Chapter 2
Evolution of 2.4 GHz Wireless LANs

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This chapter considers the recently successful IEEE 802.11b standard for high performance wireless Ethernet and a proposed extension that provides for 22 Mbps transmission. The IEEE 802.11 sets standards for wireless Ethernet or wireless local area networks. The chapter describes the history of the IEEE 802.11 standards and the market opportunities in the wireless Ethernet field. The chapter gives a brief description of the media access control (MAC) layer and then presents details about the physical layer methods, including coding descriptions and performance evaluations. The chapter also discusses the role and limitations of spread spectrum communications in wireless Ethernet. A comparison in terms of range versus rate with several alternatives is presented.

2.1 INTRODUCTION TO WIRELESS ETHERNET

In the Fall of 1999 a new high speed standard for wireless Ethernet was ratified by the IEEE 802.11 standards body [1]. This standard extended the original 1 & 2 mega-bit-per-second (Mbps) direct sequence physical layer transmission standard, [2], to break the 10 Mbps barrier. The standard, "IEEE 802.11b," established two forms of coding that

each deliver both a 5.5 Mbps and 11 Mbps data rate.

The second optional choice of coding is known as *packet binary convolu-*

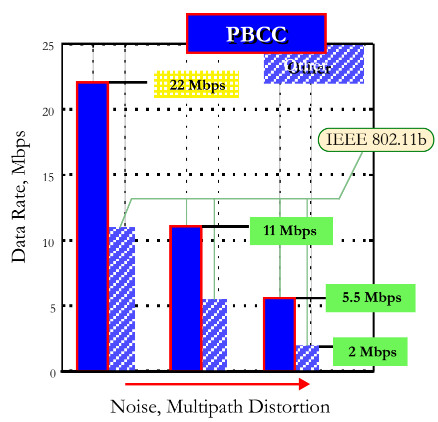

Figure 2.1 Performance Wireless Ethernet

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tional coding or “PBCC”. This PBCC option was developed by Alantro Communications, now a part of Texas Instruments Inc. This chapter describes the evolution of the standards for the 2.4GHz ISM band and the extensions developed by Alantro Communications. The Alantro PBCC system maintains a 22 Mbps data rate in the same environment as the basic 11 Mbps system of the current IEEE 802.11b standard as schematically described in Figure 2.1. The system provides for a backword compatible migration of 802.11b networks into the realm of higher throughput. This chapter greatly expands and augments the material found in [3].

2.1.1 The History and State of the Standards and Marketplace

The origins of wireless networking standardization can be traced to the late 1980’s when members of the IEEE 802.4 standards body considered extensions of token bus technology to wireless transmission. This activity was motivated by FCC spread spectrum regulations that provided for unlicensed transmission in an 83 MHz band of radio frequencies in the 2.4 GHz range. Although this activity did not produce a standard, the interest in these developments lead to the creation of IEEE 802.11 in May 1989. The charter for this group is the creation of internationally applicable standards for wireless Ethernet.

The initial standards activity was very contentious and progress was slow. In addition, as is often the case with good ideas, the technology available for the creation of robust, high performance/low cost solutions was not mature. In October of 1997, the first completed standard from the IEEE 802.11 body was ratified. Although the effort to develop the standard was tortuous and time-consuming, the results are impressive. The standard set in 1997 defined both a common *media access control* (MAC) mechanism as well as three *physical access methods* (PHYs). The three PHYs involved two radio transmission methods for the 2.4 GHz band: *frequency hopping* (FH) and *direct sequence spread spectrum* (DSSS). Both of these PHYs operated as a 1 & 2 Mbps data rate and have been deployed in products that were sold on the open market. (The third IEEE 802.11 PHY is an *infra-red* (IR) scheme; it is unclear whether any products have been produced with this technology.)

As the first standard was wrapping up, the creation of a new standards activity in IEEE 802.11 was begun. The motivation was to improve the physical layer specification to improve the data rate and throughput parameters of wireless Ethernet. There was strong consensus in the group that wireless Ethernet must be able to deliver a data rate that is comparable to the data rate offered by traditional Ethernet, 10 Mbps. It was also agreed that the new activity would concentrate on the physical layer and that changes to the common IEEE 802.11 MAC would be limited to the additions required to make the MAC aware of the parameters of the new PHY technology.

This new activity consisted of two initiatives. The first group considered the definition of a PHY for the unlicensed 5 GHz bands. This effort resulted in the IEEE 802.11a PHY for the 5 GHz band; this standard incorporates a coded multi-carrier scheme known as OFDM. The second effort produced a standard commonly known as the IEEE 802.11b standard. This standard offers a DSSS backword compatible transmission definition that added two new data rates, 5.5 Mbps and 11 Mbps, as well as two forms of coding. The mandatory coding mode is known as “CCK” modulation and is described in detail in Section 2.3.3.1 of this chapter. The optional code, known as “PBCC” and referred to as the “high performance mode” of the standard, is described in Section 2.3.3.2. This standard is clearly the most successful standard of the IEEE 802.11 to date; today there are millions of “11b” compliant devices in the hands of consumers.

Recently, the main standards setting activities of the IEEE 802.11 committee involve enhancements to the MAC, “11e,” and even higher rate extensions to the existing standard, “11g.” The former activities are directed towards enhancing the MAC, most importantly to improve *quality of service* (QOS) and security. The latter activity was motivated by the work of Alantro Communications which is a central topic of this chapter (see Sections 2.3.4.1 & 2.4). The main objective of this activity is to define a backword compatible extension to the existing “11b” networks in a way that improves the data rate (>20 Mbps) and overall user experience and satisfaction with wireless Ethernet.

As organizations such as the IEEE 802 Committee continue to push the envelope on the technology front, other organizations are also playing a key role in the adoption of Wireless Ethernet technology. The *Wireless Ethernet Compatibility Alliance* (WECA) is the most notable such organization. Both the IEEE and WECA have been instrumental in advocating innovation and enhancements to the standard, which has helped fuel rapid industry adoption. WECA’s mission is to certify inter-operability of 802.11b™(IEEE 802.11) products and to promote 802.11b as the global wireless *local area network* (LAN) standard across all market segments. The alliance recently announced that 67 products have passed the rigorous 802.11b certification testing; this makes 802.11b the world’s leading wireless LAN standard. Furthermore, momentum continues growing as WECA attracts new members from around the world.

Until a year ago there were several wireless LAN standards competing for the home market, however, 802.11b has resolved this issue. In less than a year, 802.11b has become the single wireless LAN standard for the home, small business, enterprise and public access areas.

