NEAR EAST UNIVERSITY

FACULITY OF ENGINEERING

Department of Electrical and Electronic Engineering

DİGİTAL MODULATİON TECHNİQUES

Graduation Project
EE-400

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As well as I would like to thanks my parents and my family for supporting and encouraging me accomplish my dream under their custody. And I want to handset this project to my family because without their endless support and love for me, I would never achieve my current position. I wish my mother lives happy always and my father in the heaven is proud of me.

Also my especial thanks go to my brothers, shaher and shaheer .Under their guidance; I successfully overcome many difficulties and learn a lot from them. They are always helps me a lot either in my study or my life.

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Abstract

This project focuses in the application of digital modulation techniques used in many communications systems today. Emphasis is placed on explaining the tradeoffs that are made to optimize efficiencies in system design.

The move to digital modulation provides more information capacity, compatibility with digital data services, higher data security, better quality communications, and quicker system availability.

Also in my project I am going to discuss the techniques of digital modulation have revolutionized the communication industry, and I will discusses the theory behind digital communication techniques, shows how this theory is applied to electronic devices, and demonstrates their functionality in a real instrumentation application.

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CHAPTER ONE

DIGITAL MODULATION

Introduction

This project introduces the concepts of digital modulation used in many communications systems today. Emphasis is placed on explaining the tradeoffs that are made to optimize efficiencies in system design. Most communications systems fall into one of three categories: bandwidth efficient, power efficient, or cost efficient. Bandwidth efficiency describes the ability of a modulation scheme to accommodate data within a limited bandwidth. Power efficiency describes the ability of the system to reliably send information at the lowest practical power level. In most systems, there is a high priority on bandwidth efficiency. The parameter to be optimized depends on the demands of the particular system. For designers of digital terrestrial microwave radios, their highest priority is good bandwidth efficiency with low bit -error-rate. They have plenty of power available and are not concerned with power efficiency. They are not especially concerned with receiver cost or complexity because they do not have to build large numbers of them.

On the other hand, designers of hand-held cellular phones put a high priority on power efficiency because these phones need to run on a battery. Cost is also a high priority because cellular phones must be low-cost to encourage more users. Accordingly, these systems sacrifice some bandwidth efficiency to get power and cost efficiency.

Every time one of these efficiency parameters (bandwidth, power, or cost) is increased, another one decrease, becomes more complex, or does not perform well in a poor environment. Cost is a dominant system priority. Low-cost radios will always be in demand. In the past, it was possible to make a radio low-cost by sacrificing power and bandwidth efficiency. This is no longer possible. The radio spectrum is very valuable and operators who do not use the spectrum efficiently could lose their existing licenses or lose out in the competition for new ones. These are the tradeoffs that must be considered in digital RF communications design. This project covers:

1. The reasons for the move to digital modulation.
2. How information is modulated onto in-phase (*I*) and quadrate (*Q*) signals.
3. Different types of digital modulation.
4. Filtering techniques to conserve bandwidth.
5. Ways of looking at digitally modulated signals.
6. Multiplexing techniques used to share the transmission channel.
7. How a digital transmitter and receiver work.
8. Measurements on digital RF communications systems.
9. An overview table with key specifications for the major digital communications systems; and a glossary of terms used in digital RF communications.

These concepts form the building blocks of any communications system. If we understand the building blocks, then you will be able to understand how any communications system, present or future, works.

1.1 Digital Modulation Mechanism

The move to digital modulation provides more information capacity, compatibility with digital data services, higher data security, better quality communications, and quicker system availability.

Developers of communications systems face these constraints:

1. Available bandwidth.
2. Permissible power.
3. Inherent noise level of the system.

The RF spectrum must be shared, yet every day there are more users for that spectrum as demand for communications services increases. Digital modulation schemes have greater capacity to convey large amounts of information than analog modulation schemes.

1.1.1 Trading off Simplicity and Bandwidth

There is a fundamental tradeoff in communication sy stems. Simple hardware can be used in transmitters and receivers to communicate information.

However, this uses a lot of spectrum, which limits the number of users. Alternatively, more complex transmitters and receivers can be used to transmit the same in formation over less bandwidth. The transition to more and more spectrally efficient transmission techniques requires more and more complex hardware. Complex hardware is difficult to design, test, and build. This tradeoff exists whether communication is ove r air or wire, analog or digital.

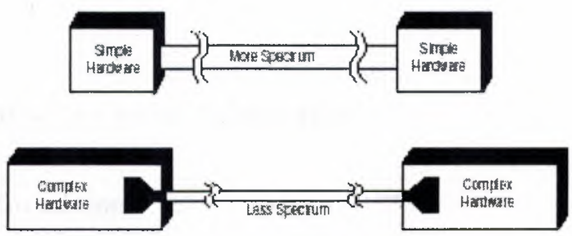


Figure 1.1.1.1: The Fundamental Tradeoff.

1.1.2 Industry Trends

Over the past few years a major transition has occurred from simple analog Amplitude Modulation (AM) and Frequency/Phase Modulation (FM/PM) to new digital modulation techniques. Examples of digital modulation include:

1. QPSK (Quadrate Phase Shift Keying).
2. FSK (Frequency Shift Keying).
3. MSK (Minimum Shift Keying).
4. QAM (Quadrate Amplitude Modulation).

Another layer of complexity in many new systems is multiplexing. Two principal types of multiplexing (or "multiple access") are TDMA (Time Division Multiple Access) and CDMA (Code Division Multiple Access). These are two different ways to add diversity to signals allowing different signals to be separated from one another

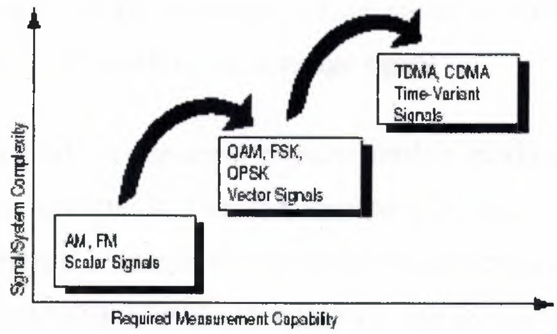


Figure 1.1.2.1: Trends in the Industry.

1.2 Using I/Q Modulation to Convey Information

1.2.1 Transmitting Information

To transmit a signal over the air, there are three main steps:

1. A pure carrier is generated at the transmitter.

2. The carrier is modulated with the information to be transmitted. Any reliably detectable change in signal characteristics can carry information.

3. At the receiver the signal modifications or changes are detected and demodulated.

1.2.2 Signal Characteristics that can be modified

There are only three characteristics of a signal that can be changed over time: amplitude, phase, or frequency. However, phase and frequency are just different ways to view or measure the same signal change.

In AM, the amplitude of a high-frequency carrier signal is varied in proportion to the instantaneous amplitude of the modulating message signal.

Frequency Modulation (FM) is the most popular analog modulation technique used in mobile communications systems. In FM, the amplitude of the modulating carrier is kept constant while its frequency is varied by the modulating message signal. Amplitude and phase can be modulated simultaneously and separately, but this is difficult to generate,

And especially difficult to detect. Instead, in practical systems the signal is separated into

Another set of independent components: *I* (In phase) and *Q* (Quadrate). These components are orthogonal and do not interfere with each other.

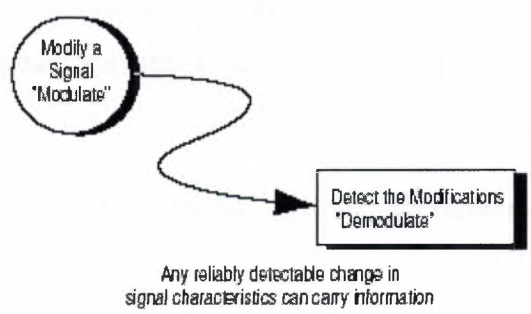


Figure 1.2.2.1: Transmitting Information (analog or digital).

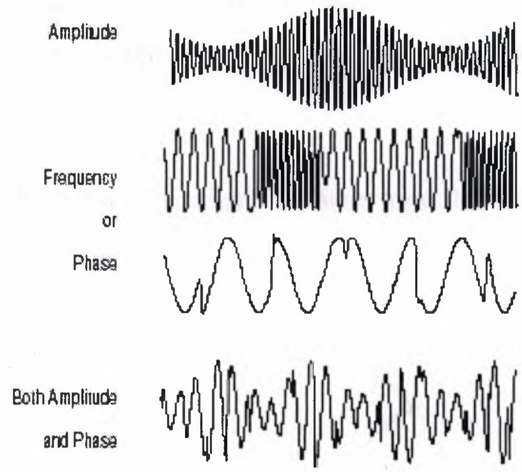


Figure 1.2.2.2: Signal Characteristics to Modify.

Signal Characteristics that can be modified a simple way to view amplitude and phase is with the polar diagram. The carrier becomes a frequency and p hase reference and the signal is interpreted relative to the carrier. The signal can be expressed in polar form as a magnitude and a phase. The phase is relative to a reference signal, the carrier in most communication systems. The magnitude is either an absolute or relative value. Both are used in digital communication systems. Polar diagrams are the basis of many displays used in digital communications,

Although it is common to describe the signal vector by its rectangular coordinates of I (In-phase) and Q (Quadrate).

1.2.3 Signal Changes or Modifications in Polar Form

Figure 1.2.3.1 shows different forms of modulation in polar form. Magnitude is represented as the distance from the center and phase is represented as the angle.

Amplitude modulation (AM) changes only the magnitude of the signal. Phase modulation (PM) changes only the phase of the signal. Amplitude and phase modulation can be used together.

Frequency modulation (FM) looks similar to phase modulation, though frequency is the controlled parameter, rather than relative phase.

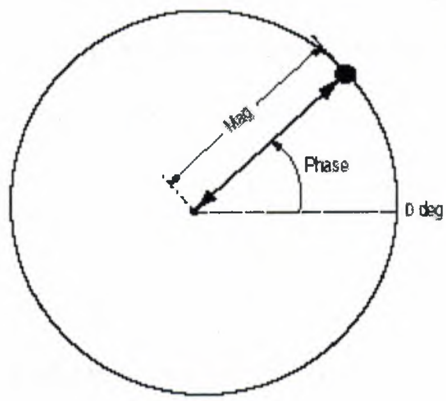


Figure 1.2.3.1: Polar Display Magnitude and Phase Represented Together

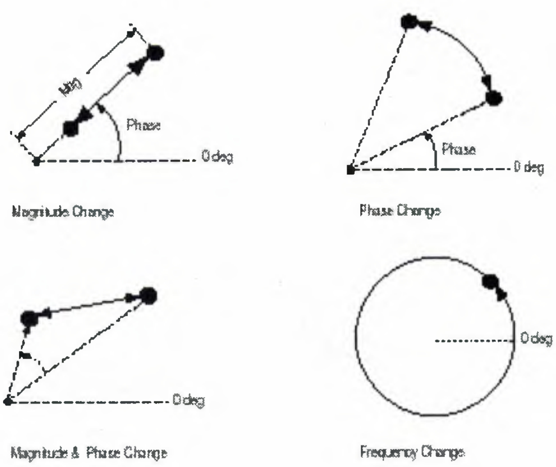


Figure 1.2.3.2: Signal Changes or Modification.

One example of the difficulties in RF design can be illustrated with simple amplitude modulation. Generating AM with no associated angular modulation should result in a straight line on a polar display. This line should run from the origin to some peak radius or amplitude value. In practice, however, the line is not straight. The amplitude modulation itself often can cause a small amount of unwanted phase modulation. The result is a curved line. It could also be a loop if there is any hysterics in the system transfer function. Some amount of this distortion is inevitable in any system where modulation causes amplitude changes. Therefore, the degree of effective amplitude modulation in a system will affect some distortion parameters.

1.2.4 *I/Q* Formats

In digital communications, modulation is often expressed in terms of I and Q. This is a rectangular representation of the polar diagram. On a polar diagram, the I axis lies on the zero degree phase reference, and the Q axis is rotated by 90 degrees. The signal vector's projection onto the I axis is its "I" component and the projection onto the Q axis is its "Q" component.

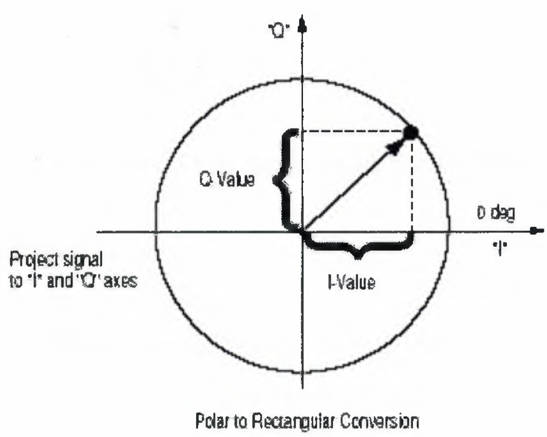


Figure 1.2.4.1: I_Q Format

1.2.5 *I* and *Q* in a Radio Transmitter

I/Q diagrams are particularly useful because they mirror the way most digital communications signals are created using an *I/Q* modulator. In the transmitter, Q signals and I are mixed with the same local oscillator (LO). A 90 degree phase shifter is placed in one of the LO paths. Signals that are separated by 90 degrees are also known as being orthogonal to each other or in quadrature.

Signals that are in quadrature do not interfere with each other. They are two independent components of the signal. When recombined, they are summed to a composite output signal. There are two independent signals in *I* and *Q* that can be sent and received with simple circuits. This simplifies the design of digital radios. The main advantage of *I/Q* modulation is the symmetric ease of combining independent signal components into a single composite signal and later splitting such a composite signal into its independent component parts.

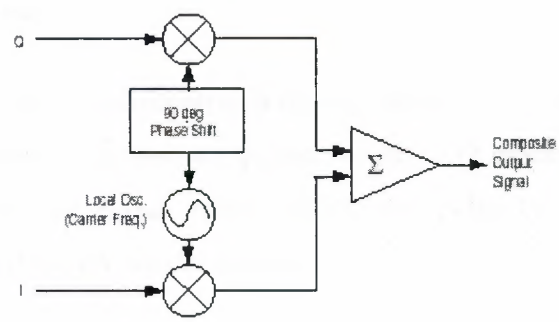


Figure 1.2.5.1*:* I and Q in A practical Radio Transmitter

1.2.6 I and Q in a Radio Receiver

The composite signal with magnitude and phase (or *I* and *Q*) information arrives at the receiver input. The input signal is mixed with the local oscillator signal at the carrier frequency in two forms. One is at an arbitrary zero phases. The other has a 90 -degree phase shift. The composite input signal (in terms of magnitude and phase) is thus broken into an in-phase, a quadrate, Q, and I component. These two components of the sign al are independent and orthogonal. One can be changed without affecting the other. Normally, information cannot be plotted in a polar format and reinterpreted as rectangular values without doing a polar-to-rectangular conversion. This conversion is exactly what is done by the in-phase and quadrature mixing processes in a digital radio.

A local oscillator, phase shifter, and two mixers can perform the conversion accurately and efficiently.

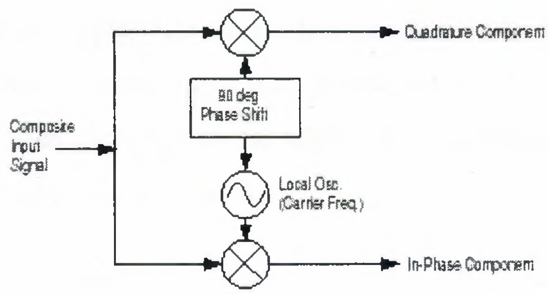


Figure 1.2.6.1: I and Q in A practical Radio Receiver

1.2.7 Q and I Mechanism

Digital modulation is easy to accomplish with *I/Q* modulators. Most digital modulation maps the data to a number of discrete points on the *I/Q* plane. These are known as constellation points. As the signal moves from one point to another, simultaneous amplitude and phase modulation usually results.

To accomplish this with amplitude modulator and a phase modulator is difficult and complex. It s also impossible with a conventional phase modulator. he signal may, in principle, circle the origin n one direction forever, necessitating infinite phase hafting capability. Alternatively, simultaneous AM and Phase Modulation is easy with an *I/Q* modulator. he *I* and *Q* control signals are bounded, but infinite has wrap is possible by properly phasing he *I* and *Q* signals.

1.3 Digital Modulation Types and Relative Efficiencies

This section covers the main digital modulation formats, their main applications, relative spectral efficiencies, and some variations of the main modulation types as used in practical systems. Fortunately, there are a limited number of modulation types, which form the building blocks of any system.

1.3.1 Applications

The table below covers the applications for different modulation formats in both wireless communications and video. Although th is note focuses on wireless communications, video applications have also been included in the table for completeness and because of their similarity to other wireless communications.

1.3.1.1 Bit rate and symbol rate

To understand and compare different modu lation format efficiencies, it is important to first understand the difference between bit rate and symbol rate. The signal bandwidth for the communications channel needed depends on the symbol rate, not on the bit rate.

$$sambol = \frac{\text{bit rate}}{\text{the number of bit trasmitted with each symbol}} \quad \ldots\ldots\ldots\ (1.3.1.1.1)$$

Modulation format	Application
MSK, GMSK	GSM, CDPD
BPSK	Deep space telemetry, cable modems
QPSK, $\pi/4$ DQPSK	Satellite, CDMA, NADC, TETRA, PHS, PDC, LMDS, DVB-S, cable (return path), cable modems, TFTS
OQPSK	CDMA, satellite
FSK, GFSK	DECT, paging, RAM mobile data, AMPS, CT2, ERMES, land mobile, public safety
8, 16 VSB	North American digital TV (ATV), broadcast, cable
8PSK	Satellite, aircraft, telemetry pilots for monitoring broadband video systems
16 QAM	Microwave digital radio, modems, DVB-C, DVB-T
32 QAM	Terrestrial microwave, DVB-T
64 QAM	DVB-C, modems, broadband set top boxes, MMDS
256 QAM	Modems, DVB-C (Europe), Digital Video (US)

Table 1.3.1.1.1: Application for Different Modulation Formats in both Wireless Communication and Video

Bit rate is the frequency of a system bit stream. Take, for example, a radio with an 8-bit sampler, sampling at 10 kHz for voice. The bit rate, the basic bit stream rate in the radio, would be eight bits multiplied by 10K samples per second or 80 Kbits per second. (For the moment we will ignore the extra bits required for synchronization, error correction, e Figure 1.3.1.2.1 is an example of a state diagram of a Quadrature Phase Shift (QPSK) signal. The states can be mapped to zeros and ones. This is a com but it is not the only one. Any mapping can be used.

