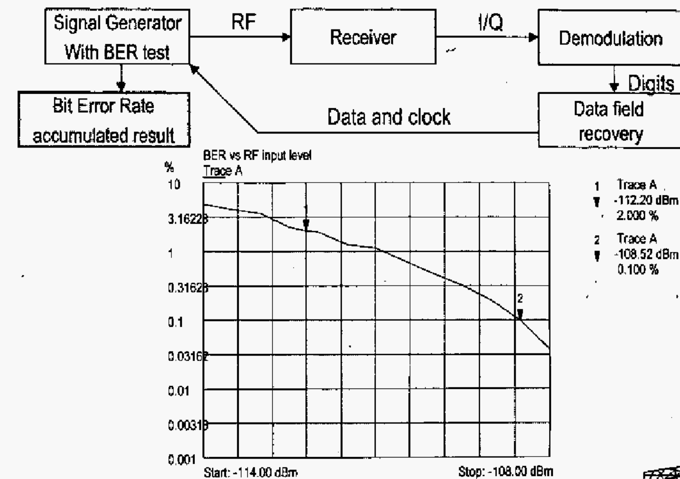
The core of the test is to transmit a known data pattern, and compare it data bit by data bit with the received data. The proportion of incorrect bits in the total number of bits tested is the bit error rate. Normally the test is run until a statistically significant number of errors is recorded, for example 100 errored bits. BER is expressed as a ratio or as a percentage, e.g. 1E-3 or 0.1%. If the link fails completely, the likelihood of a random bitstream being correct is 50%, so the highest BER normally measurable is 50%. Measurements nearing 100% would indicate that the measured data is being inverted!

In practice, synchronisation between the transmitted data and the received data is difficult, due to the need to format the test data into a standard packet structure, and the propagation delays through the RF hardware. The solution is to use a pseudo-random sequence, where it is possible to recreate the entire sequence at the receiving end given a sufficiently error-free section of data. Binary-M sequences are commonly used, having a repeating pattern of 2n –1 bits. Standard sequence definitions named PN9, PN15 and PN23 are commonly specified.

The transmitted sequence is cut into sections according to the size of the payload section of the transmitted slot, control information, checksum and guard bits are added, and the slots are assembled into standard frames. Effectively, the pseudo-random sequence replaces the speech or user data that would normally be transmitted. The digital data is then modulated onto an RF carrier and applied to the input of the receiver under test. The received data is demodulated and the data portion is extracted again from each slot. It is then re-synchronised to the original sequence such that the errors are minimised, and the error rate is calculated.

For common wireless standards, the PN9 generation, transmission, and BER testing is combined into a single piece of test equipment (signal generator or radio test set). The return signal from the receiver can in many cases be re-transmitted at high level without adding further errors, permitting a simplified RF link to the unit under test.

Receiver Sensitivity



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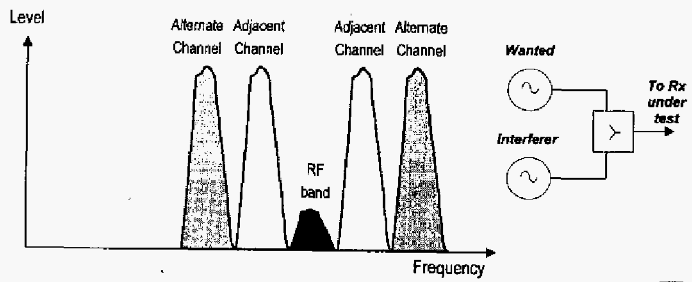
A primary requirement in receiver testing is to establish its sensitivity. This involves measuring BER as a function of input power level. The above example shows a power sweep where the 2% and 0.1% BER data points are of interest.

Generally, very low level measurements are being made, and care has to be taken to maintain the integrity of the wanted signal. Often it is necessary to use a screened room to keep interfering signals out, and leakage paths can cause the signal at the receiver to be higher than assumed.