2.1.2 Commercial Opportunities

The wide-scale availability of broadband to many homes and most businesses is accelerating the demand for wireless Ethernet. Now that users have easy access to these high-speed communications pipes, they are searching for a simple and cost-effective way to fully utilize them. In homes, a residential wireless gateway can interconnect desktop PCs, telephones, PDAs and other devices with 802.11b based wireless Ethernet. Soon, entertainment appliances like televisions, stereos and home theater systems will also be easily connected through this gateway. In the enterprise, users today are able to roam throughout their facilities while maintaining a wireless connection to the organization's network and servers.

As operators continue to roll out broadband services, they face a challenge with many customers. While bringing high bandwidth to the doorstep isn't the hurdle it once was, finding ways to effectively distribute that bandwidth once it crosses the demarcation point poses a mystery to some residential consumers. This broadband access distribution problem impacts small and medium size businesses as well. Solving this challenge has the potential to create tremendous market opportunity for communication services companies.

As home networking has gained momentum among consumers, communication companies have faced the challenge of installing new wires in their customers' homes. For example, many older homes have been particularly hard pressed to accommodate traditional Ethernet or LAN wiring; more often than not, the cost of installing it has been prohibitive. Even if it is physically feasible to re-wire an existing structure, installing new cabling has meant disruptions and lost productivity in the workplace or at home, in addition to being a major expense.

Recently deployed home networking technology, such as home phone networking, suffers from low user acceptance due to the inconvenience of the technology. It is often the case that the existing phone outlets installed in the home do not match the desired locations for the networked equipment. There are also conflicts in the use of the existing phone wires as the popularity of broadband access methods such as ADSL become more popular.

However, there is a more attractive solution – one that is rapidly gaining acceptance: *wireless Ethernet*. The challenge for communication service companies is to offer the best broadband distribution products to their users. Wireless networking systems are rapidly becoming a more and more affordable and the preferred choice for consumers. Recent developments surrounding a proposed performance extension to the Wireless Ethernet specification (IEEE 802.11b) hold great promise for an alternative to traditional wired networking. In fact, as the per-user cost of *Wireless LANs* is anticipated to drop sharply over the coming years, the market is likewise expected to explode, growing from $624 million in 1999 to $3 billion by 2002, according to Cahners/In-Stat [4].

For communication service companies, all of these performance improvements mean more robust wireless Ethernet installations. High data rates will not only accommodate today's most demanding applications, such as graphically-intense interactive gaming or high-definition television, but higher performance wireless Ethernet installed today will have the performance headroom it needs to accommodate new, even more demanding applications that have yet to be invented. A high-performance wireless Ethernet has the inherent scalability it will need to meet escalating application requirements for years to come.

Advanced technologies have expanded the effective operational range of 802.11b wireless LANs. Users have greater freedom to roam an environment and still be assured that their wireless device will be able to maintain a connection to the network.

This can be extremely important for users of all sorts of devices, such as notebook computers, PDAs or even wireless bar-code readers that are used frequently in warehouses or retail locations for inventory management. Highly efficient wireless Ethernet technology promises to make effective use of these broadband pipes, in addition to being an enabler of new and exciting applications. Multimedia applications like high-definition digital streaming video, cordless VoIP telephony, music distribution, connected always-on PDAs and other appliances are concepts that are just now beginning to tap into the potential that lies beneath the surface of wireless networking technology. Innovation, which has led to the availability of these high-performance, next-generation Wireless Ethernet products, is fulfilling the promise of broadband communication for consumers.

2.2 WIRELESS ETHERNET BACKGROUND

The IEEE 802.11 wireless LAN standard, commonly referred to as "wireless Ethernet," is part of a family of IEEE local and metropolitan networking standards, of which 802.3 ("Ethernet") and 802.5 (Token Ring Local Area Network) are two well-known, widely deployed examples. The IEEE 802 standards deal with the Physical and Data Link layers in the ISO *Open Systems Interconnection* (OSI) Basic Reference Model. IEEE 802 specifies the Data Link Layer in two sub-layers, *Logical Link Control* (LLC) and *Medium Access Control* (MAC). The IEEE 802 LAN MACs share a common LLC layer (IEEE standard 802.2) and Link Layer address space utilizing 48-bit addresses.

It is relatively straightforward to bridge between IEEE 802.11 wireless LANs and IEEE 802 wired LANs and to construct extended interconnected wired and wireless 802 LAN networks. Through this means all the services typically offered on wired LANs, such as file sharing, email transfer, and internet browsing, are made available

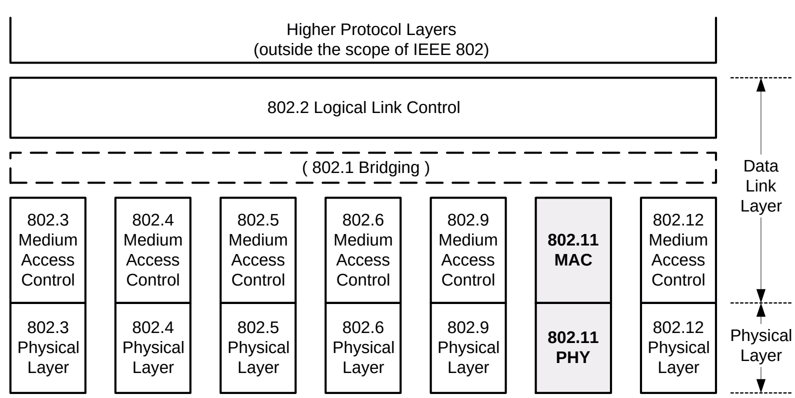


Figure 2.2 IEEE 802 Standards

to wireless stations. Transparent un-tethered LAN connectivity, high data rates (currently 11 Mbps and increasing to 22 Mbps as described in this chapter), acceptable cost, as well as the inherent inter-operability afforded by an international standard, are contributing factors to the rapidly increasing popularity of 802.11b wireless LANs.

2.2.1 Wireless Ethernet Topology

Wireless Ethernet Topology IEEE 802.11 mobile stations (end user client stations) may be mobile, portable, or stationary. Mobile stations dynamically associate with wireless LAN cells, or *Basic Service Sets* (BSSs). The 802.11 MAC protocol supports the formation of two distinct types of BSSs.

The first is the independent BSS, or "ad-hoc" BSS. Ad-hoc BSSs are typically self-forming; they are created and maintained as needed without prior administrative arrangements, often for specific purposes (such as transferring a file from one personal computer to another). Stations in an ad-hoc BSS establish MAC layer wireless links with those stations in the BSS with which they desire to communicate, and frames are transferred directly from source to destination stations. Therefore, stations in an ad-hoc BSS must be within range of one another in order to communicate. Furthermore, no architectural provisions are made for connecting the ad-hoc BSS to external networks, so communications is limited to the stations within the ad-hoc BSS.