The symbol rate is the bit rate divided by the number of bits t each symbol. If one bit is transmitted per symbol, as w would be the same as the bit rate of 80 Kbits per se symbol, as in QPSK, then the symbol rate wo second. Symbol rate is sometimes called b bit rate. These terms are often confuse same amount of data can be sen that are more complex and over a narrower piece of t

1.3.1.2 Spectrum (bandwidth) requirements

An example of how symbol rate influences spectrum requirements can be seen in eight-state Phase Shift Keying (8PSK). It is a variation of PSK. There are eight possible states that the signal can transition to at any time. The phase of the signal can take any of eight values at any symbol time. Since 23 = 8, there are three bits per symbol. This means the symbol rate is one third of the bit rate. This is relatively easy to decode.

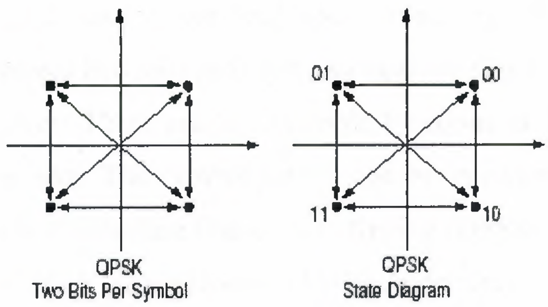


Figure 1.3.1.2.1: Bit Rate and Symbol Rate

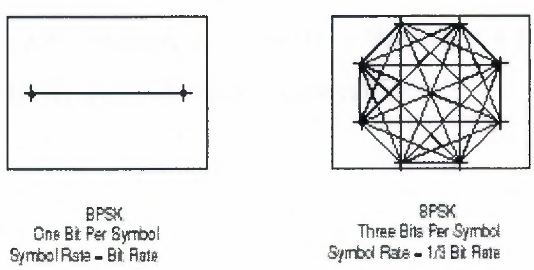


Figure 1.3.1.2.2: Spectrum Requirements

1.3.1.3 Symbol Clock

The symbol clock represents the frequency and exact timing of the transmission of the individual symbols. At the symbol clock transitions, the transmitted carrier is at the correct *I/Q* (or magnitude/ phase) value to represent a specific symbol (a specific point in the constellation).

1.3.2 Phase Shift Keying

One of the simplest forms of digital modulation is binary or Bi-Phase Shift Keying (BPSK). One application where this is used is for deep space telemetry. The phase of a constant amplitude carrier signal moves between zero and 180 degrees. On an *I* and *Q* diagram, the *I* state has two different values. There are two possible locations in the state diagram, so a binary one or zero can be sent. The symbol rate is one bit per symbol. A more common type of phase modulation is Quadrature Phase Shift Keying (QPSK). It is used extensively in applications including CDMA (Code Division Multiple Access) cellular service, wireless local loop, Iridium (a voice/data satellite system) and DVB-S (Digital Video Broadcasting — Satellite).

Quadrature means that the signal shifts between phase states, which are separated by 90 degrees. The signal shifts in increments of 90 degrees from 45 to 135, –45, or –135 degrees. These points are chosen as they can be easily implemented using an *I/Q* modulator. Only two *I* values and two *Q* values are needed and this gives two bits per symbol. There are four states because 22 = 4. It is therefore a more bandwidth-efficient type of modulation than BPSK, potentially twice as efficient.

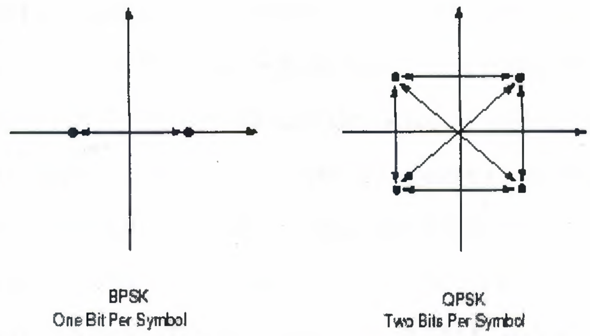


Figure 1.3.2.1: Phase Shift Keying

1.3.3 Frequency Shift Keying

Frequency modulation and phase modulation are closely related. A static frequency shift of +1 Hz means that the phase is constantly advancing at the rate of 360 degrees per second (2˜ rad/sec), relative to the phase of the unshifted signal. FSK (Frequency Shift Keying) is used in many applications including cordless and paging systems. Some of the cordless systems include DECT (Digital Enhanced Cordless Telephone) and CT2 (Cordless Telephone 2). In FSK, the frequency of the carrier is changed as a function of the modulating signal (data) being transmitted. Amplitude remains unchanged. In binary FSK (BFSK or 2FSK), a "1" is represented by one frequency and a "0" is represented by another frequency.

1.3.4 Minimum Shift Keying

Since a frequency shift produces an advancing or retarding phase, frequency shifts can be detected by sampling phase at each symbol period. Phase shifts of (2N + 1)˜/2 radians are easily detected with an *I/Q* demodulator. At even numbered symbols, the polarity of the *I* channel conveys the transmitted data; while at odd numbered symbols the polarity of the *Q* channel conveys the data. This orthogonal between *I* and *Q* simplifies detection algorithms and hence reduces power consumption in a mobile receiver. The minimum frequency shift which yields orthogonal of *I* and *Q* is that which results in a phase shift of ±˜/2 radians per symbol (90 degrees per symbol). FSK with this deviation is called MSK (Minimum Shift Keying). The deviation must be accurate in order to generate repeatable 90-degree phase shifts. MSK is used in the GSM (Global System for Mobile Communications) cellular standard. A phase shift of +90 degrees represents a data bit equal to "1," while –90 degrees represents a "0." The peak-to peak frequency shift of an MSK signal is equal to one -half of the bit rate. FSK and MSK produce constant envelope carrier signals, which have no amplitude variations. This is a desirable characteristic for improving the power efficiency of transmitters. Amplitude variations can exercise nonlinear ties in an amplifier's amplitude-transfer function, generating spectral regr owth, a component of adjacent channel power. Therefore, more efficient amplifiers (which tend to be less linear) can be used with constant-envelope signals, reducing power consumption.

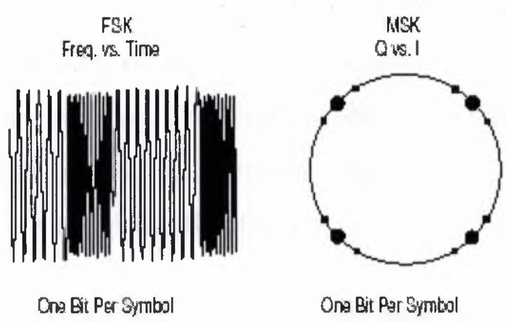


Figure 1.3.4.1: Frequency Shift Keying.

MSK has a narrower spectrum than wider deviation forms of FSK. The width of the spectrum is also influenced by the waveforms causing the frequency\ shift. If those waveforms have fast transitions or a high slew rate, then the spectrum of the transm itter will be broad. In practice, the waveforms are filtered with a Gaussian filter, resulting in a narrow spectrum. In addition, the Gaussian filter has no time -domain overshoot, which would broaden the spectrum by increasing the peak deviation. MSK with a Gaussian filter is termed GMSK (Gaussian MSK).

1.3.5 Quadrature Amplitude Modulation

Another member of the digital modulation family is Quadrature Amplitude Modulation (QAM). QAM is used in applications including microwave digital radio, DVB- C (Digital Video Broadcasting—Cable), and modems. In 16-state Quadrature Amplitude Modulation (16QAM), there are four I values and four Q values. This results in a total of 16 possible states for the signal. It can transition from any state to any other state at every symbol time. Since 16 = 24, four bits per symbol can be sent. This consists of two bits for *two bits and I* for Q. The symbol rate is one fourth of the bit rate. So this modulation format produces a more spectrally efficient transmission. It is more efficient than BPSK, QPSK, or 8PSK. Note that QPSK is the same as 4QAM. Another variation is 32QAM. In this case there are six I values and six Q values resulting in a total of 36 possible states (6x6=36). This is too many states for a power of two (the closest power of two is 32).

So the four corner symbol states, which take the most power to transmit, are omitted. This reduces the amount of peak power the transmitter has to generate. Since 2 5 = 32, there are five bits per symbol and the symbol rate is one fifth of the bit rate. The current practical limits are approximately 256QAM, though work is underway to extend the limits to 512 or 1024 QAM. A 256QAM system uses 16 I-values and 16 Q-values, giving 256 possible states. Since 28 = 256, each symbol can represent eight bits. A 256QAM signal that can send eight bits per symbol is very spectrally efficient.

However, the symbols are very close together and are thus more subject to errors due to noise and distortion. Such a signal may have to be transmitted with extra power (to effectively spread the symbols out more) and this reduces power efficiency as compared to simpler schemes.

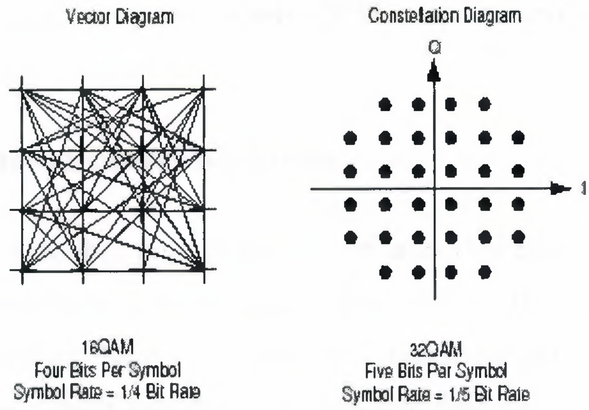


Figure 1.3.5.1: Quadrature Amplitude Modulation.

Compare the bandwidth efficiency when using 256QAM versus BPSK modulation in the radio example in section 1.3.1.1 (which uses an eight-bit sampler sampling at 10 kHz for voice). BPSK uses 80 KSymbols-per- second sending 1 bit per symbol. A system using 256QAM sends eight bits per symbol so the symbol rate would be 10 Ksymbols per second. A 256QAM system enables the same amount of information to be sent as BPSK using only one eighth of the bandwidth. It is eight times more bandwidth efficient. However, there is a tradeoff. The radio becomes more complex an d is more susceptible to errors caused by noise and distortion. Error rates of higher-order QAM systems such as this degrade more rapidly than QPSK as noise or interference is introduced. A measure of this degradation would be a higher Bit Error Rate (BER).

In any digital modulation system, if the input signal is distorted or severely attenuated the receiver will eventually lose symbol lock completely. If the receiver can no longer recover the symbol clock, it cannot demodulate the signal or recover any information. With less degradation, the symbol clock can be recovered, but it is noisy, and the symbol

Locations themselves are noisy. In some cases, a symbol will fall far enough away from its intended position that it will cross over to an adjacent position . The *I* and *Q* level detectors used in the demodulator would misinterpret such a symbol as being in the wrong location, causing bit errors. QPSK is not as efficient, but the states are much farther apart and the system can tolerate a lot more noise before suffering symbol errors. QPSK has no intermediate states between the four corner-symbol locations, so there is less opportunity for the demodulator to misinterpret symbols. QPSK requires less transmitter power than QAM to achieve the same bit error rate.

1.3.6 Theoretical Bandwidth Efficiency Limits

Bandwidth efficiency describes how efficiently the allocated bandwidth is utilized or the ability of a modulation scheme to accommodate data, within a limited bandwidth. The table below shows the theoretical band width efficiency limits for the main modulation types. Note that these figures cannot actually be achieved in practical radios since they require perfect modulators, demodulators, filter, and transmission paths. If the radio had a perfect (rectangular in the frequency domain) filter, then the occupied bandwidth could be made equal to the symbol rate. Techniques for maximizing spectral efficiency include the following:

1. Relate the data rate to the frequency shift(as in GSM).
2. Use premodulation filtering to reduce the occupied bandwidth. Raised cosine filters, as used in NADC, PDC, and PHS, give the best spectral efficiency.
3. Restrict the types of transitions.

Modulation format	Theoretical bandwidth efficiency limits
MSK	1 bit/second/Hz
BPSK	1 bit/second/Hz
QPSK	2 bits/second/Hz
8PSK	3 bits/second/Hz
16 QAM	4 bits/second/Hz
32 QAM	5 bits/second/Hz
64 QAM	6 bits/second/Hz
256 QAM	8 bits/second/Hz

Table 1.3.6.1: Theoretical Bandwidth Efficiency Limits for the Main Modulation Type.

Effects of going through the origin

Take, for example, a QPSK signal where the normalized value changes from 1, 1 to –1, 1. When changing simultaneously from I and Q values of +1 to I and Q values of –1, the signal trajectory goes through the origin (the I/Q value of 0 ,0). The origin represents 0 carrier magnitude. A value of 0 magnitude indicates that the carrier amplitude is 0 for

A moment. Not all transitions in QPSK result in a trajectory that goes through the origin. If I changes value but Q does not (or vice-versa) the carrier amplitude changes a little, but it does not go through zero. Therefore some symbol transitions will result in a small amplitude variation, while others will result in a very large amplitude variation. The clock recovery circuit in the receiver must deal with this amplitude variation uncertainty if it uses amplitude variations to align the receiver clock with the transmitter clock. Spectral regrowth does not automatically result from these trajectories that pass through or near the origin. If the amplifier and associated circuits are perfectly linear, the spectrum (spectral occupancy or occupied bandwidth) will be unchanged. The problem lies in nonlinearities in the circuits.

A signal which changes amplitude over a very large range will exercise these nonlinearities to the fullest extent. These nonlinearities will cause distortion products. In continuously modulated systems they will cause "spectral regrowth" or wider modulation sidebands (a phenomenon related to intermodulation distortion). Another term which is sometimes used in this context is "spectral splatter." However this is a term that is more correctly used in association with the increase in the bandwidth of a signal caused by pulsing on and off.

1.3.7 Spectral Efficiency Examples in Practical Radios

The following examples indicate spectral efficiencies that are achieved in some practical radio systems.

The TDMA version of the North American Digital Cellular (NADC) system, achieves a 48 Kbits-per second data rate over a 30 kHz bandwidth or 1.6 bits per second per Hz. It is a ˜/4

QPSK based system and transmits two bits per symbol. The theoretical efficiency would be two bits per second per Hz and in practice it is 1.6 bits per second per Hz. Another example is a microwave digital radio using 16QAM. This kind of signal is more susceptible to noise and distortion than something simpler such as QPSK. This type of signal is usually sent over a direct line-of-sight microwave link or over a wire where there is very little noise and interference. In this microwave-digital-radio example the bit rate is 140 Mbits per second over a very wide bandwidth of 52.5 MHz. The spectral efficiency is 2.7 bits per second per Hz. To implement this, it takes a very clear line-of-sight transmission path and a precise and optimized high-power transceiver.

Digital modulation types—variations

The modulation types outlined in sections 1.3.2 to 1.3.4 form the building blocks for many systems. There are three main variations on these basic building blocks that are used in communications systems: *I/Q* offset modulation, differential modulation, and constant envelope modulation.

I/Q Offset Modulation

The first variation is offset modulation. One example of this is Offset QPSK (OQPSK). This is used in the cellular CDMA (Code Division Multiple Access) system for the reverse (mobile to base) link. In QPSK, the *I* and *Q* bit streams are switched at the same time. The symbol clocks, or the *I* and *Q* digital signal clocks, are synchronized. In Offset QPSK (OQPSK), the *I* and *Q* bit streams are offset in their relative alignment by one bit period (one half of a symbol period).

Since the transitions of *I* and *Q* are offset, at any given time only one of the two bit streams can change values. This creates a dramatically different constellation, even though there are still just two *I/Q* values. This has power efficiency advantages. In OQPSK the signal trajectories are modified by the symbol clock offset so that the carrier amplitude does not go through or near zero (the center of the constellation). The spectral efficiency is the same with two *I* states and two *Q* states. The reduced amplitude variations (perhaps 3 dB for OQPSK, versus 30 to 40 dB for QPSK) allow a more power- efficient, less linear RF power amplifier to be used.

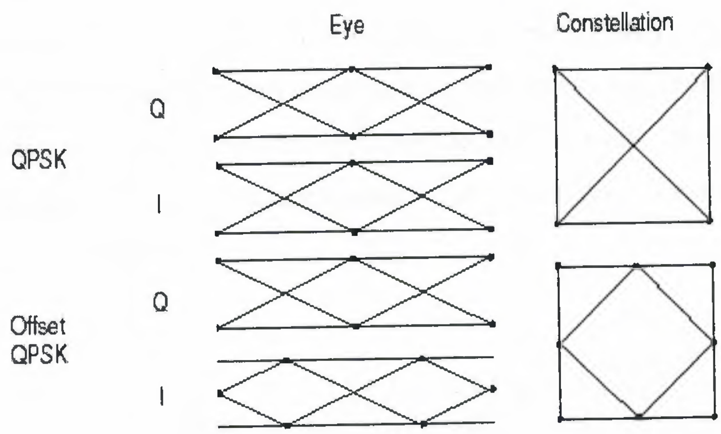


Figure 1.3.8.1.1: I _Q offset Modulation.