When performing a power sweep, it is easier to establish the call at a higher level and gradually lower it until the call is dropped. However, some specifications call for raising the signal level upwards from the noise floor, to characterise the performance during fading conditions. It is possible to obtain different results by arriving at a measurement level from above or below, due to the way that synchronisation and equalisation is performed in the baseband.

Adjacent / Alternate Channel Selectivity

- Set the wanted signal to e.g. 6dB above sensitivity level
- Test with the interferer at each adjacent and alternate channel, adjust the level until e.g. 0.1% BER



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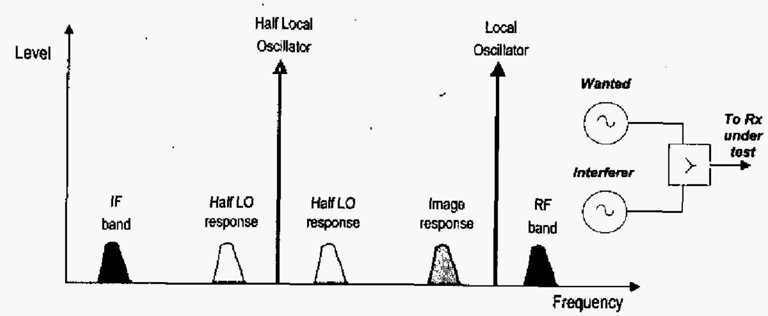
Selectivity is a measure of the receiver's ability to receive a desired signal while rejecting a strong signal in an adjacent channel or alternate channel. The test is performed with two modulated signal sources, combined together and applied to the input of the receiver under test. One signal generator provides a test signal at the desired channel frequency at a level typically 6 dB above the sensitivity of the receiver. The second signal generator provides either the adjacent channel signal, offset by one channel spacing, or the alternate channel signal, offset by two channel spacings. The level of the out-of-channel signal is increased until the sensitivity is degraded to a specified level, typically 0.1% Bit Error Rate.

In order to make accurate measurements of selectivity, the two sources need good level and modulation accuracy. The most important characteristic is the spectral accuracy of the modulation sidebands, particularly for the adjacent-channel source.

Spurious Response Measurements

- Undesired RF frequencies can mix with LO1, LO2 or baseband *clock frequencies and their harmonics / sub-harmonics*



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A spurious response measurement involves exposing the receiver to an interferer signal and observing the ability to receive a wanted signal.

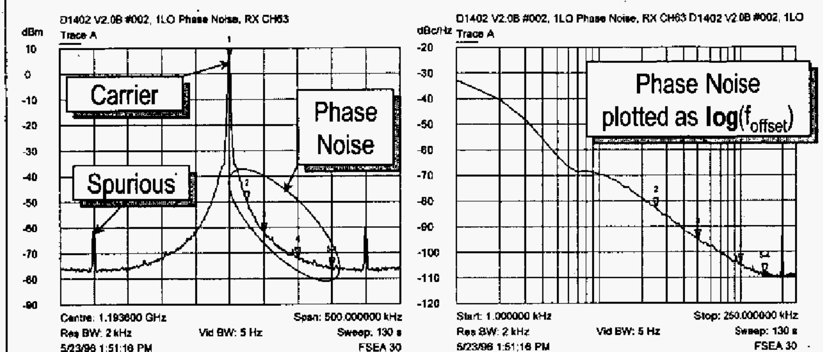
To make this measurement requires two sources, one provides a modulated test signal at the desired on-channel frequency at a level 6 to 10 dB above the sensitivity of the receiver. The second source outputs an interfering signal over a broad range of frequencies. The interfering signal may be modulated or unmodulated, depending upon the frequency range and the communication standard.

The frequency of the interfering signal is swept first over a wide frequency range (often up to 12GHz for 0.9 to 2.4 GHz systems). The level of the interferer is set initially to 10dB above the required specification limit, and the interferer is stepped in small increments (e.g. 0.5x channel spacing). At each search frequency a BER measurement is made, and this first sweep is designed to find "weaknesses" in the receiver. This sweep can take many hours, due to the narrow steps and wide frequency range. Then at each frequency where a response was found, a more detailed power sweep is performed, to characterise the level at which a failure in BER occurs. The difference between the test signal and the interfering signal is the spurious immunity of the receiver.