The second type of BSS is the infrastructure BSS; this is the more common type used in practice. This type supports extended interconnected wireless and wired networking. Within each infrastructure BSS is an *Access Point* (AP), a special central traffic relay station that normally operates on a fixed channel and is stationary. APs connect the infrastructure BSS to an IEEE 802.11 abstraction known as the *Distribution System* (DS). Multiple APs connected to a common DS form an *Extended Service Set* (ESS). The IEEE 802.11 standard portal function connects the DS to non-802.11 LANs, and ultimately to the rest of the network system if present. The DS is responsible for forwarding frames within the ESS, between APs and portals, and it may be implemented with wired or wireless links.

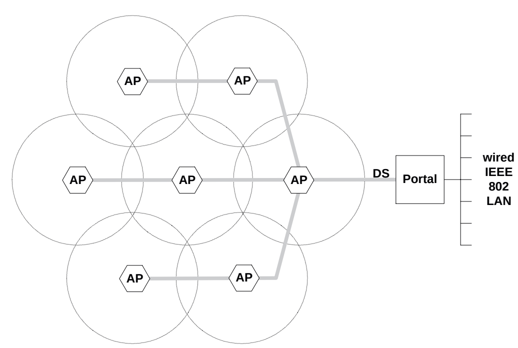


Figure 2.3 IEEE 802.11 Network

The ESS allows wireless LAN connectivity to be offered over an extended area, such as a large campus environment. APs may be placed such that the BSSs they service overlap slightly in order to provide continuous coverage to mobile stations.

In practice Distribution Systems are typically implemented using ordinary wired Ethernet. Commercially available APs include an embedded Ethernet portal, and they are therefore essentially wireless LAN to Ethernet bridges.

Mobile stations in an infrastructure BSS establish MAC layer links with an AP. Furthermore, they only communicate directly to and from the selected AP. The AP / DS utilizes store-and-forward retransmission for intra-BSS traffic in order to provide connectivity between the mobile stations in a BSS. Typically, at most a small fraction of the frames flow between mobile stations within an infrastructure BSS, therefore retransmission results in a small overall bandwidth penalty. The effective physical span of the BSS is on the order of twice the maximum mobile station to station range; mobile stations must be within range of the AP to join a BSS, but may not be within range of all other mobile stations in the BSS.

Mobile stations utilize the 802.11 architected scan, authentication, and association processes in order to join an infrastructure BSS and connect to the wireless LAN system. Scanning allows mobile stations to discover existing BSSs that are within range. APs periodically transmit beacon frames that, among other things, may be used by mobile stations to discover BSSs. Prior to joining a BSS, a mobile station must demonstrate through authentication that it has the credentials to do so. The actual BSS join occurs through association. Mobile stations may authenticate with multiple APs, but may be associated with only one AP at a time. Roaming mobile stations initiate handoff from one BSS to another through reassociation. The reassociation management frame is both a request by the sending mobile station to disassociate from the currently associated BSS, and a request to join a new BSS.

2.2.2 Medium Access Control

The IEEE 802.11 MAC is similar to wired Ethernet in that both utilize a "listen before talk" mechanism to control access to a shared medium. However, the wireless medium presents some unique challenges not present in wired LANs that must be dealt with by the 802.11 MAC. The wireless medium is subject to interference and is inherently less reliable. The medium is susceptible to possible unwanted interception. Wireless networks suffer from the "hidden station" problem; a station transmitting to a receiving station may be interfered with by a third "hidden" station which is within range of the receiver but out of range of the transmitter and therefore does not defer. Finally, wireless stations cannot reliably monitor the idle / busy state of the medium while transmitting. The 802.11 MAC protocol is designed to provide robust, secure communications over the wireless medium.

The fundamental access mechanism is *Carrier Sense Multiple Access / Collision Avoidance* (CSMA/CA) with truncated binary exponential back off. A station with a frame to transmit contends for the medium by first sensing the medium and deferring until it is idle for a minimum period of time, at which point the station transmits a frame. If the frame is a unicast frame and is received without error by the destination station, the destination station immediately returns a positive acknowledgement frame. If the originating station does not successfully receive the acknowledgement frame, the station assumes that a collision or other event producing a lost packet has occurred. In response to a lost packet, the transmitting station selects a random back off interval from a uniform range. The range is doubled for every lost packet experienced, until an administratively configurable maximum is reached. The transmitting station then re-queues the frame for transmission and contends for the medium after the back off interval has been satisfied. Multicast and broadcast frames do not use the acknowledgement protocol, and other mechanisms provide protection from lost packets for these frames.

Multiple MAC layer mechanisms contribute to collision avoidance and efficient use of the wireless medium. In contrast to wired Ethernet, if the medium is sensed busy for the first transmission attempt a random back off is selected and applied. In addition, the back off counters in deferring 802.11 stations are not decremented when the medium is sensed busy. These two mechanisms reduce the probability of contention when it is most likely to occur, immediately following a transmission.

In addition to the basic contention access mechanism described above, IEEE 802.11 offers an optional contention-free access method. Contention-free access is available only in infrastructure BSSs. Currently, contention-free access is not commonly utilized. However, contention-free access is expected to play an increasingly important role in the future for implementation of quality of service.

With contention-free access, APs gain and maintain control of the wireless medium for extended periods using virtual carrier sense and IFS timing hierarchy (described below). During contention-free periods polling by the AP is used to grant to mobile stations access to the medium.

The IEEE 802.11 MAC adheres to a strict *inter-frame space* (IFS) timing hierarchy; four different IFS durations are specified, separated by a minimum of one slot time. These IFS durations establish the length of the gap between non-deferred transmissions, both for frame burst from a single station, and for listen-then-talk transmissions. Due to the listen-then-talk access method, transmissions utilizing a given IFS preempt, without contention, those queued transmissions using a longer IFS.

Two types of inter-frame spacings, the SIFS and the PIFS, are applied when normally only one station in the BSS has permission to transmit, and are therefore intended to result in contention-free access. The *short inter-frame spacing* (SIFS) is the smallest IFS and it is used betweeen certain multi-frame exchange sequences, such as acknowledgement frames sent in response to the error-free reception of a unicast frame. The remaining IFS intervals in order of increasing duration are the DIFS, used by APs to gain priority access to the medium, the PIFS, used by contending stations whose back off interval has been satisfied, and the EIFS, an IFS enforced after an erroneous reception.