1.4 Different Ways of Looking at a Digitally Modulated Signal Time and Frequency Domain View

There are a number of different ways to view a signal. This simplified example is an RF pager signal at a center frequency of 930.004 MHz. This pager uses two-level FSK and the carrier shifts back and forth between two frequencies that are 8 kHz apart (930.000 MHz and 930.008 MHz). This frequency spacing is small in proportion to the center frequency of 930.004 MHz. This is shown in Figure 1.4.1(a). The difference in period between a signal at 930 MHz and one at 930 MHz plus 8 kHz is very small. Even with a high performance oscilloscope, using the latest in high-speed digital techniques, the change in period cannot be observed or measured. In a pager receiver the signals are first down converted to an IF or baseband frequency. In this example, the 930.004 MHz FSK-modulated signal is mixed with another signal at 930.002 MHz. The FSK modulation causes the transmitted signal to switch between 930.000 MHz and 930.008 MHz. The result is a baseband signal that alternates between two frequencies, –2 kHz and +6 kHz. The demodulated signal shifts between –2 kHz and +6 kHz. The difference can be easily detected. This is sometimes referred to as "zoom" time or IF time. To be more specific, it is a band converted signal at IF or baseband. IF time is important as it is how the signal looks in the IF portion of a receiver. This is how the IF of the radio detects the different bits that are present. Most pagers use a two-level, Frequency-Shift-Keying (FSK) scheme. FSK is used in this instance because it is less affected by multipath propagation, attenuation and interference, common in urban environments. It is possible to demodulate it even deep inside modern steel/concrete buildings, where attenuation, noise and interference would otherwise make reliable demodulation difficult.

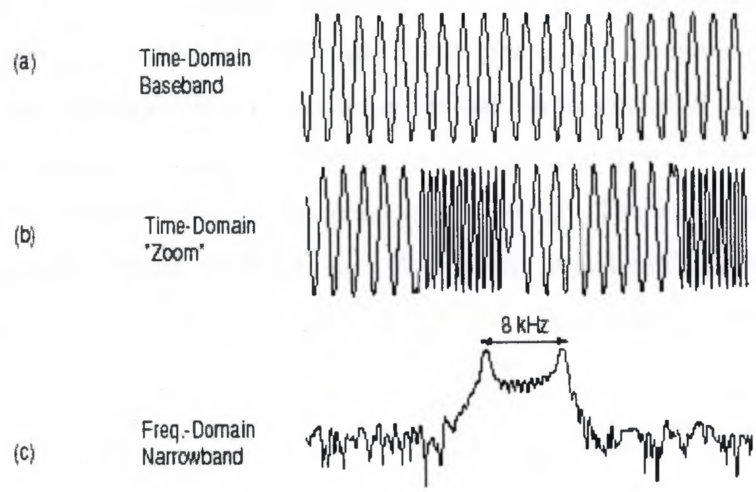


Figure 1.4.1: Time and Frequency Domain View.

1.4.1 Power and Frequency View

There are many different ways of looking at a digitally modulated signal. To examine how transmitters turn on and off, a power-versus-time measurement is very useful for examining the power level changes involved in pulsed or bursted carriers. For example, very fast power changes will result in frequency spreading or spectral regrowth. This is also known as frequency "splatter." Very slow power changes waste valuable transmit time, as the transmitter cannot send data when it is not fully on. Turning on too slowly can also cause high bit error rates at the beginning of the burst. In addition, peak and average power levels must be well understood, since asking for excessive power from an amplifier can lead to compression or clipping. These phenomena distort the modulated signal and usually lead to spectral regrowth as well.

1.4.2 Constellation Diagrams

As discussed, the rectangular *I/Q* diagram is a polar diagram of magnitude and phase. A two-dimensional diagram of the carrier magnitude and phase (a standard polar plot) can be represented differently by superimposing rectangular axes on the same data and interpreting the carrier in terms of in-phase (*I*) and quadrature-phase (*Q*) components. It would be possible to perform AM and PM on a carrier at the same time and send data this way; it is easier for circuit design and signal processing to generate and detect a rectangular, linear set of values (one set for *I* and an independent set for *Q*). The example shown is a ˜/4 Differential Quadrature Phase Shift Keying (˜/4 DQPSK) signal as described in the North American Digital Cellular (NADC) TDMA standard. This example is a 157 symbol DQPSK burst.

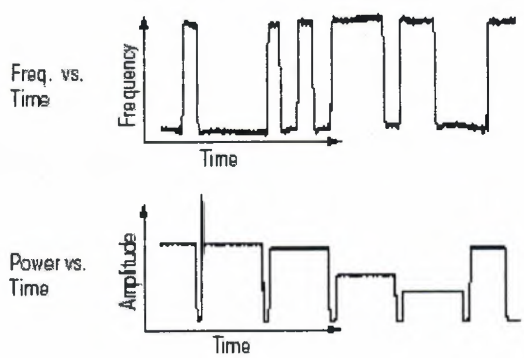


Figure 1.4.2.1: Power and Frequency View.

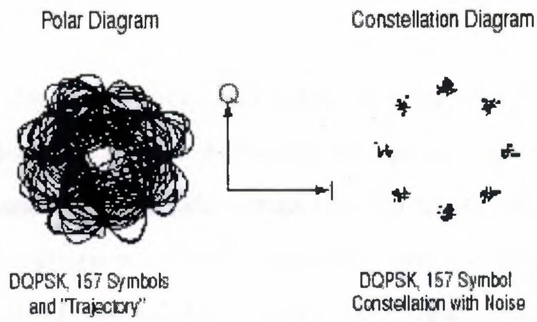


Figure 1.4.2.2: Constellation Diagram.

The polar diagram shows several symbols at a time. That is, it shows the instantaneous value of the carrier at any point on the continuous line between and including symbol times, represented as *I/Q* or magnitude/phase values. The constellation diagram shows a repetitive "snapshot" of that same burst, with values shown only at the decision points. The constellation diagram displays phase errors, as well as amplitude errors, at the decision points. The transitions between the decision points affects transmitted bandwidth. This display shows the path the carrier is taking but does not explicitly show errors at the decision points. Constellation diagrams provide insight into varying power levels, the effects of filtering, and phenomena such as Inter-Symbol Interference. The relationship between constellation points and bits per symbol is,

$$M{=}2n\text{.........}\ (1.4.2.1)$$

Where

M = number of constellation points.

n = bits/symbol

or n = log2 (M)

This holds when transitions are allowed from any constellation point to any other.

1.4.3 Eye Diagrams

Another way to view a digitally modulated signal is with an eye diagram. Separate eye diagrams can be generated, one for the *I*-channel data and another for the *Q*-channel data. Eye diagrams display *I* and *Q* magnitude versus time in an infinite persistence mode, with retraces. The *I* and *Q* transitions are shown separately and an "eye" (or eyes) is formed at the symbol decision times. QPSK has four distinct *I/Q* states, one in each quadrant. There are only two levels for *I* and two levels for *Q*. This forms a single eye for each *I* and *Q*. Other schemes use more levels and create more nodes in time through which the traces pass. The lower example is a 16QAM signal, whic h has four levels forming three distinct "eyes." The eye is open at each symbol. A "good" signal has wide-open eyes with compact crossover points.

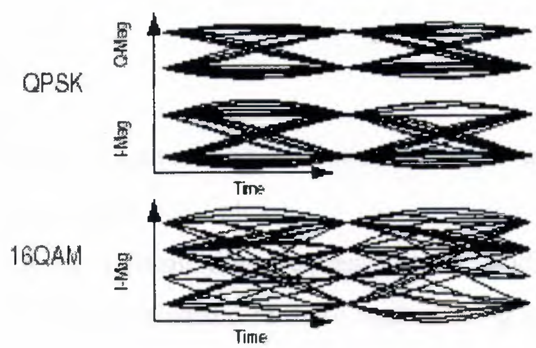


Figure 1.4.3.1: I and Q Eye Diagrams.

1.4.4 Trellis Diagrams

Figure 1.4.4.1 is called a "trellis" diagram, because it resembles a garden trellis. The trellis diagram shows time on the X-axis and phase on the Y- axis. This allows the examination of the phase transitions with different symbols. In this case it is for a GSM system. If a long series of binary ones were sent, the result would be a series of positive phase transitions of, in the example of GSM, 90 degrees per symbol. If a long series of binary zeros were sent, there would be a constant declining phase of 90 degrees per symbol. Typically there would be intermediate transmissions with random data. When troubleshooting, trellis diagrams are useful in isolating missing transitions, missing codes, or a blind spot in the *I/Q* modulator or mapping algorithm.

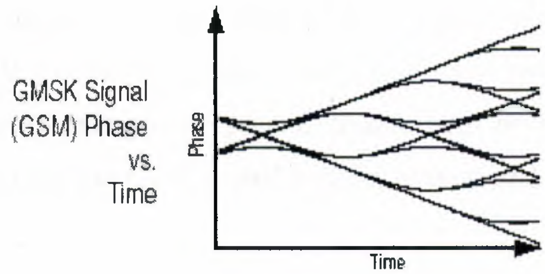


Figure 1.4.4.1: Trellis Diagram

1.5 Sharing the Channel

The RF spectrum is a finite resource and is shared between users using multiplexing (sometimes called channelization). Multiplexing is used to separate different users of the spectrum. This section covers multiplexing frequency, time, code, and geo graphy. Most communications systems use a combination of these multiplexing methods.

1.5.1 Multiplexing—Frequency

Frequency Division Multiple Access (FDMA) splits the available frequency band into smaller fixed frequency channels. Each transmitter or receiver uses a separate frequency. This technique has been used since around 1900 and is still in use today. Transmitters are narrowband or frequency-limited. A narrowband transmitter is used along with a receiver that has a narrowband filter so that it can demodulate the desired signal and reject unwanted signals, such as interfering signals from adjacent radios.

1.5.2 Multiplexing—Time

Time-division multiplexing involves separating the transmitters in time so that they can share the same frequency. The simplest type is Time Division Duplex (TDD). This multiplexes the transmitter and receiver on the same frequency. TDD is used, for example, in a simple two-way radio where a button is pressed to talk and released to listen. This kind of time division duplex, however, is very slow. Modern digital radios like CT2 and DECT use Time Division Duplex but they multiplex hundreds of times per second. TDMA (Time Division Multiple Access) multiplexes several transmitters or receivers on the same frequency. TDMA is used in the GSM digital cellular system and also in the US NADC-TDMA system.

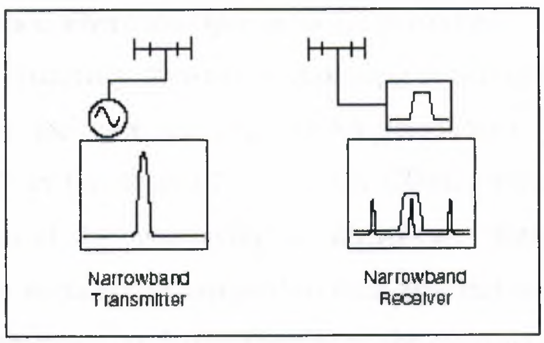


Figure 1.5.2.1: Multiplexing Frequency.

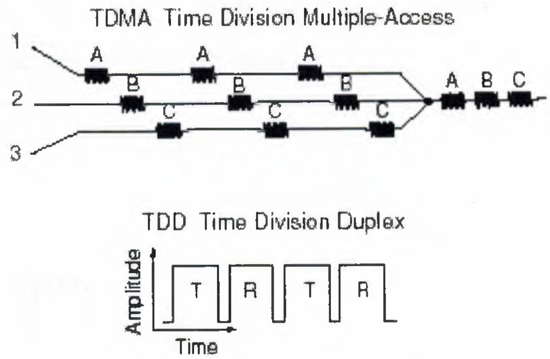


Figure 1.5.2.2: Multiplexing Time.

1.5.3 Multiplexing—Code

CDMA is an access method where multiple users are permitted to transmit simultaneously on the same frequency. Frequency division multiplexing is still performed but the channel is 1.23 MHz wide. In the case of US CDMA telephones, an additional type of channelization is added, in the form of coding. In CDMA systems, users timeshare a higher-rate digital channel by overlaying a higher-rate digital sequence on their transmission. A different sequence is assigned to each terminal so that the signals can be discerned from one another by correlating them with the overlaid sequence. This is based on codes that are shared between the base and mobile stations. Because of the choice of coding used, there is a limit of 64 code channels on the forward link. The reverse link has no practical limit to the number of codes available.

1.5.4 Multiplexing—Geography

Another kind of multiplexing is geographical or cellular. If two transmitter/receiver pairs are far enough apart, they can operate on the same frequency and not interfere with each other. There are only a few kinds of systems that do not use some sort of geographic multiplexing. Clear-channel international broadcast stations, amateur stations, and some military low frequency radios are about the only systems that have no geographic boundaries and they broadcast around the world.

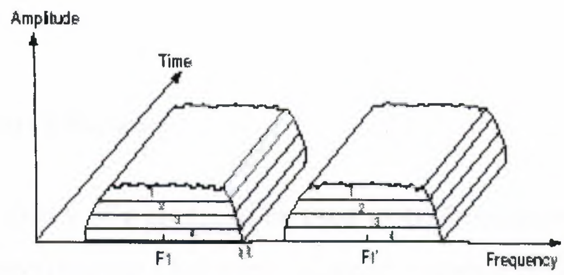


Figure 1.5.4.1: Multiplexing code.

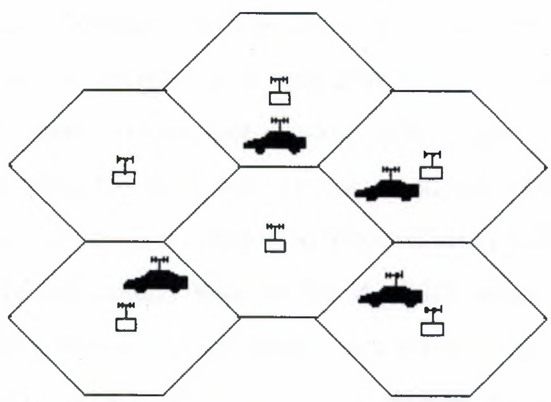


Figure 1.5.4.2: Multiplexing_ Geography.

1.5.5 Combining Multiplexing Modes

In most of these common communications systems, different forms of multiplexing are generally combined. For example, GSM uses FDMA, TDMA, FDD, and geographic. DECT uses FDMA, TDD, and geographic multiplexing.

1.5.6 Penetration Versus Efficiency

Penetration means the ability of a signal to be used in environments where there is a lot of attenuation, noise, or interference. One very common example is the use of pagers versus cellular phones. In many cases, pagers can receive signals even if the user is inside a metal building or a steel-reinforced concrete structure like a modern skyscraper. Most pagers use a two-level FSK signal where the frequency dev iation is large and the modulation rate (symbol rate) is quite slow. This makes it easy for the receiver to detect and demodulate the signal since the frequency difference is large (the symbol locations are widely separated) and these different frequencies persist for a long time (a slow symbol rate). However, the factors causing good pager signal penetration also cause inefficient information transmission. There are typically only two symbol locations. They are widely separated (approximately 8 kHz), and a small number of symbols (500 to 1200) are sent each second. Compare this with a cellular system such as GSM which sends 270,833 symbols each second. This is not a big problem for the pager since all it needs to receive is its unique address and perhaps a short ASCII text message. A cellular phone signal, however, must transmit live duplex voice. This requires a much higher bit rate and a much more efficient modulation technique. Cellular phones use more complex modulation formats (such as ~/4 DQPSK and 0.3 GMSK) and faster symbol rates. Unfortunately, this greatly reduces penetration and one way to compensate is to use more power. More power brings in a host of other problems, as described previously.

1.6 Digital Transmitters and Receive rs Work

1.6.1 A digital communications transmitter

Figure 1.6.1.1 is a simplified block diagram of a digital communications transmitter. It begins and ends with an analog signal. The first step is to convert a continuous analog signal to a discrete digital bit stream. This is called digitization. The next step is to add voice coding for data compression. Then some channel coding is added. Channel coding encodes the data in such a way as to minimize the effects of noise and interference in the communications channel. Channel coding add extra bits to the input data stream and removes redundant ones. Those extra bits are used for error correction or sometimes to send training sequences for identification or equalization. This can make synchronization (or finding the symbol clock) easier for the receiver. The symbol clock represents the frequency and exact timing of the transmission of the individual symbols. At the symbol clock transitions,

The transmitted carrier is at the correct *I/Q* (or magnitude/phase) value to represent a specific symbol (a specific point in the constellation). Then the values (*I/Q* or magnitude/phase) of the transmitted carrier are changed to represent another symbol. The interval between these two times is the symbol clock period. The recip rocal of this is the symbol clock frequency. The symbol clock phase is correct when the symbol clock is aligned with the optimum instant(s) to detect the symbols. The next step in the transmitter is filtering. Filtering is essential for good bandwidth efficiency. Without filtering, signals would have very fast transitions between states and therefore very wide frequency spectra—much wider than is needed for the purpose of sending information. A single filter is shown for simplicity, but in reality there are two filters; one each for the *I* and *Q* channels. This creates a compact and spectrally efficient signal that can be placed on a carrier. The output from the channel coder is then fed into the modulator. Since there are independent *I* and *Q* components in the radio, half of the information can be sent on *I* and the other half on *Q*. This is one reason digital radios work well with this type of digital signal. The *I* and *Q* components are separate. The rest of the transmitter looks similar to a typical RF transmitter or microwave transmitter/ receiver pair. The signal is converted up to a higher intermediate frequency (IF), and then further up converted to a higher radio frequency (RF). Any undesirable signals that were produced by the up conversion are then filtered out.

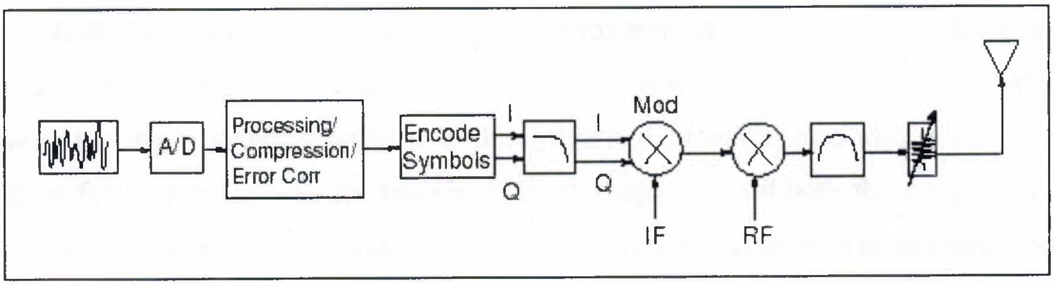


Figure 1.6.1.1: Digital Transmitter.