Occasionally, surprising interferer response frequencies occur, and it becomes necessary to calculate the frequency conversion mechanism in order to find a solution to the problem. This may involve mixing of harmonics of the RF with harmonics or sub-harmonics of the Local Oscillator. Sometimes, obscure effects involving both the First and the second Local Oscillator, or even the baseband processor clock can cause unwanted mixing effects that give an IF signal product. The 5th harmonic of the 2LO and the 14th harmonic of the TCXO reference have been known to cause spurious responses!

Phase Noise

- A measurement of oscillator purity (dBc/Hz)
  - *Affects selectivity of receivers and spurious emissions from transmitters*



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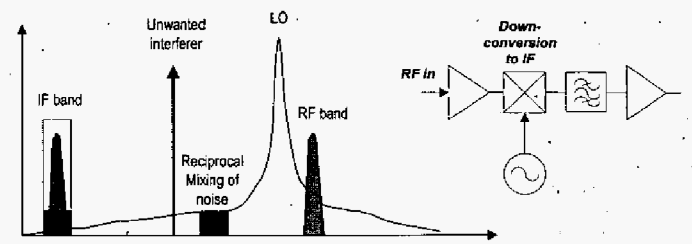
For a theoretical sinewave oscillator, all of the power is concentrated at a single frequency. In practice, random noise within the oscillator components will perturb the sinewave and cause the power to be spread over a band of frequencies. Noise in oscillator circuits predominantly modulate the phase of the oscillation rather than the amplitude due to clipping in the active device. Therefore the effect is referred to as phase noise and is modelled as a random phase modulation.

Phase noise is expressed in units of dBc/Hz. It is generally displayed on a dBc/Hz vs log-frequency axis. This enables both the close-in phase noise (offsets < 1 kHz) and the far-out phase noise (offsets > 10 kHz) to be easily examined on one plot. Key specifications for oscillators are given at one or more frequency offsets from the fundamental frequency.

The Effects of Phase Noise



- **Unwanted signals mix with phase noise profile to desensitise the receiver**
  - *Known as reciprocal mixing*

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A receiver performs the task of downconverting the RF signal to an intermediate frequency (IF) or baseband frequency such that it can be demodulated. A Local Oscillator is used in each frequency conversion stage, where a mixer performs a frequency difference operation between the LO and the RF signal.

Unfortunately there are many unwanted scenarios where the combination of frequencies result in a product that lies in the IF passband. These affect the selectivity of the receiver. An ideal receiver would be immune to signals outside the intended RF band. However, this will not be true if the LO is broader than a single pure tone. Any two frequencies having the correct difference will cause an output in the IF band. This is even true if one of the components is phase noise from the LO. The level of the IF mixing product is governed by the smaller of the two RF signals, and in the case of an interferer mixing with LO phase noise, it is the phase noise that determines the level. This is opposite to the norm, in the sense that the LO is usually the higher amplitude signal, and merely facilitates frequency translation of the other RF signal. Consequently, this scenario is referred to as reciprocal mixing.

The consequence of reciprocal mixing is that interfering signals produce unwanted noise at the output of the receiver, reducing sensitivity to the wanted signal. This effect can only be reduced by improving the phase noise of the Local Oscillator.

Conclusions

- System link block diagram – modulate, tx, channel, rx, demodulate
- Why Digital? Resistance to fading, voice vs packet data, capacity
- IQ modulation representation, constellation, eye diagram display formats
- BPSK, QPSK, MSK, properties of gaussian and RRC filtering, concept of ISI
- Spread spectrum, OFDM
- TDMA, FDMA, CDMA definitions
  - *TDD, FDD*
- Constant envelope modulation examples
  - *Bluetooth, GSM*
- Non-constant Envelope Modulation examples
  - *EDGE, W-CDMA, 802.11b (complementary code keying - spread spectrum), 802.11a (OFDM)*
- Transmitter and Receiver Measurements