Virtual carrier sense is a MAC layer mechanism that augments the physical carrier sense generated by the PHY layer. The duration / ID field in the MAC frame header indicates the expected time remaining to complete the current frame exchange sequence. Stations defer based upon previously received duration values, even if the physical carrier sense indicates the medium is idle. Virtual carrier sense mitigates the hidden station problem. For example, virtual carrier sense prevents a station that is within range of a transmitting station, but out of range of the destination station, from colliding with the acknowledgement frame returned by the destination station.

Virtual carrier sense together with the *request to send / clear to send* (RTS/CTS) protocol allows stations to place a reservation on the medium prior to transmitting a data frame. Because RTS and CTS are short control frames and therefore occupy the medium for a relatively short time, the RTS / CTS protocol increases the probability of successful transmission and reduces loss of network throughput due to collisions.

Fragmentation and Automatic Rate Fall-back improve the robustness of 802.11 networks by increasing the probability of successful packet transfer, especially for networks that experience time-varying external interference. Long transmit packets presented to the 802.11 MAC by upper protocol layers may be fragmented by the MAC into smaller packets for transmission on the medium. The smaller fragmented packets occupy the medium for shorter periods, increasing the probability that they will be successfully received. Receiving stations re-assemble fragments in order to regenerate the original transmit packet for the upper protocol layers. 802.11 MAC fragmentation is transparent to upper protocol layers.

The 802.11 wireless PHYs provide for multiple transmission bit rates. Generally speaking, lower bit rates enjoy greater range and decreased susceptibility to interference. With automatic rate fall-back 802.11 MACs automatically dynamically choose target station specific transmission rates based upon packet loss statistics and receive signal quality indications provided by the PHY in order to optimize throughput.

2.2.3 Security

Wireless LANs are subject to possible unwanted monitoring. For this reason IEEE 802.11 specifies an optional MAC layer security system known as *wired equivalent privacy* (WEP). As the name implies, WEP is intended to provide to the wireless Ethernet a level of privacy similar to that enjoyed by wired Ethernets. WEP involves a shared key authentication service with RC4 encryption. The RC4[1] is stream cipher is used to generate a pseudo-random sequence that is "XOR-ed" into the data stream (*ala* a "one time pad"). A key, derived by combining a secret key and an *initialization vector* (IV), is used to set the initial condition or state of the RC4 pseudo-random number generator. By default each BSS supports up to four 40-bit keys that are shared by all the stations in the BSS. Keys unique to a pair of communicating stations and direction of transmission may also be used (that is, unique to a transmit / receive address pair). Key distribution is outside the scope of the standard but presumably utilizes a secure mechanism.

[1]RC4 is a stream cipher designed by Ronald Rivest for RSA Data Security (now RSA Security). It is commonly known as "Ron's cipher 4".

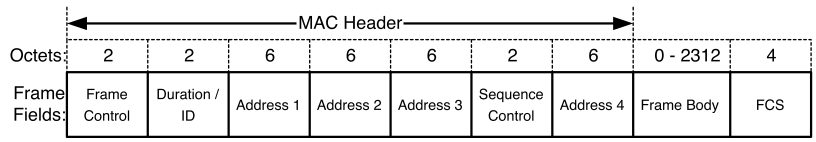


Figure 2.4 Wireless Ethernet Frame

When a station attempts to authenticate with a second station that implements WEP, the authenticating station presents to the requesting station challenge text. The requesting station encrypts the challenge text using the RC4 algorithm and returns the encrypted text to the authenticating station. The encrypted challenge text is decrypted and checked by the authenticating station prior to completing authentication. After authentication (and association), the Frame Body (the MAC payload) is encrypted in all frames exchanged between the stations. Encrypted frames are decrypted and checked by the MAC layer of receiving stations before being passed to the upper protocol layers.

2.2.4 The 802.11 MAC Frame Format

Shown in Figure 2.4 is the general 802.11 MAC frame format. (Not shown is the PHY header that is appended to the front of every frame transmission; see: Figure 2.5.) The Address 2, 3 and 4, the Sequence Control, and the Frame Body fields are not found in every frame. The frame control field is 16 bits in length and it contains basic frame control information, including the frame type (data, MAC control, or MAC management) and subtype, if the frame is originated from or is bound to the DS, and if the frame is encrypted. The duration / ID field normally indicates the duration of the remainder of a frame exchange sequence and is used to control the virtual carrier sense mechanism as previously described.

The address fields, if present, contain one of the following 48 bit IEEE 802 Link Layer addresses: Destination Address, Source Address, Receiver Address, Transmitter address, *Basic Service Set ID* (BSSID). For infrastructure networks, the BSSID is the Link Layer address of the AP. For ad-hoc networks, the BSSID is a random number generated at the time the ad-hoc network is formed. The Receiver, Transmitter, and BSSID addresses are the MAC addresses of stations joined to the BSS that are transmitting or receiving the frame over the wireless Ethernet. Destination and Source addresses are the MAC addresses of stations, wireless or otherwise, that are the ultimate destination and source of the frame. In those cases where two addresses are the same (for example, the Receiver station and the Destination station are one and the same), then a single address field is used. Four address fields are present only in the uncommon case where the DS is implemented with an 802.11 wireless Ethernet, and only for frames traversing the DS. A more typical case involves a frame originating from a wireless station in an infrastructure BSS that is bound for a station on a wired network such as an IEEE 802.3 wired Ethernet. In this situation, the Address 1 field contains the BSSID, the Address 2 field contains the address of the source / transmitter station, the Address 3 field contains the address of the destination station, and the Address 4 field is not present. Including both the BSSID and the Destination Address (or Source Address for frames flowing to the BSS) in the frame avoids requiring the AP to maintain a list of MAC addresses of stations that are not in the BSS.

The Sequence Control field is 16 bits in length and it contains the Sequence Number and Fragment Number sub-fields. Receiving stations use this field to properly reassemble multi-fragment frames and to identify and discard duplicate frame fragments.

The Frame Body is an optional field that contains the MAC frame payload. For 802.11 MAC management type frames, the Frame Body contains information elements that are specific to the subtype. The FCS field contains a 32 bit *Cyclic Redundancy Check* (CRC). The CRC calculation includes all the MAC frame fields.

2.3 THE PHYSICAL LAYER: CODING AND MODULATION

2.3.1 The Physical Layer Preamble

The IEEE 802.11b standard defines a physical layer (PHY) preamble that is transmitted before the wireless Ethernet frame depicted in Figure 2.4. The PHY preamble, as shown in Figure 2.5, consists of a preamble and a header. The header consists of three fields, the *Signal* field, the *Service* field and the *Length* field. These three fields are protected with a 16 bit CRC that is used to detect transmission errors in the header.