1.6.2 A digital Communications Receiver

The receiver is similar to the transmitter but in reverse. It is more complex to design. The incoming (RF) signal is first down converted to (IF) and demodulated. The ability to demodulate the signal is hampered by factors including atmospheric noise, competing signals, and multipath or fading. Generally, demodulation involves the following stages:

1. Carrier frequency recovery (carrier lock).

2. Symbol clock recovery (symbol lock).

3. Signal decomposition to *Q* components and I.

4. Determining *I* and *Q* values for each symbol ("slicing").

5. Decoding and de-interleaving.

6. Expansion to original bit stream.

7. Digital-to-analog conversion, if required in more and more systems.

However, the signal starts out digital and stays digital. It is never analog in the sense of a continuous analog signal like audio. The main difference between the transmitter and receiver is the issue of carrier and clock (or symbol) recovery. Both the symbol-clock frequency and phase (or timing) must be correct in the receiver in order to demodulate the bits successfully and recover the transmitted information. A symbol clock could be at the right frequency but at the wrong phase. If the symbol clo ck were aligned with the transitions between symbols rather than the symbols themselves, demodulation would be unsuccessful. Symbol clocks are usually fixed in frequency and both the transmitter and receiver accurately know this frequency. The difficulty is to get them both aligned in phase and timing. There are a variety of techniques and most systems employ two or more. If the signal amplitude varies during modulation, a receiver can measure the variations. The transmitter can send a specific synchronization signal or a predetermined bit sequence such as 10101010101010 to "train" the receiver's clock. In systems with a pulsed carrier, the symbol clock can be aligned with the power turn -on of the carrier. In the transmitter, it is known where the RF carrier and digital data clock are because they are being generated inside the transmitter itself. In the receiver there is not this luxury. The receiver can approximate where the carrier is but has no phase or timing symbol clock information. A difficult task in receiver design is to create carrier and symbol clock recovery algorithms. That task can be made easier by the channel coding performed in the transmitter.

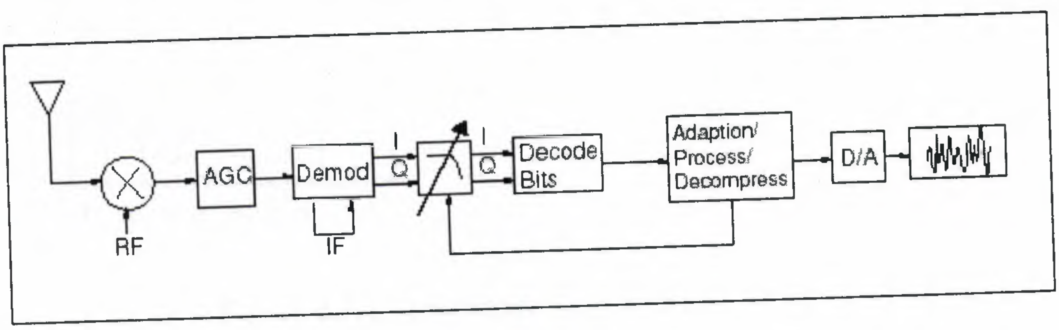


Figure 1.6.2.1: Digital Receiver

CHAPTER TWO

DIGITAL MODULAION TECHNIQUES

2.1 Digital Modulation Techniques

There are three ways in which the bandwidth of the channel carrier may be altered simply. It is worth emphasising that these methods are chosen because they are practically simple, not because they are theoretically desirable. These are the altering of the amplitude, frequency and phase of the carrier sine wave. These techniques give rise to amplitude-shift-keying (ASK), frequency-shift-keying (FSK) and phase-shift-keying (PSK), respectively.

ASK describes the technique the carrier wave is multiplied by the digital signal $f(t)$ mathematically, the modulated carrier signal $s(t)$ is:

$$s(t) = f(t)\sin(2\pi f_c + \phi) \ldots\ldots\ldots (2.1.1)$$

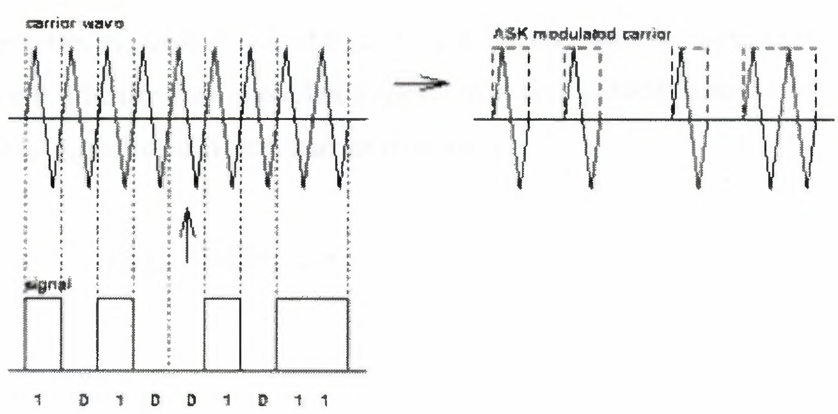


Figure 2.1.1: Amplitude Shift Keying.

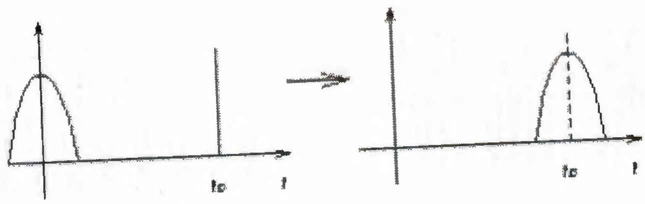


Figure 2.1.2: Amplitude Shift Keying -- Frequency Domain.

It is a special case of amplitude modulation (AM). Amplitude modulation has the property of translating the spectrum of the modulation $f(t)$ to the carrier frequency. The bandwidth of the signal remains unchanged.

The fact that AM simply shifts the signal spectrum is often used to convert the carrier frequency to a more suitable value without altering the modulation. This process is known variously as mixing, up-conversion or down-conversion. Some form of conversion will always be present when the channel carrier occupies a frequency range outside the modulation frequency range.

FSK describes the modulation of a carrier (or two carriers) by using a different frequency for a 1 or 0. The resultant modulated signal may be regarded as the sum of two amplitude modulated signals of different carrier frequency:

$$s(t) = f_1(t)\sin(2\pi f_{c1} + \phi) + f_2(t)\sin(2\pi f_{c2} + \phi) \quad \text{.........(2.1.2)}$$

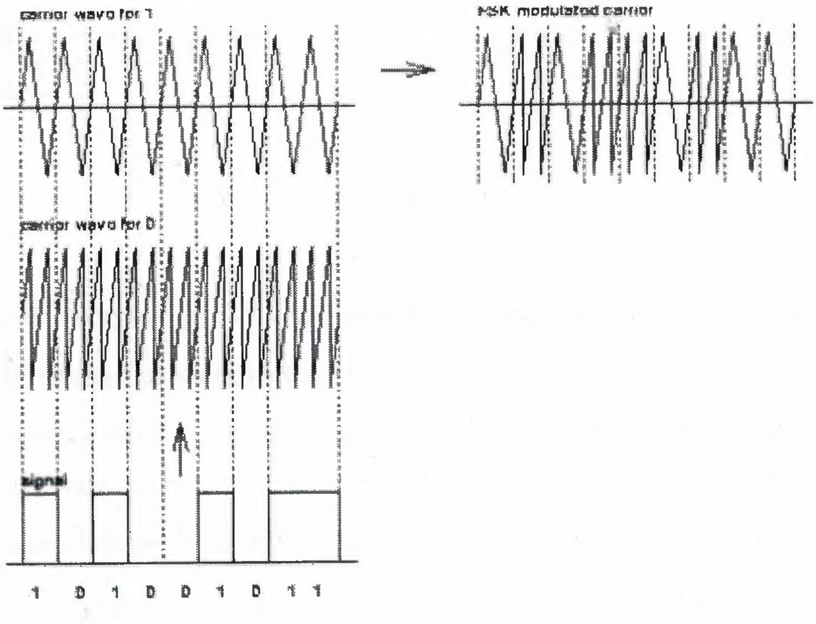


Figure 2.1.3: Frequency Shift Keying.

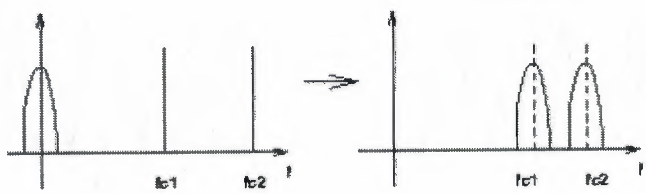


Figure 2.1.4: Frequency Shift Keying -- Frequency Domain.

FSK is classified as wide-band if the separation between the two carrier frequencies is larger than the bandwidth of the spectrums of $f1(t)$ and $f2(t)$. In this case the spectrum of the modulated signal appears as two separate ASK signals. Narrow-band FSK is the term used to describe an FSK signal whose carrier frequencies are separated by less than the width of the spectrum than ASK for the same modulation.

PSK describes the modulation technique that alters the phase of the carrier. Mathematically:

$$s(t) = \sin(2\pi f_c + \phi(t)) \text{ ....(2.1.3)}$$

Binary phase-shift-keying (BPSK) has only two phases, 0 and π. It is therefore a type of ASK with $f(t)$ taking the values -1 or 1, and its bandwidth is the same as that of ASK. Phase-shift-keying offers a simple way of increasing the number of levels in the transmission without increasing the bandwidth by introducing smaller ph ase shifts. Quadrature phase-shift-keying (QPSK) has four phases $0, \pi/2, \pi, 3\pi/2$, M-ary PSK has M phases, $2\pi m/M$ where $m = 1, \ldots M-1$ for a given bit-rate, and QPSK requires half the bandwidth of PSK and is widely used for this reason.

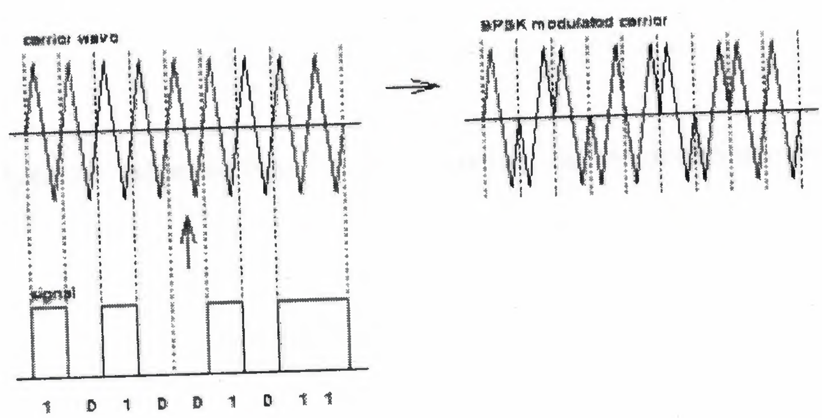


Figure 2.1.5: Binary Phase Shift Keying.

The number of times the signal parameter (amplitude, frequency, and phase) is changed per second is called the signaling rate. It is measured in baud. 1 baud = 1 change per second. With binary modulations such as ASK, FSK and BPSK, the signaling rate equals the bit-rate. With QPSK and M-ary PSK, the bit-rate may exceed the baud rate.

2.2 Quadrature Amplitude Modulation

Looking back at the QPSK signal space, we can see that the points are equally spaced on the circumference of a circle and if we combined bits into groups of four, there would be16 possible values and 16 points on the circumference in the signal space of 16-ary PSK or 16-PSK. Performance improves when these points are separated as widely as possible. Quadrature amplitude modulation (QAM) is one approach toward a different point distribution within the signal space. We chose 16-QAM as an example of this form of modulation. That is, we consider combining bits in groups of four, yielding 16 possible values. These values change every 4Tb, so we can expect a bandwidth that is one-fourth that of BPSK. If we use 16 ary- PSK, the points in signal space would be equally spaced on the circumference of a circle, and the angular spacing between adjacent points would be 22.5 degrees. In this case, both the amplitude and the phase vary, so the points no longer lie on the circumference of a single circle. The signal space diagram consists of 16 points in a uniform square array. The Individual signals are of the form:

$$s_i(t) = A_i \cos(2\pi f_c t + \theta_i) \text{..........} (2.2.1)$$

The index I take the values of 0 to 15. The equation can be re -written to the following form:

$$s_i = RE\{A_i e^{j\theta} e^{2\pi f_c t} \text{ ....} (2.2.2)$$

The following is a mathematical implementation of a 16-QAM modulator scheme:

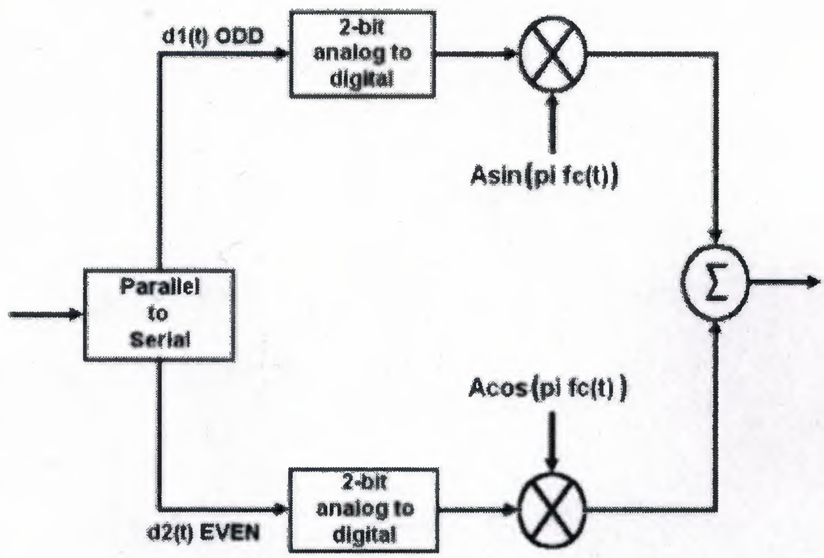


Figure 2.2.1: Mathematical Implementation of a 16_QAM Modulator Scheme.

We have implemented a 16-QAM modulator- demodulator Demonstration in Stimulant; we have used a Pseudo-Random signal as an input.

2.2.1: 16 QAM Demonstrations

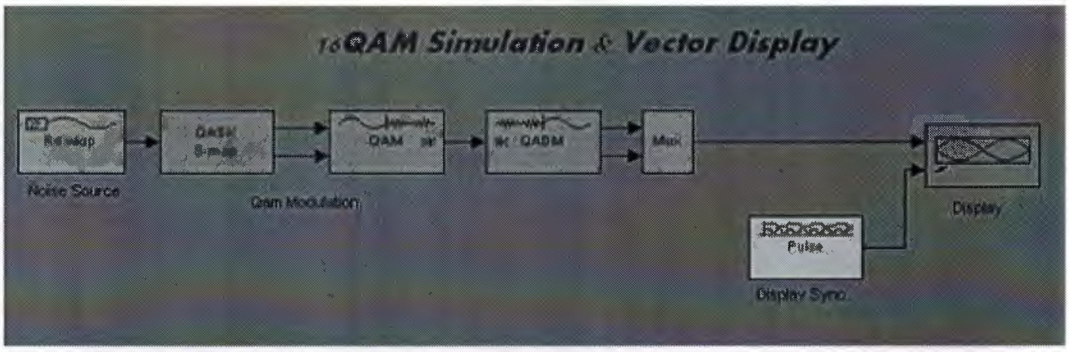


Figure 2.2.1.1: 16 QAM simulation and vector display.

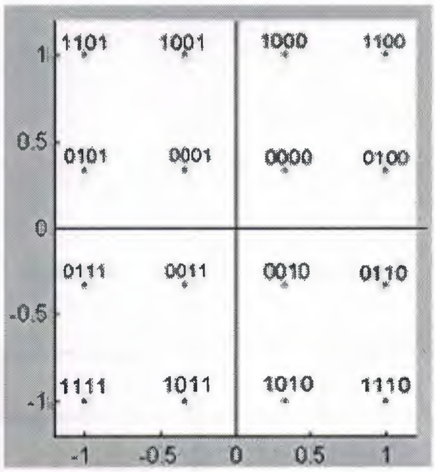
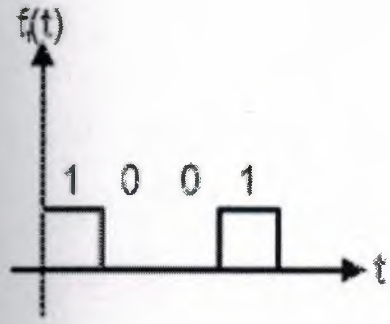
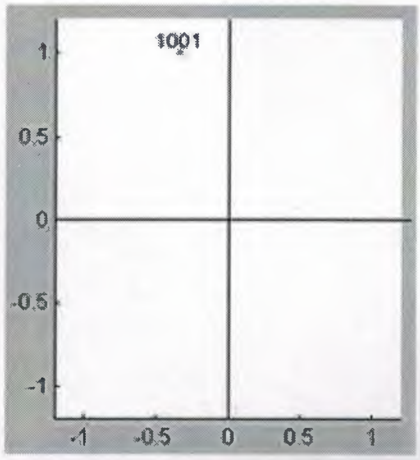


Figure 2.2.2: Visual Demonstration of 16 QAM Signal Space.