The PHY preamble provides for—

- ❖ Packet Detection and Training:
  The preamble is used to detect the presence of a packet transmission, to decide on antenna selection, and to estimate packet parameters such as signal level for automatic gain control (AGC), carrier offset for frequency tracking, symbol timing, etc.

- ❖ Detection of Frame Boundary (SFD):
  For packet frame synchronization.

- ❖ Description of Packet Body Modulation and Coding:
  The choice of coding and modulation is described by the Signal *field.*

- ❖ Virtual Carrier Sense:
  The Length *field describes the length of transmission for the body of the packet. This Length field measures the transmission in time duration (rather than bits); it is used to initialize a timer in each receiver that detects the packet and is used to time the transmission period. This allows unintended receivers, that*

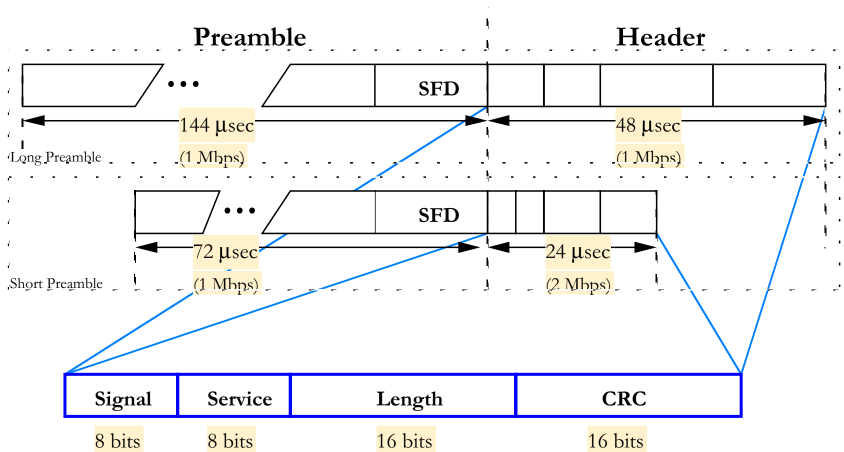


Figure 2.5 The Physical Layer Preamble

> *may be incapable of demodulating/decoding the type of packet specified in the Signal field, to refrain from transmission during the duration of the packet. This mechanism avoids packet collisions and allows for the introduction of new forms of coding and modulation, in an existing network, in a backword compatible way.*

The original DSSS (1 & 2 Mbps) standard defined a PHY preamble with a length of 192 μsecs; this preamble is encoded using the 1 Mbps encoding method described in Section 2.3.2.1. The "11b" standard added an optional "short preamble" with a duration that is half as long, 96 μsecs. The short preamble uses a shorter, 1 Mbps encoded preamble, followed by a 2 Mbps encoded header.

2.3.2 The Low Rate DS Standards: *The Past*

The original low rate *direct sequence* (DS) modulation forms a basis for the high rate extension. This method of coding and modulation is used for preamble generation in all rates and coding combinations. The low rate system is a direct sequence spread spectrum signal with a "chip rate" of 11 MHz and a data rate of 1 Mbps (BPSK) or 2 Mbps (QPSK).

2.3.2.1 Barker 1 & 2 Mbps The basis for the original 1 and 2 Mbps transmission is the incorporation of an 11 bit Barker code (or sequence)

$$B_{11} = [-1, +1, -1, -1, +1, -1, -1, -1, +1, +1, +1]$$

with QPSK or BPSK modulation. This code has the desirable property that the auto-correlation function is minimal (0 or -1) except at the origin (where it is 11) as seen in Figure 2.14 in Section 2.4.1. This means the modulated waveform essentially occupies the same spectrum (see 2.4.1) as an 11 MHz uncoded chip signal and that a matched filter receiver, matched to the Barker sequence, will experience a processing gain of 11 = 10.41 dB.

Table 2.1 QPSK Mapping (CCK)

Code Symbol c_i	Signal x_i
0	+1+i
1	-1+i
2	-1-i
3	+1-i

From a coding point of view, the Barker code can be described in terms of a *linear block code* over the set of integers modulo 4, $\mathcal{Z}_4 = \{0, 1, 2, 3\}$. Consider the $k \times n = 1 \times 11$ repetition *generator matrix*

$$G = [1, 1, 1, 1, 1, 1, 1, 1, 1, 1, 1]$$

and the length 11 *cover vector*

$$\mathbf{b} = [2, 0, 2, 2, 0, 2, 2, 2, 0, 0, 0].$$

Then the four Barker codewords for the 2 Mbps case are generated by the codeword equation

$$\mathbf{c} = m \cdot G + \mathbf{b} = [c_1, c_2, c_3, c_4, c_5, c_6, c_7, c_8, c_9, c_{10}, c_{11}] \quad (\text{modulo } 4) \tag{2.1}$$

where the message symbol $m \in \mathcal{Z}_4$. The transmitted signal is generated with the QPSK mapping in Table 2.1 which produces the signal vector

$$\mathbf{c} \to \mathbf{x} = [x_1, x_2, x_3, x_4, x_5, x_6, x_7, x_8, x_9, x_{10}, x_{11}]. \tag{2.2}$$

The complex values in Table 2.1 are used to represent the “in-phase” (the real part or “cosine”) and “quadrature-phase” (the imaginary part or “sine”) of the pulse amplitude modulated carrier. For example, the A_k’s in Equation (2.6), on page 54, would take on the 4 complex QPSK values given in Table 2.1. The resulting complex baseband signal $x(t)$ would be modulated to a a prescribed carrier frequency in the 2.4-2.483 GHz range. The transmitter would “mix” (*i.e.*, multiply) the real part of $x(t)$ by a cosine wave at the carrier frequency and mix the imaginary part of $x(t)$ by a sine wave (the cosine wave shifted by $90°$) and add the two mixed signals together.

This composite signal would be a bandpass signaled centered at the carrier frequency as described in [5].

Notice that the 2Mbps Barker code is 90° rotationally invariant (*i.e.*, the rotation of a codeword vector $\mathbf{x}$ by 90° is another codeword). This follows from the fact the addition of 1 (modulo 4) to a message symbol $m \in \mathcal{Z}_4$ will add the all 1's vector (modulo 4) to the codeword $\mathbf{c}$ and that incrementing by 1 (modulo 4) in the QPSK mapping (Table 2.1) corresponds to rotation by 90° (counter-clockwise). This rotational invariance is exploited in the standard by using a differential encoding method that involves "precoding" at the transmitter[2]

$$\tilde{m}_k = m_k + \tilde{m}_{k-1} \quad (\text{modulo } 4)$$

and "differential" decoding at the receiver

$$m_k = \tilde{m}_k - \tilde{m}_{k-1} \quad (\text{modulo } 4)$$

(the sliding window nature of the differential decoder limits error propagation).