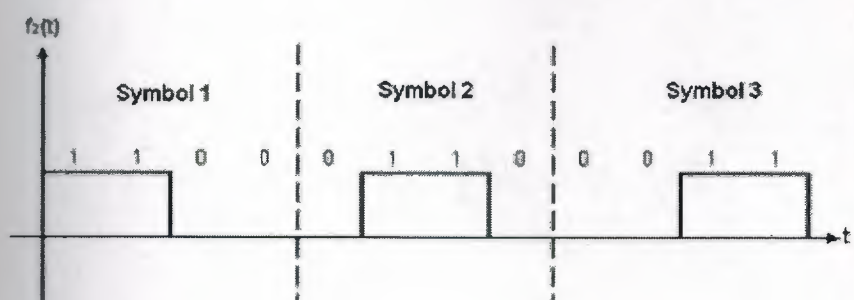


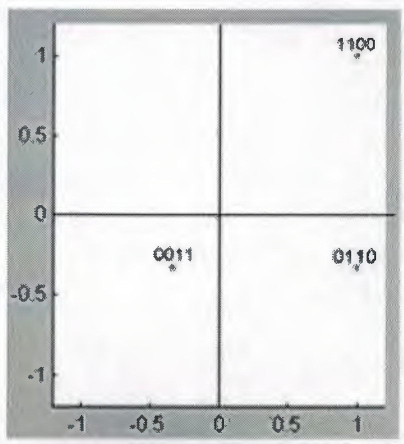


A "1001" Symbol in the time domain

The symbol's vector constellation

Figure 2.2.3: A"1001" Symbol in the Time Domain and Vector Constellation.





3 consecutive symbols in the time domain

Constellation of the 3 symbols

Figure 2.2.4: 3 consecutive symbols in the time domain.

2.3 QPSK Modulation

Since the early days of electronics, as advances in technology were tak ing place, the boundaries of both local and global communication began eroding, resulting in a world that is smaller and hence more easily accessible for the sharing of knowledge and information. The pioneering work by Bell and Marconi formed the cornerstone of the information age that exists today and paved the way for the future of telecommunications.

Traditionally, local communication was done over wires, as this presented a cost-effective way of ensuring a reliable transfer of information. For long-distance communications, transmission of information over radio waves was needed. Although this was convenient from a hardware standpoint, radio -waves transmission raised doubts over the corruption of the information and was often dependent on high -power transmitters to overcome weather conditions, large buildings, and interference from other sources of electromagnetic.

The various modulation techniques offered different solutions in terms of cost-effectiveness and quality of received signals but until recen tly were still largely analog. Frequency modulation and phase modulation presented certain immunity to noise, whereas amplitude modulation was simpler to demodulate. However, more recently with the advent of low-cost microcontrollers and the introduction of domestic mobile telephones and satellite communications, digital modulation has gained in popularity. With digital modulation techniques come all the advantages that traditional microprocessor circuits have over their analog counterparts. Any shortfalls in the communications link can be eradicated using software. Information can now be encrypted, error correction can ensure more confidence in received data, and the use of DSP can reduce the limited bandwidth allocated to each service.

As with traditional analog systems, digital modulation can use amplitude, frequency, or phase modulation with different advantages. As frequency and phase modulation techniques offer more immunity to noise, they are the preferred scheme for the majority of services in use today and will be discussed in detail below.

2.3.1 Digital Frequency Modulation

A simple variation from traditional analog frequency modulation (FM) can be implemented by applying a digital signal to the modulation input. Thus, the output takes the form of a sine wave at two distinct frequencies. To demodulate this waveform, it is a simple matter of passing the signal through two filters and translating the resultant back into logic levels. Traditionally, this form of modulation has been called frequency-shift keying (FSK).

2.3.2 Digital Phase Modulation

Spectrally, digital phase modulation, or phase-shift keying (PSK), is very similar to frequency modulation. It involves changing the phase of the transmitted waveform instead of the frequency, these finite phase changes representing digital data. In its simplest form, a phase-modulated waveform can be generated by using the digital data to switch between two signals of equal frequency but opposing phase. If the resultant waveform is multiplied by a sine wave of equal frequency, two components are generated: one cosine waveform of double the received frequency and one frequency-independent term whose amplitude is proportional to the cosine of the phase shift. Thus, filtering out the higher-frequency term yields the original modulating data prior to transmission. This is difficult to picture conceptually.

2.3.3 Quadra phase-Shift Modulation

Taking the above concept of PSK a stage further, it can be assumed that the number of phase shifts is not limited to only two states. The transmitted "carrier" can undergo any number of phase changes and, by multiplying the received signal by a sine wave of equal frequency, will demodulate the phase shifts into frequency-independent voltage levels.

This is indeed the case in quadraphase-shift keying (QPSK). With QPSK, the carrier undergoes four changes in phase (four symbols) and can thus represent 2 binary bits of data per symbol. Although this may seem insignificant initially, a modulation scheme has now been supposed that enables a carrier to transmit 2 bits of information instead of 1, thus effectively doubling the bandwidth of the carrier.

The proof of how phase modulation, and hence QPSK, is demodulated is shown below. The proof begins by defining Euler's relations, from which all the trigonometric identities can be derived. Euler's relations state the following:

$$\sin \omega t = \frac{e^{j\omega t} - e^{-j\omega t}}{2j} \quad \text{..........} \ (2.2.5)$$

$$\cos \omega t = \frac{e^{j\omega t} + e^{-j\omega t}}{2} \quad \text{..........} \ (2.2.6)$$

Now consider multiplying two sine waves together, thus

$$\sin^2 \omega t = \frac{e^{j\omega t} - e^{-jwt}}{2j} \times \frac{e^{j\omega t} - e^{-j\omega t}}{2j} = \frac{e^{2j\omega t} - 2e^{o} + e^{-2j\omega t}}{-4} \quad \text{..........} \ (2.2.7)$$

$$= \frac{1}{2} - \frac{1}{2}\cos 2\omega t \quad \text{..........} \ (2.2.8)$$

From Equation 2.2.8, it can be seen that multiplying two sine waves together (one sine being the incoming signal, the other being the local oscillator at the receiver mixer) results in an output frequency ($\frac{1}{2}\cos 2\omega t$) double that of the input (at half the amplitude) superimposed on a dc offset of half the input amplitude.

Similarly, multiplying $\sin \omega t$ by $\cos \omega t$ gives:

$$\sin \omega t \times \cos \omega t = \frac{e^{2j\omega t} - e^{-2j\omega t}}{4j}$$

$$= \sin 2\omega t \qquad \text{..........} (2.2.8)$$

Which gives an output frequency ($\sin 2\omega t$)double that of the input, with no dc offset.

It is now fair to make the assumption that multiplying $\sin \omega t$ by any phase-shifted sine wave ($\sin \omega t + \phi$)yields a "demodulated" waveform with an output frequency double that of the input frequency, whose dc offset varies according to the phase shift, ϕ.

To prove this,

$$\sin \omega t \times \sin(\omega t + \phi) = \frac{e^{j\omega t} - e^{-j\omega t}}{2j} \times \frac{e^{j(j\omega t + \phi)} - e^{-j(\omega t + \phi)}}{2j} \quad \text{...2.2.9}$$

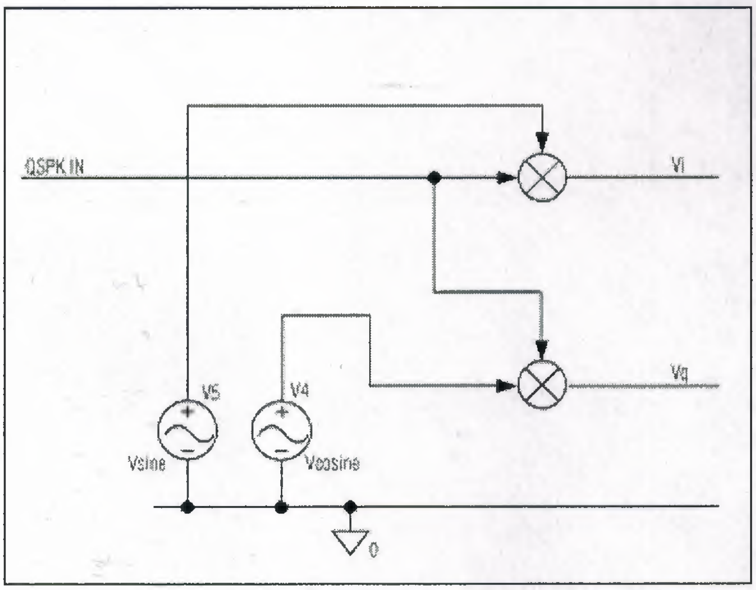
$$\sin \omega t \times \sin(\omega t + \phi) = \frac{\cos \phi}{2} - \frac{\cos(2\omega t + \phi}{2} \quad \text{..........} (2.2.10)$$

Thus, the above proves the supposition th at the phase shift on a carrier can be demodulated into a varying output voltage by multiplying the carrier with a sine -wave local oscillator and filtering out the high-frequency term. Unfortunately, the phase shift is limited to two quadrants; a phase shift of π/2 cannot be distinguished from a phase shift of -π/2. Therefore, to accurately decode phase shifts present in all four quadrants, the input signal needs to be multiplied by both sinusoidal and cosinusoidal waveforms, the high frequency filtered out, and the data reconstructed. The proof of this, expanding on the above mathematics, is shown below. Thus,

$$\cos \omega t \times \sin(\omega t + \phi) = \frac{e^{j\omega t} + e^{-j\omega t}}{2} \times \frac{e^{j(\omega t + \phi)} - e^{-j(\omega t + \phi)}}{2j} \quad \text{..........} (2.2.11)$$

$$\cos \omega t \times \sin(\omega t + \phi) = \frac{\sin(2\omega t + \phi)}{2} + \frac{\sin \phi}{2} \text{..........} (2.2.12)$$

A SPICE simulation verifies the above theory. Figure 2.3.3.1shows a block diagram of a simple demodulator circuit. The input voltage, QPSK IN, is a 1M Hz sine wave whose phase is shifted by 45°, 135°, 225°, and then 315° every 5μs.



Figures 2.3.3.1: Block Diagram of a Simple Demodulator Circuit.

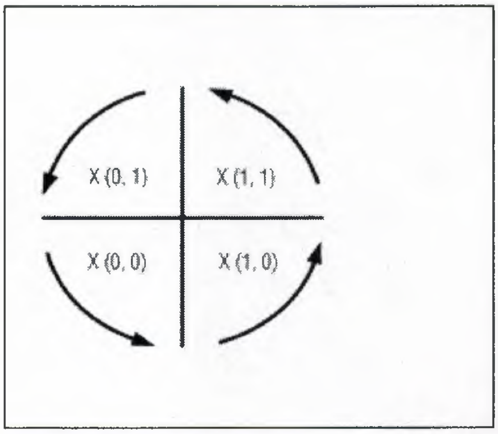


Figure 2.3.3.2: Pharos Diagram Showing the Phase Shift of QPSK IN and the Demodulated Data.

The above theory is perfectly acceptable, and it would appear that removing the data from the carrier is a simple process of low-pass filtering the output of the mixer and reconstructing the 4 voltages back into logic levels. In practice, getting a receiver loc al oscillator exactly synchronized with the incoming signal is not easy. If the local oscillator varies in phase with the incoming signal, the signals on the phasor diagram will undergo a phase rotation, its magnitude equal to the phase difference. Moreover, if the phase *and* frequency of the local oscillator are not fixed with respect to the incoming signal, there will be a continuing rotation on the phasor diagram.

Therefore, the output of the front-end demodulator is normally fed into an ADC and any rotation resulting from errors in the phase or frequency of the local oscillator are removed in DSP.

With the advances in monolithic silicon germanium (SiGe) technology, all of the above front-end circuitry can be integrated to reduce the problems outlined. A good example of how much of the front-end circuitry can be integrated is illustrated in the MAX2450, ultra - low-power quadrature modulator/demodulator IC. This is one of many devices from Maxim Integrated Products that incorporates the quadraphase shifter, the on -chip oscillator, and the mixer. Once the data has been demodulated, the output can be applied to a high - frequency dual-channel ADC (such as the MAX1002 or the MAX1003) before processing the signal in DSP.

As the MAX2450 is designed to be used at an IF of 35MHz to 80MHz, RF signals up to 2.5GHz can be down converted using the MAX2411A. This is a high -frequency up/down converter with a low-noise amplifier (LNA) local oscillator, and it has access to the output of the LNA for image-reject filtering.

Alternatively, an effective way of converting straight to base band is using a direct - conversion tuner IC. The MAX2102 is designed to take RF inputs from 2150MHz and convert directly down to base band I and Q signals, thus providing cost savings over multiple-stage devices.

The above devices are part of the rapidly expanding RF chipsets from Maxim Integrated Products. With five high-speed processes, more than 70 high-frequency standard products, and 52 ASICs in development, Maxim is committed to being a major player in the RF/wireless, fiber/cable, and instrumentation markets.

2.4. Quadriphase Shift keying (QPSK).

2.4.1 Coherent Quadrature-Modulation Techniques.

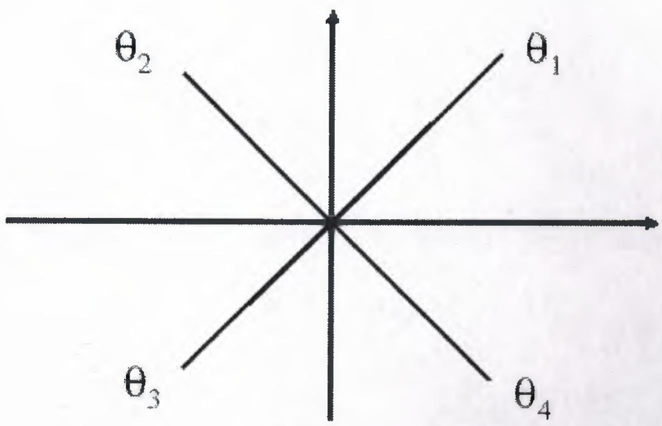


Figure 2.4.1.1:Quadrature_Modulation Diagram.

i	?*i*	bit in phase	cos(?*i*)	−sin(?*i*)	bit quad.
1	p/4	1	$1/\sqrt{2}$	$-1/\sqrt{2}$	0
2	3p/4	0	$-1/\sqrt{2}$	$-1/\sqrt{2}$	0
3	5p/4	0	$-1/\sqrt{2}$	$1/\sqrt{2}$	1
4	7p/4	1	$1/\sqrt{2}$	$1/\sqrt{2}$	1

Table 2.4.1.1: Figure the Corresponding Chases and the Messages.

symbol i	$?_i$	s_{i1}	s_{i2}	Gray code
1	p/4	$\sqrt{E/2}$	$-\sqrt{E/2}$	10
2	3p/4	$-\sqrt{E/2}$	$-\sqrt{E/2}$	00
3	5p/4	$-\sqrt{E/2}$	$\sqrt{E/2}$	01
4	7p/4	$\sqrt{E/2}$	$\sqrt{E/2}$	11

Table 2.4.1.2: The Corresponding Phases and the Signal Points.

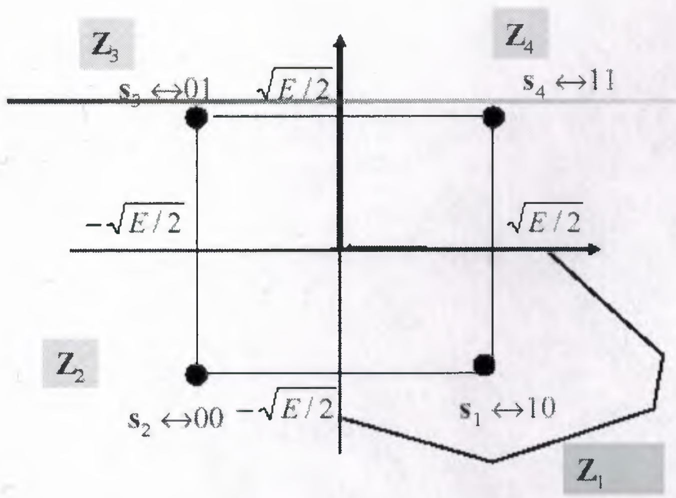


Figure 2.4.1.2: Signal Constellations of QPSK under the Gray Code.

2.4.2 Probability of Error of QPSK

The average probability of error of QPSK is given by:

$$P_e = \frac{1}{4}\sum_{i=1}^{4} P_e(m_1) \text{ .... (2.4.1.1)}$$

The decision regions are equal and the signal points are symmetric. Hence all Pe (mi) are equal. In the following, we give the result on $m1 = 10$.

For $m1 = 10$,

$$P_e(10) = 1 - P_C(10) \text{ .... (2.4.1.2)}$$

$$P_c(10) = P(X = (x_1, x_2) \in Z_1/10) = P\{0 < x_1 < \infty, -\infty < x_2 < 0/10\} \text{ .... (2.4.1.2)}$$

Where

$$x_j/10 \sim N(s_{1j}, \frac{N_O}{2}), \quad j = 1,2$$

and these two Gaussian random variables are independent. Thus

$$P_c(10) = P\{x_1 > 0/10\} P\{x_2 < 0/10 \text{ .... (2.4.1.3)}$$

$$= \left(1 - \phi\left(\frac{-\sqrt{E/2}}{\sqrt{N_O/2}}\right)\right)\phi\left(\frac{\sqrt{E/2}}{\sqrt{N_O/2}}\right) = \phi\left(\sqrt{\frac{E}{N_O}}\right)^2 \text{ .... (2.4.1.4)}$$

$$= \left(1 - Q\left(\sqrt{\frac{E}{N_O}}\right)\right)^2 = 1 - 2P_1 + P_1^2 \text{ .... (2.4.1.5)}$$

Where

$$P_1 = Q\left(\sqrt{\frac{E}{N_O}}\right) = P_{e,BPSK} \text{ .... (2.4.1.6)}$$

$$\Rightarrow P_e(10) = 2P_1 - P_1^2$$

$$P_e = \frac{1}{4}4(2P_1 - P_1^2) = 2P_1 - P_1^2 \text{ .... } (2.4.1.7)$$

Where

$$\sqrt{\frac{E}{N_O}} >> 1$$

The second term can be ignored so that for the coherent QPSK, the average probability of symbol error is

$$P_e \approx 2Q\left(\sqrt{\frac{E}{N_O}}\right) \text{.......... } (2.4.1.8)$$

2.5 M-ary Modulation Techniques

2.5.1 M-ary PSK Scheme:

The phase of carrier takes on one of M possible values, namely,

$$\theta_i = (2i-1)\pi / M, \quad i = 1, \ldots M \text{ .... } (2.5.1.1)$$

An M-ary signal set is represented as

$$s_i(t) = \sqrt{\frac{2E}{T}} \cos(2\pi f_c + \frac{(2i-1)\pi}{M}), \quad i = 1, \ldots M, 0 \leq t \leq T \text{ .... } (2.5.1.2)$$

Where T is the symbol duration and E is the signal energy per symbol. The carrier frequency

$$f_c = n_c / T \text{ .... (2.5.1.3)}$$

Where nc is a fixed integer. Similarly, we did for QPSK; each signal si (t) can be represented by the followin g two orthogonal functions with unit energy:

$$\phi_1(t) = \sqrt{\frac{2}{T}}\cos(2\pi f_c t) \quad \text{and } \phi_2(t) = \sqrt{\frac{2}{T}}\sin(2\pi f_c t) \text{ .... (2.5.1.3)}$$

Thus, the signal constellation of M-ary PSK is two-dimensional. The M messages are equally spaced one circle of radius and center at the origin. The coordinates of the received signal given $s_i(t)$ was transmitted is;

$$x_I = \sqrt{E}\cos\left(\frac{(2i-1)\pi}{M}\right) + n_I \text{ .... (2.5.1.4)}$$

$$x_Q = -\sqrt{E}\cos\left(\frac{(2i-1)\pi}{M}\right) + n_Q \quad \text{.......... (2.5.1.5)}$$

$$i = 1, 2, \ldots M$$

Where n_I and n_Q are Gaussian random variables with zero mean.