The 1 Mbps mode is defined by using a repetition generator matrix

$$G = [2, 2, 2, 2, 2, 2, 2, 2, 2, 2, 2]$$

which incorporates a binary message symbol, $m \in \mathcal{Z}_2 = \{0, 1\}$ and produces a BPSK signal, $x_j \in \{+1 + i, -1 - i\}$. This produces a code that is 180° rotationally invariant.

Example *Barker*
To encode $[m_1 = 1, m_2 = 0, m_3 = 2, m_4 = 3, \dots]$ the precoder would produce (taking $\tilde{m}_0 = 0$)

$$[\tilde{m}_1 = 1, \tilde{m}_2 = 1, \tilde{m}_3 = 3, \tilde{m}_4 = 2, \dots].$$

This would be encoded into the barker codewords according to Equation (2.1)

$$[31331333111, 31331333111, 13113111333, 02002000222, \dots]$$

and then translated to the QPSK symbols as in Equation (2.2)

$$\begin{gathered}
[1 - i, -1 + i, 1 - i, 1 - i, -1 + i, 1 - i, 1 - i, 1 - i, -1 + i, -1 + i, -1 + i, \\
1 - i, -1 + i, 1 - i, 1 - i, -1 + i, 1 - i, 1 - i, 1 - i, -1 + i, -1 + i, -1 + i, \\
-1 + i, 1 - i, -1 + i, -1 + i, 1 - i, -1 + i, -1 + i, -1 + i, 1 - i, 1 - i, 1 - i, \\
1 + i, -1 - i, 1 + i, 1 + i, -1 - i, 1 + i, 1 + i, 1 + i, -1 - i, -1 - i, -1 - i, \\
\dots]
\end{gathered}$$

[2] The precoded symbol at time k, $\tilde{m}_k$, is used in the encoding Equation (2.1)

according to Table 2.1. ❒

In any communications system, the reliability of transmission can be improved with a correspond reduction in transmission rate. For example, by sending a given signal n times, an energy gain factor of n (or $10 \cdot \log_{10}(n)$ dB) can be achieved in signal to noise ratio (SNR) at the expense of a factor of $1/n$ in rate (since the same signal is transmitted n times). However, coding theory predicts that for a given reduction in rate R, the improvement in SNR can be greater that $1/R$. An improvement in excess of the simple "repeation gain" is commonly known as "coding gain".

The *minimum squared distance* (MSD) of QPSK is $2\mathrm{E_s}$ (where $\mathrm{E_s}$ is the average symbol energy) while BPSK has an MSD of $4\mathrm{E_s}$. Both the 1 & 2 Mbps transmissions schemes show an energy improvement in minimum distance squared, at the cost of rate. In the case of 2 Mbps, the minimum distance squared is $22\mathrm{E_s}$ since the Barker encoder has a repeation effect of length 11. This results in an energy gain of 11 = 10.41 dB over uncoded QPSK with a factor of 1/11 in the data rate. From a coding gain perspective, there is no coding gain *w.r.t.* QPSK since the the minimum distance squared normalized by the data rate is the same as QPSK. The *asymptotic coding gain* (ACG) of a coded system (C) relative to an uncoded system (U) is defined as the ratio

$$\mathrm{ACG} = \frac{\mathrm{d}^2_{\min}(C) \cdot R(C)/\mathrm{E_s}(C)}{\mathrm{d}^2_{\min}(U) \cdot R(U)/\mathrm{E_s}(U)}.$$

In the 2 Mbps case, $\mathrm{d}^2_{\min}(C)/\mathrm{E_s}(C) = 22$ and $R(C) = 2/11$ (bits/symbol), while for uncoded QPSK, $\mathrm{d}^2_{\min}(U)/\mathrm{E_s}(U) = 2$ and $R(U) = 2$; in this case $\mathrm{ACG} = 1 =$ 0 dB. Similarly in the 1 Mbps case, which uses BPSK, there is an energy gain of $2 \cdot 11 = 22 = 13.42$ dB (over QPSK) but 0 dB of coding gain since the data rate factor is 1/22 of uncoded QPSK.

2.3.3 The "High Rate" Standards: *The Present*

The standard calls for two choices of coding each involving a "symbol rate" of 11 MHz and data rates of 5.5 Mbps and 11 Mbps. One code uses a short blocklength code, known as "CCK" that codes over 8 QPSK symbols and the other choice incorporates a 64 state, packet based binary convolutional code (PBCC). The main difference between the two involves the much larger coding gain of the PBCC over CCK at a cost of computation at the receiver.

2.3.3.1 CCK 5.5 & 11 Mbps The *complementary code keying* (CCK) code can be considered as a block code generalization of the low rate Barker code. For CCK-11, the code is an $(n = 8, k = 4)$ linear block code over $\mathcal{Z}_4$. At the 11 Mbps rate, 8 bits (4-$\mathcal{Z}_4$ symbols) of information is encoded via the $k \times n = 4 \times 8$ CCK-11 *generator* *matrix*

$$G = \begin{bmatrix} 1 & 1 & 1 & 1 & 1 & 1 & 1 & 1 \\ 1 & 1 & 1 & 1 & 0 & 0 & 0 & 0 \\ 1 & 1 & 0 & 0 & 1 & 1 & 0 & 0 \\ 1 & 0 & 1 & 0 & 1 & 0 & 1 & 0 \end{bmatrix}$$

using the matrix equation

$$\mathbf{c} = \mathbf{m} \cdot G + \mathbf{b} = [c_1, c_2, c_3, c_4, c_5, c_6, c_7, c_8] \quad (\text{modulo } 4). \tag{2.3}$$

In this case, the length 8 *cover vector* is given by

$$\mathbf{b} = [0, 0, 0, 2, 0, 0, 2, 0]$$

and the message vector, $\mathbf{m} = [m_1, m_2, m_3, m_4]$, $m_j \in \mathcal{Z}_4$, represents 8 bits of information. Applying the QPSK mapping, shown in Table 2.1, produces the signal vector

$$\mathbf{c} \to \mathbf{x} = [x_1, x_2, x_3, x_4, x_5, x_6, x_7, x_8].$$

At the 5.5 Mbps rate, 4 bits of information is encoded via the $k \times n = 3 \times 8$ CCK-5.5 *generator matrix*

$$G = \begin{bmatrix} 1 & 1 & 1 & 1 & 1 & 1 & 1 & 1 \\ 2 & 2 & 2 & 2 & 0 & 0 & 0 & 0 \\ 2 & 0 & 2 & 0 & 2 & 0 & 2 & 0 \end{bmatrix}$$

using the matrix equation

$$\mathbf{c} = \mathbf{m} \cdot G + \mathbf{b} = [c_1, c_2, c_3, c_4, c_5, c_6, c_7, c_8] \quad (\text{modulo } 4).$$

In this case, the length 8 *cover vector* is given by

$$\mathbf{b} = [1, 0, 1, 2, 1, 0, 3, 0]$$

and the message vector, $\mathbf{m} = [m_1, m_2, m_3]$, $m_1 \in \mathcal{Z}_4, m_2 \in \mathcal{Z}_2, m_3 \in \mathcal{Z}_2$, represents 4 bits of information.