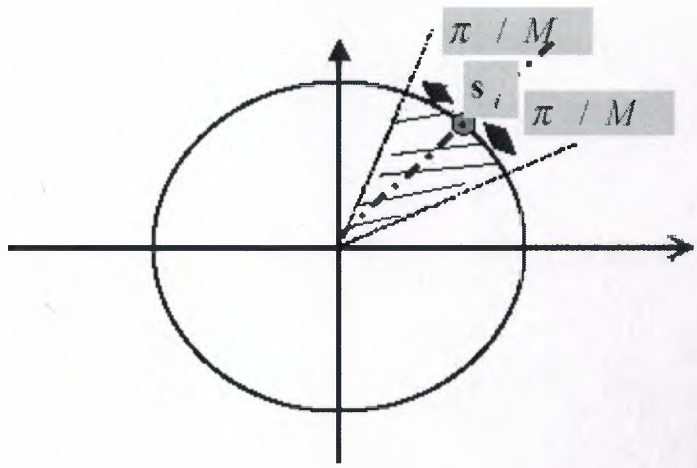


Figure 2.5.1.1: Signal Constellation for Octa- PSK ($M = 8$).

The probability of correct reception is to integrate the shaded area. This probability can be bounded by some bound. Therefore, for large values of $E/N0$, the probability of symbol error is approximately given by

$$P_{e,M-PSK} \approx 2Q\left(\sqrt{\frac{E}{N_o}}\sin\frac{\pi}{M}\right) \qquad M \geq 4 \text{..........} (2.5.1.6)$$

2.5.2 M-ary FSK

In an M-ary FSK scheme, the transmitted signals are defined by

$$s_i(t) = \sqrt{\frac{2E}{T}}\cos 2\pi\left[f_o + (i-1)\Delta f\right]t, \quad 0 \leq \mathrm{t} \leq \mathrm{T},\ i = 1,\ldots,M \text{ ....} (2.5.2.1)$$

Where is taken as an integer for convenience and is th e minimum frequency spacing such that adjacent signals are orthogonal.

For coherent M-ary FSK, the optimum receiver consists of a bank of M correlations or matched filters. At the sampling times $t = kT$, the receiver makes decisions based on the largest matched filter output. The probability of symbol error can be upper bounded by

$$P_{e,M-FSK} \leq (M-1)Q\left(\sqrt{\frac{E}{N_O}}\right) \quad .... (2.5.2.2)$$

Where $E = E_b(\log_2 M)$ the energy per symbol and M is is the size of the symbol set.

2.5.3 Bandwidth Efficiencies of M-ary Digital Comm. Systems (DCS)

Consider the bandwidth efficiencies in terms of bits per second per hertz (bps/Hz) of bandwidth of various digital modulation schemes. For a M-ary DCS, let Rb denote the bit rate and Rs symbol rate. Then

$$R_b = (\log_2 M)R_s \; .... (2.5.3.1)$$

For an M-ary PSK, QAM, DPSK, the null to null bandwidth is

$$B_{M,X} = \frac{2R_b}{\log_2 M} \; .... (2.5.3.2)$$

$$\rho = \frac{R_b}{B_{M,X}} = 0.5(\log_2 M) \;\; (bps/Hz) \; .... (2.5.3.3)$$

For a M-ary FSK, consider the spacing between frequency is minimum. Then the bandwidth is

$$B_{coh,MFSK} = \frac{(M+3)R_b}{2\log_2 M} \quad \text{..........(2.5.3.4)}$$

$$\rho = \frac{R_b}{B_{coh,MFSK}} = \frac{2\log_2 M}{M+3} \quad (bps / Hz) \text{ .... (2.5.3.5)}$$

M	2	4	6	16	32	64
ρ: PSK DPSK QAM	0.5	1	1.5	2	2.5	3
ρ : FSK	1	1	.75	.5	.3125	.1875

Table 2.5.3: Bandwidth Efficiencies of M- ary Signals.

M-ary PSK and M-ary QAM have 2-dimensional signal space and they are both bandwidth efficient (or called spectral efficient). MFSK has M-dimensional signal space and it is bandwidth inefficient.

2.6 Digital Modulation Techniques in Cellular Radio

Two broad classes of modulation techniques are available for use in mobile radio systems, analog and digital. First generation cellular radio systems use analog modulation, which itself can be divided into two major categories, amplitude modulation (AM) and frequency modulation (FM). In any transmission system, various performance factors must be considered, and certain factors must be traded against others. Two of the primary performance measures are susceptibility to noise and spectrum usage. FM generally has better noise immunity than AM, while AM generally uses less bandwidth. Both schemes however can be demodulated by inexpensive detector.

2.6.1 Digital Modulation

Advancing technology in semiconductors and digital signal processing techniques is making digital modulation more cost effective than ever before, even more so than analog systems. Because of this, digital modulation schemes are beginning to be deployed in newer cellular systems. Digital modulation can offer several advantages over analog modulation, such as improved noise immunity, higher resistance to channel problems (fading and multipath), multiplexing of different types of information (voice, video, data), higher security, and error correction. Furthermore, digital signal processors allow digital modulation schemes to be implemented completely in software, leading to dramatically reduced development and upgrade costs.

Digital modulation involves representing information with a symbol which can be in one of *m* finite states, where each state represents *n* binary bits of information. The choice of the particular modulation scheme, as with analog modulation, involves tradeoffs of various performance factors. Power efficiency and bandwidth efficiency are two of the primary measures traditionally associated with digital modulation. Power efficiency is essentially a measure of how well the information fidelity is maintained at low power levels. Bandwidth efficiency measures the amount of data that can be transmitted within a certain bandwidth. One example of this type of tradeoff is adding error correction to a digital modulation scheme. This increases power efficiency, while decreasing bandwidth efficiency. On the other hand, higher-level modulation techniques (more bits per symbol) increase bandwidth efficiency while decreasing power efficiency. For cellular systems, where spectrum has been the more limited resource, bandwidth efficiency usually has received greater emphasis. Variations of Phase Shift Keying (PSK), Frequency Shift Keying (FSK), and Quadrature Amplitude Modulation (QAM) are examples of digital modulation techniques which trade off power and bandwidth efficiencies. However, a different type of digital modulation, known as spread spectrum, deliberately uses a very wide transmission bandwidth for each user, but allows multiple users simultaneous access to the same portion of the spectrum. Cellular radio systems can be well served by spread spectrum modulation's ability to support a relatively large number of users within the same transmission spectrum.

2.6.2 Spread Spectrum Modulation

Code Division Multiple Access, or CDMA, is a particular spread spectrum technique that is gaining popularity in cellular systems because of the large capacity increase it brings. The CDMA standard (TIA standard IS-95) supports almost 16,000 users in the normal 800 MHz cellular band, as opposed to the less than 2,500 users supported by normal analog cellular standards or other digital cellular standards. The majority of cellular and PCS providers in the U.S. have committed to deploying CDMA systems, and operational trials of a CDMA base station began in 1994. Bes ides the increased number of users, other important features of spread spectrum modulation are interference rejection and resistance to multipath fading.

In spread spectrum modulation, the narrowband information signal is spread into a very wideband signal by use of a pseudo-noise (PN) sequence which has a much higher bit rate than the information signal. This PN sequence is so named because its autocorrelation is similar to the autocorrelation of a random binary sequence, although it is actually deterministic. The symbol rate of the PN sequence is also known as the chip rate, and any one of the symbols in the PN sequence is called a chip. Each user is assigned a different PN code, which is generated to be approximately orthogonal to the codes of other users, allowing each receiver to demodulate the intended signal with little interference from the other users. PN codes are usually generated by a sequential logic circuit consisting of shift registers and XOR feedback logic, a circuit known as a linear PN se quence generator.

Two primary techniques are used in spread spectrum modulation, direct sequence spread spectrum (DS-SS), and frequency hopped spread spectrum (FH- SS). Direct sequence spread spectrum is the method implemented by the aforementioned CDMA s tandard. The primary difference between DS-SS and FH-SS is the way the PN sequence is utilized. In direct sequence systems, the PN code is directly combined with the information signal to produce a new, wider bandwidth base band signal to modulate the carr ier. In frequency hopped systems, however, the PN code is used to determine which carrier frequency is used for each hop. DS-SS and FH-SS each have its own advantages and disadvantages. In some cases a hybrid of the two techniques is used to minimize the d isadvantages of either single method.

2.6.3 Performance Tradeoffs for CDMA DS-SS Systems

As discussed earlier, direct sequence spread spectrum modulation offers CDMA systems several advantages over traditional cellular systems. Performance tradeoffs must be considered, however, even in CDMA systems. In terms of the performance measures normally applied to digital modulation, DS-SS does very well in terms of power efficiency when compared with more conventional digital modulation. In terms of bandwidth efficiency, DS-SS also does well when compared to other types of digital modulation, as long as the number of users is large. From a system-level viewpoint, the primary parameters to be considered in implementing a CDMA system are coverage, quality, and capacity. These general categories are actually similar to the tradeoffs in any cellular system, although the specific parameters affecting each category may be different between CDMA and other types of cellular radio.

2.6.4 Digital Sampling Power Analyzer for GSM and CDMA

The digital modulation methods employed for cellular telephone and other wireless communications present a challenge for making accurate peak power measurements. Although there are many different implementations, the two basic types of modulation systems in use are time division multiple access (TDMA), and code division multiple access (CDMA).

Of the many TDMA systems, GSM (Global System for Mobile communication) is in widespread use in Europe and Asia and will soon appear in the USA. The RF envelope is in the form of 542.8 μsec pulses which are located within a 576.9 μsec timeslot, each containing 147 bits of information. The power- versus time relationship for each pulse is controlled within narrow limits for both turn -on and turn-off. This is necessary to prevent interference between adjacent time slots which are assigned to different transmitters. A GSM transmitter has only 28 μsec to ramp up to full power, a 70dB dynamic range, while remaining within a specified power/time profile. The profile defines limits for overshoot and rise-time as well as fall-time. A peak power video bandwidth of at least 1 MHz is required to assure compliance with the profile.

2.6.4.1 CW Power Measurement

The average power of an unmodulated RF carrier can be measured accurately by a CW type power meter with a thermoelectric or diode detector. The thermoelectric detector offers good accuracy over a dynamic range of about 50dB. The diode detector can provide a much larger dynamic range, about 90dB. The average power of a modulated RF carrier which has constant envelope amplitude, e.g. FM, can also be measured accurately using these techniques.

For modulated RF carriers with non- constant envelope amplitude, e.g. pulse modulation, the thermoelectric detector will still respond accurately to the average power of the signal. The long time constant associated with thermal effects prevents this type of detector from following the envelope at the modulation rate, and therefore, is unable to provide any measure of instantaneous power.

The conventional CW type diode detector will also respond accurately, provided that it is used at low power in its square-law response region. This usually corresponds to a power at the diode of no more than -20 dBm or 10 μW. Higher input power is accommodated by placing an attenuator between the input signal and the diode. In a CW type detector the diode is loaded by a fairly large capacitance which filters the noise and improves sensitivity. The resulting time constant is long compared with modulation frequencies and prevents the detector from following the instantaneous value of the envelope.

2.6.4.2 Pulse Power Measurement

Pulse power is determined traditionally by adjusting the average power reading of a CW type power detector for the duty cycle of a modulating pulse. In this way, a peak power measurement of moderate accuracy can be obtained from an average power value, provided certain conditions are met. First, the modulation must consist of constant amplitude rectangular pulses of known duty cycle (on/off ratio). Second, the linear power range of the detector must not be exceeded by the peak power applied. This requirement is often overlooked, resulting in invalid readings or damage to the detector. The pulse power measurement technique is not suitable for digital modulation systems in which the duty cycle is not constant and pulse amplitude and shape varies.

2.6.4.3 Peak Power Measurement

What is needed for complex digital modulation is true instantaneous power measurement with a bandwidth of at least 1 MHz. The Boonton Model 4400A RF Peak Power Meter and Model 4500A RF Peak Power Meter / Analyzer provide the capability to measure peak power accurately with a dynamic range of as much as 60 dB and a demodulated video bandwidth as large as 35 MHz, with the Model 56518 Peak Power Sensor. Knowledge of the modulation method or modulating signal is not required for accurate average and peak power measurements.

In simplified form the Model 4400A/4500A peak power measuring s ystem consists of the following:

1. A Peak Power Sensor containing a dual diode detector with wide RF bandwidth (up to 40 GHz) and a narrower video bandwidth (3 to 35 MHz), and a precision log amplifier compatible with the video bandwidth.
2. A fast sample and hold amplifier, asynchronous with respect to the input signal.
3. An analog to digital converter which operates at the sampling rate.
4. A Digital Signal Processor (DSP) for processing the samples at high speed.

A built-in, digitally controlled, 1 GHz precision CW power calibrator with internal pulse modulation. A host processor to control I/O interfaces and all sub -processes. A VGA processor for full color display of numerical and graphical data.

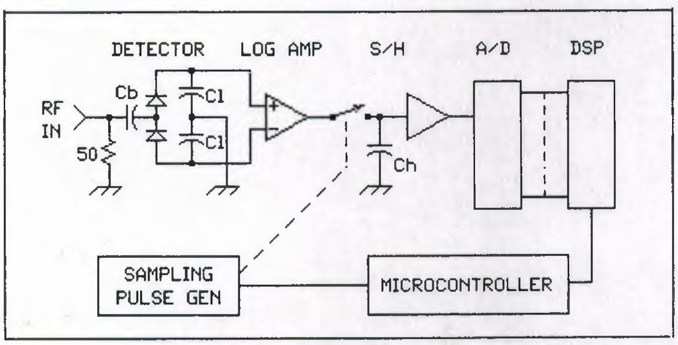


Figure 2.6.4.3: Simplified view of peak power measurement system.

2.6.5 Precision Digitally Controlled Calibrator.

In order to eliminate the error associated with diode non -linearity, a calibration table is created for each sensor, which stores the response to a series of precision power levels covering the effective dynamic range along with additional data. This is accomplished automatically by a precision, digitally controlled, 1 GHz RF power source and control program. The resulting calibration table is extended by interpolation to create a power entry for all possible A/D converter values. This allows the DSP to calculate the instantaneous power of each individual sample of the RF envelope.

Average power is calculated by summing the instantaneous power values. The non-linear relationship between instantaneous RF power level and diode output is resolved before any averaging is done, thus, the result is correct for an arbitrary waveform. It is this characteristic which separates this method of power measurement from the conventional average power method in which the output of the detector is averaged before A/D conversion.

Random power samples in time can be processed by the DSP to provide results in any form needed by an application. This includes peak power versus time, peak power relative to a trigger event, average power over various time intervals, and peak to average ratio, and maximum peak power in a time interval, etc. It does not matter that the samples are disordered in time. For a stationary signal, the sum of the random samples over arbitrary, equal length time intervals is the same, provided there is no periodic relationship between the sampling rate and the modulating signal. In addition, there must be a sufficient quantity of samples taken to ensure adequate coverage. The advantage of a high sampling rate is the ease of accumulating a large number of sample points for each reading.

If the detected signal is stationary or quasi-stationary in time, the waveform of the RF envelope can be re-constructed from the random samples. In conventional pulse or linear amplitude modulation, the RF carrier envelope and thus the detected signal correspond closely to the modulating signal waveform. In TDMA systems, the exact shape of the pulsed RF envelope is critical for optimum performance.

The Boonton Model 4400A is particularly suited to applications in which peak power versus time is the primary concern. Statistical Methods Using the Model 4500A Digital modulation methods in which amplitude and phase modulation are combined in a multilevel arrangement to represent a group of bit values from one or more data streams, and multiple carrier spread spectrum systems, such as CDMA, do not have simple envelope waveforms which can be directly related to modulation parameters. Traditional parameters such as modulation depth and modulation index are not meaningful because the peak to average power ratio of the modulated carrier is a complex function of the data stream content, rather than the amplitude of the modulating signal. The resulting noise- like character of these signals suggests a statistical approach to analysis.

The Boonton Model 4500A RF Peak Power Meter / Analyzer is designed to extract the statistical properties of these signals in addition to the time related properties discussed above.