The CCK code is rotationally invariant since the first row of the generator matrix G is the all 1's vector. This implies that a rotation by a multiple of $90°$ at the receiver will affect only the first symbol m_1 of the message vector. This is exploited in the standard by differential encoding/decoding on the first symbol m_1, using the same method as in the low rate case.

$\boxed{\textit{Example}}$ *CCK-11*

The encoding of the 8 bits $[m_1 = 1, m_2 = 0, m_3 = 2, m_4 = 3]$ (ignoring the precoding function on m_1) produces the CCK codeword according to Equation (2.3)

$$\begin{aligned}\mathbf{c} = \; & 1\cdot[11111111] + 0\cdot[11110000] + 2\cdot[11001100] + 2\cdot[10101010] \\ & + [10121030] \\ = \; & [11111111] + [00000000] + [22002200] + [20202020] + [10121030] \\ = \; & [23032321]\end{aligned}$$

and then translated to the QPSK symbols as in Equation (2.2)

$$[-1-i, 1-i, 1+i, 1-i, -1-i, 1-i, -1-i, -1+i]$$

according Table 2.1. ❐

Table 2.2 CCK Weight Distribution

Wt/2E_s:	0	4	6	8	10	12	16
Number (CCK-11):	1	24	16	174	16	24	1
Number (CCK-5.5):	1			14			1
Number (CCK-6.875):	1			30			1

The minimum distance squared of the 11 Mbps CCK code is $8\mathrm{E_s}$; two codewords at minimum distance are generated by the messages $\mathbf{m}_1 = [0000]$

$$\begin{aligned}\mathbf{c}_1 &= \begin{bmatrix}0 & 0 & 0 & 2 & 0 & 0 & 2 & 0\end{bmatrix} \rightarrow \\ \mathbf{x}_1 &= \begin{bmatrix}+1+i & +1+i & +1+i & -1-i & +1+i & +1+i & -1-i & +1+i\end{bmatrix}\end{aligned}$$

and $\mathbf{m}_2 = [0001]$,

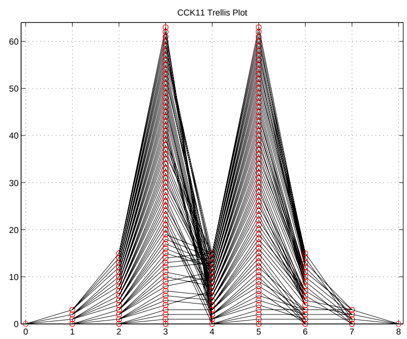
$$\begin{aligned}\mathbf{c}_1 &= \begin{bmatrix}1 & 0 & 1 & 2 & 1 & 0 & 3 & 0\end{bmatrix} \rightarrow \\ \mathbf{x}_1 &= \begin{bmatrix}-1+i & +1+i & -1+i & -1-i & -1+i & +1+i & +1-i & +1+i\end{bmatrix}\end{aligned}$$

for example. The minimum distance squared of the 5.5 Mbps CCK code is $16\mathrm{E_s}$. It is interesting to note that a 6.875 Mbps CCK code, with the same minimum distance of $16\mathrm{E_s}$, is possible by using the generator

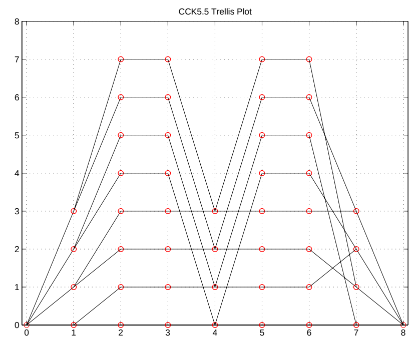
$$G = \begin{bmatrix}1 & 1 & 1 & 1 & 1 & 1 & 1 & 1 \\ 2 & 2 & 2 & 2 & 0 & 0 & 0 & 0 \\ 2 & 2 & 0 & 0 & 2 & 2 & 0 & 0 \\ 2 & 0 & 2 & 0 & 2 & 0 & 2 & 0\end{bmatrix};$$

this code is *not* part of the standard.

The asymptotic coding gain for CCK is 3 dB ($\mathrm{ACG} = 2$) over uncoded QPSK. This follows the fact that the code rate is 1/2 while the minimum distance is 4; the product



(a) CCK 11 Mbps Trellis



(b) CCK 5.5 Mbps Trellis

Figure 2.6 The Trellis for CCK

is 2. However, the practical coding gain is about 2 dB (as shown in Section 2.4.2 on page 56). The reduction in coding gain from the asymptote is due to the number of "nearest neighbors" at the minimum distance as shown in Table 2.2. This table shows that at the minimum distance of the code ($8\mathrm{E_s}$ for CCK-11 and $16\mathrm{E_s}$ for CCK-5.5/6.875) there are 24/14/30 codewords. This large number of nearest neighbors (compared to 2 nearest neighbors for the 2 Mbps Barker) accounts for the 1 dB reduction in practical coding gain.

Since the CCK codes are affine translations of linear block codes, the codewords can be compactly described in terms of a trellis with $n = 8$ sections as shown in Figure 2.6. A description of the generation of the trellis for a linear block code is given in the Chapter 2 of [6]. In the case of CCK-11, the number of states of the trellis follow a $[1, 4, 16, 64, 16, 64, 16, 4, 1]$ pattern; there are 296 branches in the trellis.