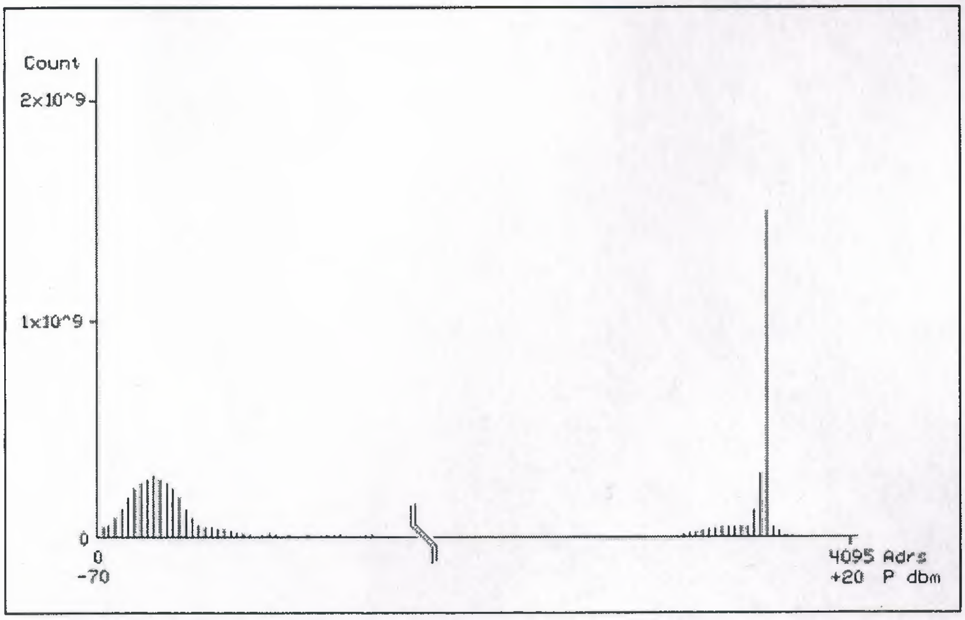


Figure 2.6.5.1: Sample Count Array in Memory.

Since the power of the individual random samples is know n, they can be sorted and counted by power level. For a 12-bit A/D converter system there are 4096 possible power levels. If a memory array of this size is established, each address corresponds to one of the possible power levels. With the array initially cleared to zeroes, the value of each sample taken is interpreted as an offset address into the array, and the count stored at that location is incremented by one. As this process is repeated millions of times, the array contents approaches N times the probability function for the signal, where N is the total count of the entire array. The count at any address divided by N is equal to the probability of occurrence of the power level represented by that address.

The measurement process must keep track of the total number of samples taken in order to scale the results properly and to estimate the statistical uncertainty, which is inversely proportional to the square root of the number of samples. A high degree of confidence is assured by a very large number of samples and a long running time. A word size of 31 -bits will account for at least 2.1 billion (2.1×10^9) samples without overflowing any counter. Even at 500,000 samples per second, the running time will be at least 4,200 seconds or 1.17 hours before any possible overflow can occur. The measurement could be allowed to run indefinitely with a suitable decimation process, but not without loss of some information. Unfortunately, ordinary right shifting of the data results in the loss of the small counts, which are typically the most important ones. As a result, the measurement is automatically stopped before any overflow occurs.

Well-defined modulation processes may show convergent results after only a few million samples are collected. Running times of only a few seconds may be completely adequate for these applications.

It is convenient for analytical reasons to organize statistical data into one of several standard forms. The Boonton Model 4500A displays the data both numerically and graphically on a color VGA CRT. The following symbols are used throughout the formulas:

Y is a discrete random variable with a range equal to all possible sampled values of carrier power. Y is a specific power value contained in Y. PDF.

The probability distribution functions of Y. The PDF is the percentage of time that the power is equal to a specific value, y. The percentage ranges from 0 to 100%, and the power extends over the entire dynamic range of the system. PDF expressed as a percentage is:

PDF = P(y) = 100*P[Y=y] where y ranges over all values in Y, $0 \leq P(y) \leq 100\%$. As samples are continuously taken, the sample space is rescaled to 100%. This conforms to the requirement that all P(y) add up to 100%. $\Sigma P(y) = 100\%$ where y ranges over all values in Y.

The PDF is useful for analyzing the nature of modulating signals. Sustained power levels such as the flat tops of pulses or steps show up as lines. Random noise produces a Gaussian shaped curve.

CDF. The cumulative distribution functions of Y. The CDF is the probability that the power is less than or equal to a specific value, y. The CDF is non-decreasing in y, that is, the graph of CDF versus y cannot have negative slope. The maximum power sample taken will lie at 100%. CDF expressed as a percentage is:

CDF = Q(y) = 100*P[$Y \leq y$] where y ranges over all values in Y, $0 \leq Q(y) \leq 100\%$

$Q(y_{max}) = 100\%$, and also, just as for PDF above, P(y) = 100%

CCDF. It is often more convenient to use the complementary CDF, or CCDF, or 1 -CDF, sometimes called the "upper tail area". The CCDF is the probability that the power is greater than a specific power value. CCDF is non-increasing in y and the maximum power sample lies at 0%. CCDF expressed as a percentage is:

CCDF = 1-Q(y) = 100*P[Y>y] where y ranges over all values in Y is:

$0 \leq 1-Q(y) \leq 100\%$, $1-Q(y_{max}) = 0\%$

In a non-statistical peak power measurement the peak-to-average ratio is the parameter which describes the headroom required in linear amplifiers to prevent clipping or compressing the modulated carrier. The meaning of this ratio is easy to visualize in the case of simple modulation in which there is close correspondence between the modulating waveform and the carrier envelope. When this correspondence is not present, the peak -to-average ratio alone does not provide adequate information. It is necessary to know what fraction of time the power is above (or below) particular levels. For example, some digital modulation schemes produce narrow and relatively infrequent power peaks which can be compressed with minimal effect. The peak-to-average ratio alone would not reveal anything about the fractional time occurrence of the peaks, but the CDF or CCDF clearly show this information. See in Figure 2.6.5.2

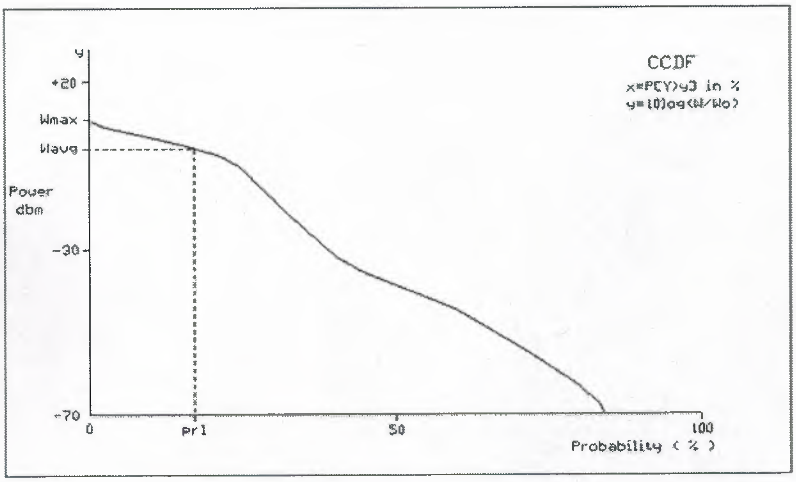


Figure 2.6.5.2:A CCDF with Expanded Time Axis.

Note that the CCDF plot in Figure 2.6.5.2 has probability in percent on the X-axis and instantaneous envelope power in dBm on the Y-axis. The usual practice in texts on statistics is to show probability on the Y-axis. This change is made so that power always appears on the Y-axis in all instrument display modes. Keep in mind that power is the independent variable. The probability scale in Figure 2.6.5.2 has been expanded to better show the region around zero. Assume this plot represents a full length run of one hour plus with $2x10^9$ samples counted. On the Y-axis at probability = 0% is the maximum peak power which occurred during the entire run. Or, there is zero probability that a power level higher than W_{max} occurred during the run. At probability = 1% is the power level W_{clp} which was exceeded only 1% of the time during the entire run.

Note that this analysis does not depend upon any particular test signal, nor upon synchronization with the modulating signal and there is no time base involved. In fact, the analysis can be done using actual communication system signals. Normal operation is not disturbed by the need to inject special test signals. This type of analysis is particularly suited to the situation in which the bit error rate (BER) or some other error rate measure is correlated with the percentage of time that the signal is corrupted. If known short intervals of clipping are tolerable, the CCDF can be used to determine optimum transmitter power output. The CCDF is also used to evaluate various modulation schemes to determine the demands that will be made on linear amplifiers and transmitters and the sensitivity to non - linear behavior.

The Boonton Model 4500A provides the CCDF as well as the CDF and PDF graphs along with power and pulse parameters for a comprehensive analysis of pulse or spread spectrum digital modulation.

CHAPTER THREE

DESIGN AND TECHNIQUES IN DIGITAL COMMUNICATION

3.1 Techniques in Digital Communication

Digital techniques have revolutionized the communication industry. New applications and techniques in communication electronics, including digital quadrature sampling, digital error correction, direct digital synthesis, digital modulation, and digital frequency analysis, to name a few, have allowed smaller, lighter, less expensive, more accurate circuits and instruments to be developed.

Digital communication theory can be traced back to the development of information and sampling theory at Bell Laboratories in the 1940's. At that time, digital techniques, although useful, required expensive, cumbersome equipment to implement. Modern advances in integrated circuits have made digital techniques much more applicable to communication applications. Digital and mixed-signal integrated circuits that provide high speed and high performance have become inexpensive. This project shall describe three categories of digital communication theory: digital detection, digital processing, and digital synthesis.

3.1.1 Digital Detection

Digital detection techniques involve more than simple analog-to-digital conversion of conventionally detected analog signals. Digital receivers employ digital sampling to down-convert and detect RF and IF signals directly. This technique bypasses the baseband analog portion of a conventional analog receiver. Consequently, the noise, drift, and sensitivity issues related to base band analog signal processing are eliminate d.

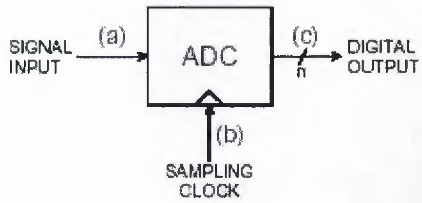


Figure 3.1.1.1: Simple Digital down Conversion.

In order to explain the digital down conversion process, it is helpful to examine digital sampling in the simple analog-to-digital converter (ADC) shown in figure 3.1.1.1 The ADC is used to sample a sinusoidal input signal to produce a digital data stream. The sampling rate (*fS*) is less than the carrier frequency (*fC*) and consequently, below the Nyquist frequency of the input signal. This under sampling technique is employed intentionally to digitally down-convert the sinusoid to a lower intermediate frequency.

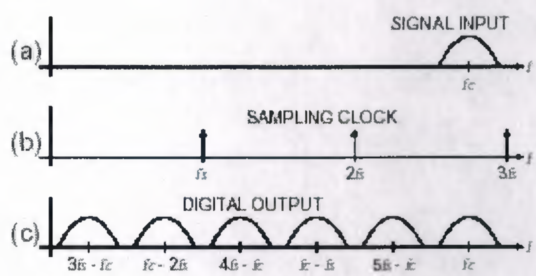


Figure 3.1.1.2: Spectrum of Digital down Conversion.

Examining the frequency domain of the simple ADC sampling clearly explains this technique. Figure 3.1.1.2 shows the spectrum of (a) the input sinusoidal signal,

(b) the ideal sampling clock, and (c) the resulting digital output. Notice that the digital output signal retains the spectral properties of the input signal repeated at the sampling clock frequency. Because the sampling frequency is below the Nyquist frequency of the input signal, the digital output data contains signal components in frequency bands lower than the original input signal. By choosing the sampling frequency properly, this down conversion technique can effectively translate a high frequency input to lower frequency digital data. Note that there is a bandwidth limitation on the input signal to avoid spectral overlap and signal aliasing. The input signal must not have signal components outside of a limited frequency band. Equation 3.1.1.1 defines the bandwidth limitation for digital down conversion, where *fMAX* is maximum frequency offset of any signal component from the carrier frequency, and *fS* is the sampling frequency.

$$f\max = \pm\frac{fs}{4} \;\ldots\ldots\ldots\; (3.1.1.1)$$

Another technique that is employed within digital receivers is quadrature demodulation. Quadrature demodulation involves the separation of a sinusoidal signal into its orthogonal in-phase and quadrature (I/Q) components. In analog receivers, this involves splitting the received signal into two signals and mixing those with t wo local oscillator signals having a 90degree phase difference. Again, the analog signal is susceptible to errors associated with offsets, noise, drift, and channel imbalance. The DC offset and drifts, along with the low - frequency noise translate directly into errors in the detected I/Q signals. Also, separating the signal into two channels for I and Q makes the analog circuit susceptible to errors if the two paths are not exactly matched in magnitude and maintained at the required 90degree phase difference. The errors associated with analog quadrature demodulation are shown graphically in figure 3.1.1.3. Note that an ideal quadrature demodulator will have an I/Q plot that traces a perfect circle with its center at the origin (0,0). DC offsets in the analog demodulator circuit cause the center to move from the origin. Elliptical characteristics in the I/Q plot are due to gain imbalance and quadrature phase error. The imperfections in the analog quadrature demodulator cause harmonic distortion and decrease character seperation in the demodulated data output.

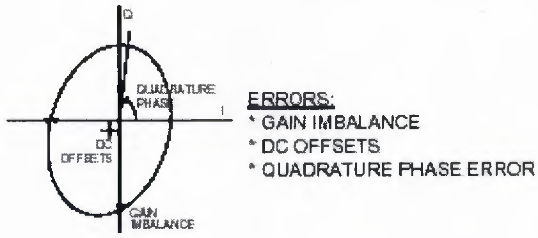


Figure 3.1.1.3: Errors Involved in Quadrature Demodulation.

An alternative to analog quadrature demodulation involves the digital quadrature demodulation shown in the block diagram of figure 3.1.1.4 within the digital quadrature demodulator; an input signal is sampled at a local oscillator (LO) clock rate that is exactly four times the input carrier frequency. The time between consecutive ADC samples is exactly equal to one-fourth the period of the carrier, which corresponds to 90degree. in phase. The digital data represents the input sampled at 90degree. intervals. The ADC output provides a data stream consisting of a repetitive pattern of I, Q, -I, and –Q measurements. The digital data is then separated into I and Q components by a demultiplexer that switches every other sample into two parallel digital paths. The sign inversion in each path is removed by multiplying each data stream by +1 and –1 alternately. The resulting I and Q outputs are digitally filtered and decimated. Because the I and Q components of the signal are detected digitally in a single ADC channel, without requiring a baseband analog step, the errors associated with gain imbalance, DC offsets, and quadrature phase are eliminated. Also, the flexibility and simplification of this approach make it very attractive for digital receivers.

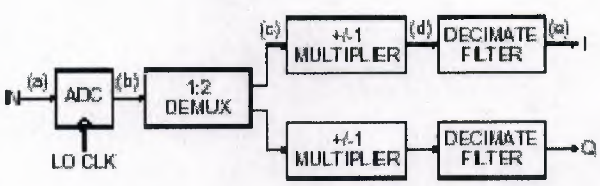


Figure 3.1.1.4: Digital Quadrature Demodulation Block.

Diagram again, examining the spectral properties of the digital quadrature demodulation process is informative. Figure 3.1.1.5 shows the spectrum of the various stages in the digital quadrature demodulation process. Sampling reproduces the input spectrum at the sampling rate. The I/Q signal separation is equivalent to decimation by two and causes the signal spectrum to be spread accordingly. Because of this decimation by two, the bandwidth of the input signal must be further constrained. Equation 3.1.1.2 defines the bandwidth limitation for quadrature demodulation, where *fMAX* is maximum frequency offset of any signal component from the carrier frequency, and *fS* is the sampling frequency.

$$f\max = \pm\frac{fs}{8} \text{ .... } (3.1.1.2)$$

Multiplication by a repeating pattern of +1 and −1 is comparable to mixing the data stream with a sampled sinusoid at the carrier frequency. In effect, the ± 1 multiplier causes a frequency shift of the input signal equal to the carrier frequency. Because the frequency shift to baseband is preformed digitally, the data does not contain errors normally associated with analog DC offsets. The resulting digital I/Q data are then lowpass filtered in a programmable digital filter. Due to the precise nature of digital electronics, the digital processing does not introduce any errors in the signal data.

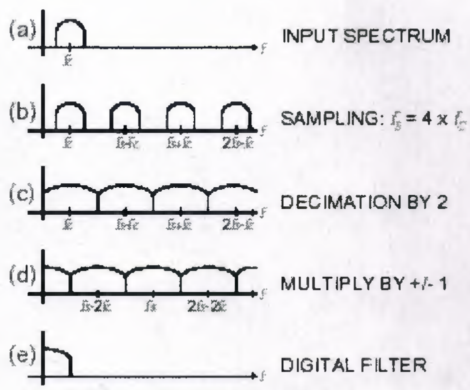


Figure 3.1.1.5: Spectrum of Digital Quadrature Demodulation.

The digital quadrature demodulator may apply the digital down conversion technique in conjunction with quadrature sampling to simultaneously down convert and quadrature detect a received signal. This is accomplished by sampling at a frequency that is not exactly four times the carrier frequency. The frequencies that provide quadrature demodulation and digital downconversion are calculated with equation 3.1.1.3, where fS is the sampling frequency, fC is carrier frequency, and n is a positive integer.

$$fs = \frac{4 * fc}{2 * (n+1)} \text{..........} \quad (3.1.1.3)$$

3.1.2 Digital Processing

Powerful, real-time digital processing is made possible by the current speed and performance of digital signal processor (DSP) integrated circuits. DSP algorithms are applied to reduce noise and distortion, extract information, separate signals, measure signal parameters, encode new information, or modulate the signal for outgoing transmission. A few examples of DSP algorithms include digital filtering, fast Fourier transforms, modulation analysis, adaptive noise cancellation, error detection and correction, and data compression. DSP devices include both dedicated -purpose and general -purpose devices. Dedicated-purpose DSPs tend to be faster but less flexible than the general-purpose DSPs. Dedicated-purpose DSPs are common on digital communication instruments due to the high data throughputs that are often necessary to stream continuous or high duty -cycle signals. General -purpose DSP devices are often employed to accomplish customized post-processing functions and user-defined algorithms. With a general -purpose DSP and the correct hardware peripherals, a wide variety of custom applications can be implemented in DSP firmware.

3.1.4 Digital Synthesis

Digital synthesis techniques involve the creation of complicated waveforms using digital electronics. The basic building block for digital synthesis is the direct digital synthesizer

(DDS), which is shown in the block diagram of figure 3.1.4.1 The DDS produces a waveform at a given frequency by accumulating phase at a higher frequency. The value stored in the frequency register is added to an accumulator once every clock cycle. The resulting accumulator value is applied to a sine look-up table that computes the sine of that accumulator value. The full-scale range of the accumulator corresponds to one period or 360 degree of a sinusoid. The digital sinusoidal data output of the look-up table is applied to a digital to- analog converter (DAC) and reconstruction filter for conversion to an analog waveform. In order to reconstruct the sinusoidal waveform, the synthesized frequency must be less than one-half the LO clock frequency.

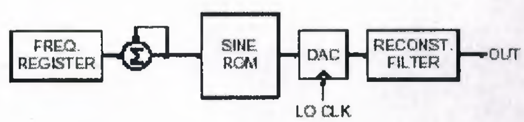


Figure3.1.4.1: Block Diagram of DDS.

The digital nature of the DDS limits the output frequency to discrete values that are defined by equation 3.1.4.1, where *fOUT* is the output frequency, *fLO* is LO clock frequency, *n* is the number of bits in the accumulator, and Δ is an integer between 1 and 2n-1 that is used to select the output frequency.

$$fout = \frac{flo * \Delta}{2^n} \quad .... (3.1.4.1)$$

Another device used for digital synthesis is the DDS modulator which combines a digital quadrature modulator with a DDS synthesizer. Complex modulation functions are accomplished in digital circuitry with high precision and without errors or noise. The resulting analog output consists of a carrier with programmable frequency upon which a programmable modulation is applied. Flexible modulations such as AM, FM, PM, FSK, PSK, QPSK, QAM, and Chirp are easily accomplished with the DDS modulator.

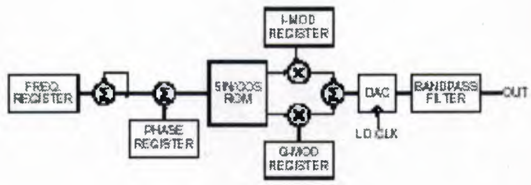


Figure 3.1.4.2: Block Diagram of DDS Modulator.

A hybrid DDS/phase-locked loop (PLL) combination is another frequency synthesizer implementation that provides broad output frequency angle, excellent spectral purity, and extremely fine frequency resolution. The DDS -driven PLL of figure 3.1.4.3 uses a DDS as the reference for the frequency multiplying PLL. This technique creates a sinusoidal output with a frequency that is N times the DDS frequency. Thus, the output frequency resolution in N times the DDS frequency resolution. Note that the frequency resolution of the DDS-driven PLL varies with N, and is not constant throughout the entire output frequen cy range. Also, because the spurious content of the DDS output exceeds that of a high -grade reference oscillator, the output may be degraded if steps are not taken to filter the spurious tones from the DDS output.

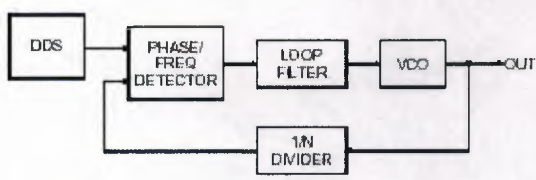


Figure 3.1.4.3: Block Diagram of DDS- Driven PLL.

Another type of hybrid DDS/PLL is the DDS-offset PLL shown in figure 3.1.4.4 The DDS-offset PLL implementation uses a DDS to offset the PLL output frequency, providing a finer frequency resolution than the DDS-driven PLL. Also, because a high -grade oscillator is used as the frequency reference, the output spectrum is less sensitive to spurious tones in the DDS output. The PLL loop filter will reject the noise and spurious content of the DDS as well as other sources of error in the PLL. The DDS-offset PLL creates a highfidelity sinusoidal output with extremely fine frequency resolution.

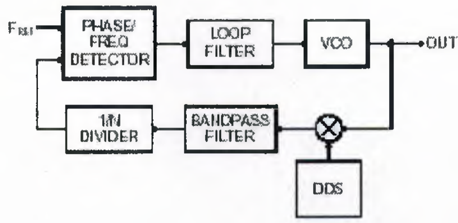


Figure3.1.4.4: Block Diagram of DDS-Offset PLL.

3.1.5. Implementation

The implementation of digital communication techniques relies upon the recent advancements in semiconductor devices for data conversion and digital processing. Fast ADCs and DACs, dedicated digital communication devices, and powerful DSPs make modern electronic instruments smaller, less expensive, and more flexible than their analog predecessors. An example of an implementation that applies many of the previously defined digital communication techniques is ZTEC Inc.'s ZT200VME IF Processor module, shown in block diagram form in figure 3.1.5.1 The ZT200VME module contains three downconverting digital quadrature demodulators, six dedicated digital filter DSPs, one general-purpose DSP, one DDS modulator, and all necessary peripheral memory and interface electronics. The general-purpose DSP is the core of this instrument, allowing it to function in a variety of applications.

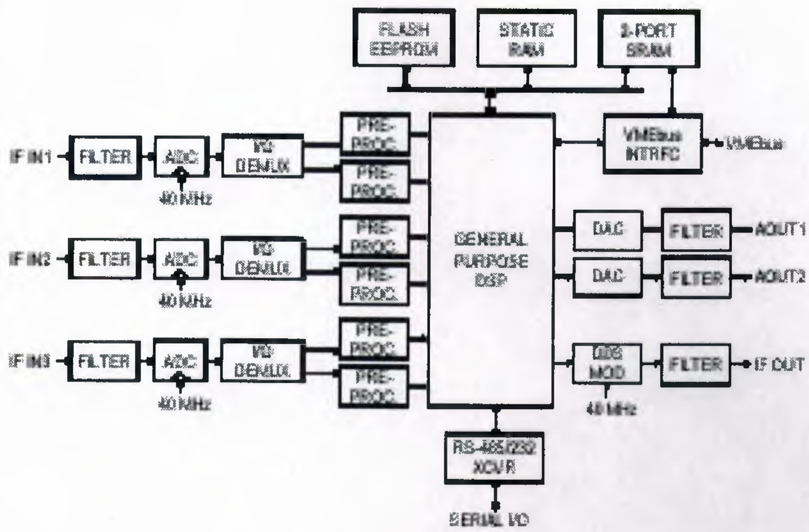


Figure 3.1.5.1: Block Diagram of ZT200VME.

The flexibility of the digital processing on the ZT200VME is demonstrated by its ability to function in significantly different operating modes. A change in its on-board DSP firm ware allows it to function as a digital receiver, a modulation analyzer, a digital transmitter, an adaptive beam former, or a feedback controller. The applications are varied and limitless. Also, many ZT200VME modules may be interconnected through the front-panel time-domain multiplexed high-speed serial port. This port enables multi-processing communication without tying up the bandwidth of the VMEbus.

3.1.6 Applications

The applications for digital communication electronics are varied and growing. Although many of ZTEC Inc.'s products are applied in test and measurement applications, others perform fixed functions as part of a permanent installation. One example is the ZT201VXI Field Control Module that has been applied to the feedback/feedforward contro l of the microwave fields in a charged particle accelerator .A photograph of the ZT201VXI Module is shown in figure 3.1.6.1. This module applies the three techniques of digital detection, digital processing, and digital synthesis. The application of these techniques to the control of the microwave fields in a charged-particle accelerator accomplishes more precise and complicated algorithms than were previously possible with analog instruments.

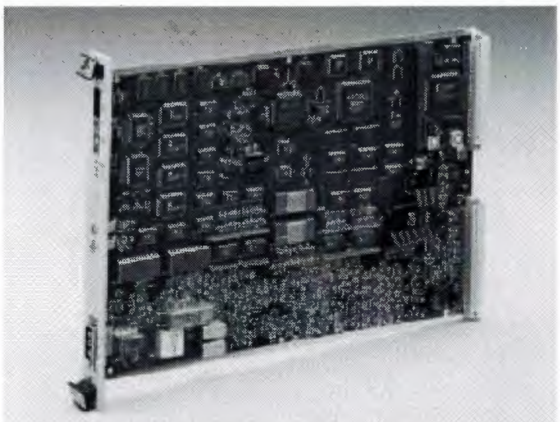


Figure 3.1.6.1: ZT201VXI Module the ZT201VXI Detects Two Received IF Signals (one feedback, one feedforward) in Digital Quadrature Detectors.

The resulting digital waveforms are processed in a Texas Instruments TMS320C50 DSP. The processing involves digital filtering, amplitude saturation detection, PID control, integrator antiwindup, stimulus and response characterization, history buffering, and transient recording. The ZT201VXI's control output signals are used to modulate an IF signal that drives the forward path of the closed-loop control system.

3.2 Crystal Oscillators

Crystal oscillators are one of the fundamental building blocks of electronic systems.Where microprocessors are used in personal computers and embedded applications, there crystal oscillators help to maintain stable frequencies. In cellular ph ones, temperature-compensated oscillators (TCXOs) allow the devices to be used anywhere in the world and withstand extreme heat or cold. Crystals are also being used in video cameras and optical systems, automotive electronic systems, and networking applications. The crystal is far from a static component. Discrete oscillator modules such as TCXOs and voltage- controlled oscillators

Programmable oscillators, which first made a splash approximately a year ago, are becoming stable enough and affordable enou gh to be used in some volume-production applications, primarily communications, to reduce cycle times.

3.3 Digital to Analog Conversion

One common requirement in electronics is to convert signals back and forth between analog and digital forms. Most such conversions are ultimately based on a *digital-to-analog converter* circuit. Therefore, it is worth exploring just how we can convert a digital number that represents a voltage value into an actual analog voltage.

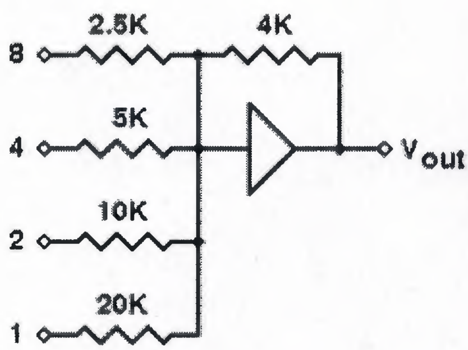


Figure 3.3.1: A Basic Digital to Analog Converter.

The circuit in figure 3.3.1 is a basic digital-to-analog (D to A) converter. It assumes a 4-bit binary number in Binary-Coded Decimal (BCD) format, using +5 volts as a logic 1 and 0 volts as a logic 0. It will convert the applied BCD number to a matching (inverted) output voltage. The digits 1, 2, 4, and 8 refer to the relative weights assigned to each input. Thus, 1 is the Least Significant Bit (L SB) of the input binary number, and 8 is the Most Significant Bit (MSB).

If the input voltages are accurately 0 and +5 volts, then the "1" input will cause an output voltage of $-5 \times (4k/20k) = -5 \times (1/5) = -1$ volt whenever it is a logic 1. Similarly, the "2," "4," and "8" inputs will control output voltages of -2, -4, and -

8 volts, respectively. As a result, the output voltage will take on one of 10 specific voltages, in accordance with the input BCD code.

Unfortunately, there are several practical problems with this circuit. First, most digital logic gates do not accurately produce 0 and +5 volts at their outputs. Therefore, the resulting analog voltages will be close, but not really accurate. In addition, the different input resistors will load the digital circuit outputs differently, which will almost certainly result in different voltages being applied to the summer inputs.

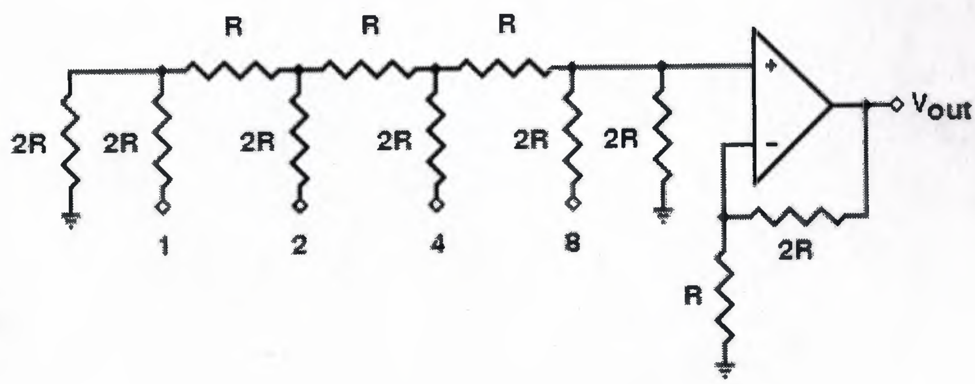


Figure 3.3.2: Another Circuit D to A conversion.

The circuit above performs D to A conversion a little differently. Typically the inputs are driven by CMOS gates, which have low but equal resistance for both logic 0 and logic 1. Also, if we use the same logic levels, CMOS gates re ally do provide +5 and 0 volts for their logic levels.

The input circuit is a remarkable design, known as an R -2R ladder network. It has several advantages over the basic summer circuit we saw first:

1. Only two resistance values are used anywhere in the entire circuit. This means that only two values of precision resistance are needed, in a resistance ratio of 2:1. This requirement is easy to meet, and not especially expensive.

2. The input resistance seen by each digital input is the same as for every other input. The actual impedance seen by each digital source gate is 3R. With a CMOS gate resistance of 200 ohms, we can use the very standard values of 10k and 20k for our resistors.
3. The circuit is indefinitely extensible for binary numbers. Thus, if we use binary inputs instead of BCD, we can simply double the length of the ladder network for an 8-bit number (0 to 255) or double it again for a 16-bit number (0 to 65535). We only need to add two resistors for each additional binary input.
4. The circuit lends itself to a non-inverting circuit configuration. Therefore we need not be concerned about intermediate inverters along the way. However, an inverting version can easily be configured if that is appropriate.

One detail about this circuit: Even if the input ladder is extended, the output will remain within the same output voltage limits. Additional input bits will simply allow the output to be subdivided into smaller increments for finer resolution. This is equivalent to adding inputs with ever-larger resistance values (doubling the resistance value for each bit), but still using the same two resistance values in the extended ladder.

The basic theory of the R-2R ladder network is actually quite simple. Current flowing through any input resistor (2R) encounters two possible paths at the far end. The effective resistances of both paths are the same (also 2R), so the incoming current splits equally along both paths. The half-current that flows back towards lower orders of magnitude does not reach the op amp, and therefore has no effect on the output voltage. The half that takes the path towards the op amp along the ladder can affect the output.

The most significant bit (marked "8" in the figure) sends half of its current toward the op amp, so that half of the input current flows through that final 2R resistance and generates a voltage drop across it. This voltage drop (from bit "8" only) will be one-third of the logic 1 voltage level, or 5/3 = 1.667 volts. This is amplified by the op amp, as controlled by the feedback and input resistors connected to the "-" input. For the components shown, this gain will be 3. With a gain of 3, the amplifier output voltage for the "8" input will be 5/3 × 3 = 5 volts.

The current from the "4" input will split in half in the same way. Then, the half going towards the op amp will encounter the junction from the "8" input. Again, this current "sees" two equal-resistance paths of 2R each, so it will split in half again. Thus, only a quarter of the current from the "4" will reach the op amp. Similarly, only 1/8 of the current from the "2" input will reach the op amp and be counted. This continues backwards for as many inputs as there are on the R -2R ladder structure.

The maximum output voltage from this circuit will be one step of the least significant bit below 10 volts. Thus, an 8-bit ladder can produce output voltages up to 9.961 volts (255/256 × 10 volts). This is fine for many applications. If you have an application that requires a 0-9 volt output from a BCD input, you can easily scale the output upwards using an amplifier with a gain of 1.6 (8/5).

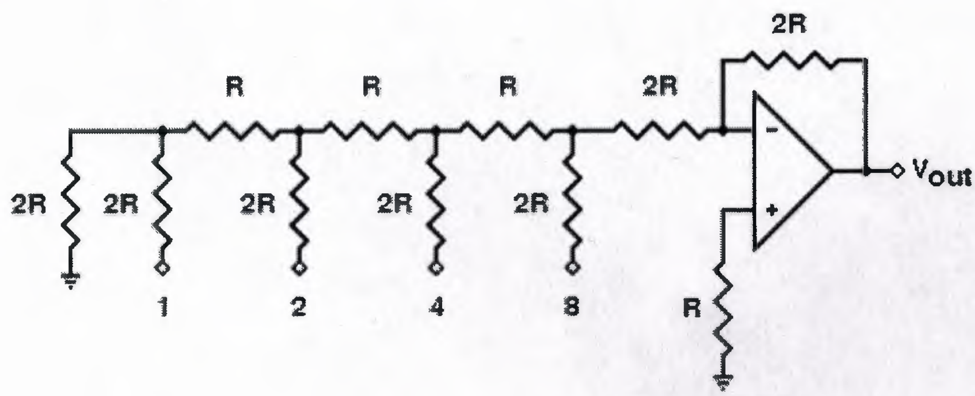


Figure 3.3.3: Another Circuit D to A Conversion.

If you want an inverting D to A converter, the circuit shown above will work well. You may need to scale the output voltage, depending on your requirements.

Also, it is possible to have a bipolar D to A converter. If you apply the most significant bit to an analog inverter and use that output for the MSB position of the R-2R ladder, the binary number applied to the ladder will be handled as a two's-complement number, going both positive and negative.

CONCLUSION

Digital communication techniques are allowing higher per formance, smaller, less expensive instruments and devices to be developed for all aspects of wireless communication. Hardware that applies digital communication techniques can provide a powerful and flexible platform for custom communication signal acquisition, processing, and control. ZTEC Inc.

The modulator implementation technique describe is versatile .The pulse shape, and consequently, the spectral shape of the resulting signal can be changed simply changing the contents of a memory ship. Any source s ymbol rate ($R_{s(\max)}$) can be accommodated by simply changing the master clock rate .Simple content modification of the memory ship in the sample generator and the coefficient resolver would result in a new modulation format.

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