The trellis can be derived from a parity check matrix

$$H = \begin{bmatrix} 0 & 1 & 0 & 3 & 3 & 0 & 1 & 0 \\ 1 & 3 & 3 & 1 & 0 & 0 & 0 & 0 \\ 0 & 0 & 1 & 3 & 3 & 1 & 0 & 0 \\ 0 & 0 & 0 & 0 & 1 & 3 & 3 & 1 \end{bmatrix},$$

a 4×8 matrix over $\mathcal{Z}_4$. Note that on 1st, 2nd, 3rd and 5th trellis sections, there is 4-way branching and on the 4th, 6th, 7th and 8th trellis sections there is 1-way branching. The trellis for CCK-5.5 has a $[1, 4, 8, 8, 4, 8, 8, 4, 1]$ state pattern with 56 branches and 4-way branching on the 1st trellis section and 2-way branching on the

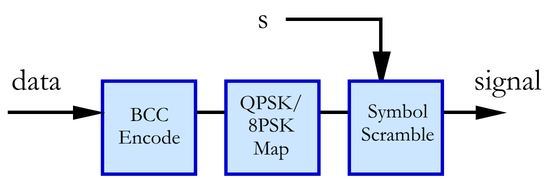


Figure 2.7 Packet Binary Convolutional Coding

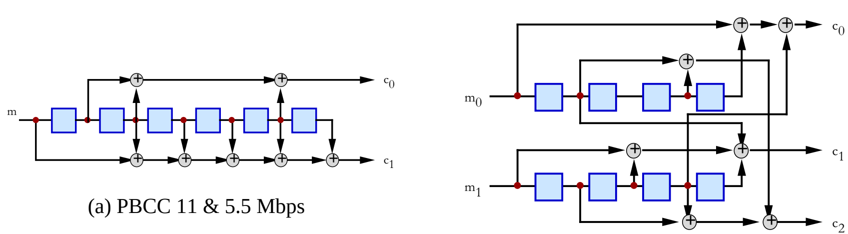
2nd and 5th. A parity check that generates this trellis is given by the 7×8 matrix

$$H = \begin{bmatrix} 1 & 0 & 3 & 0 & 0 & 0 & 0 & 0 \\ 0 & 1 & 0 & 3 & 0 & 0 & 0 & 0 \\ 0 & 0 & 1 & 3 & 3 & 1 & 0 & 0 \\ 0 & 0 & 0 & 0 & 1 & 0 & 3 & 0 \\ 0 & 0 & 0 & 0 & 0 & 1 & 0 & 3 \\ 0 & 0 & 1 & 1 & 1 & 1 & 0 & 0 \\ 1 & 2 & 1 & 0 & 0 & 0 & 0 & 0 \end{bmatrix}.$$

2.3.3.2 PBCC 5.5 & 11 Mbps The IEEE 802.11b standard specifies an optional choice of coding and modulation and is considered the "high performance" mode for 11 & 5.5 Mbps transmission. The optional mode, termed *packet binary convolutional coding* (PBCC), involves a BCC combined with a symbol scrambling method as shown in Figure 2.7. This structure is also used for the higher rate, 22 Mbps, encoding described in Section 2.3.4.1 on page 52.

The 802.11b PBCC mode (11Mbps & 5.5Mbps) uses a 1×2 generator matrix over $\mathcal{Z}_2[D]$, the set of polynomials (in the variable D) with binary coefficients:

$$G = \begin{bmatrix} D + D^2 + D^5 & 1 + D^2 + D^3 + D^4 + D^5 + D^6 \end{bmatrix} \tag{2.4}$$



(a) PBCC 11 & 5.5 Mbps

(b) PBCC 22 Mbps

Figure 2.8 The Binary Convolutional Encoders

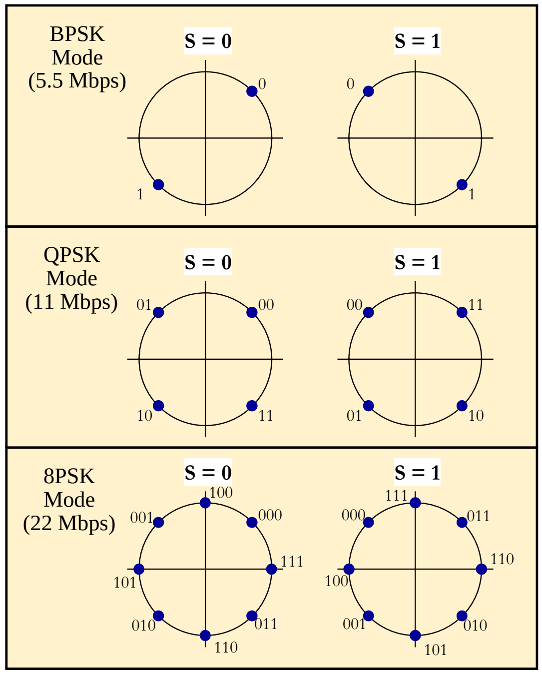


Figure 2.9 The Coding Symbol Scrambler

as shown in Figure 8(a) (in octal notation $G = [46, 175]$.) For 11 Mbps operation, this 64 state encoder is followed by a mapping onto QPSK modulation directly as specified in Table 2.3[3]. For 5.5 Mbps, the two binary outputs are bit serialized and mapped onto BPSK.

[3]The mapping given in Table 2.3 does not map "Hamming distance" to "Euclidean distance". An equivalent encoder would map $00 \rightarrow +1+i$, $01 \rightarrow -1+i$, $11 \rightarrow -1-i$, $10 \rightarrow +1-i$ and use a BCC generator $G = [133, 175]$.

Table 2.3 QPSK/BPSK Mapping (PBCC)

Code Label $c_1 c_0$	QPSK Signal x_i
00	+1+i
01	-1+i
10	-1-i
11	+1-i

Code Label c_1	BPSK Signal x_i
0	+1+i
1	-1-i

The last operation of the encoder is the "symbol scrambling." A specified, 256 bit periodic binary sequence is used to control the symbol scrambler. When the binary "s" value into the symbol scrambler is "0," the QPSK/BPSK symbol out of the symbol mapper is sent directly, while an $s = 1$ tells the symbol scrambler to rotate the mapped symbol by 90° (counter-clockwise) as shown in Figure 2.9.

Generation of the period 256 "s" sequence can be described in a two step process. First, a balanced (half 0's, half 1's) vector of length 16 is given by

$$\mathbf{u} = [u_0, u_1, , u_{15}] = [0011001110001011].$$

This vector is repeatedly concatenated with an order 3 circular rotation of the previous vector

$$\mathbf{s} = \mathbf{u} \circ \sigma^3(\mathbf{u}) \circ \sigma^6(\mathbf{u}) \circ \sigma^9(\mathbf{u}) \circ \dots$$

where the 3rd circular shifted is defined by

$$\begin{aligned}
\sigma(\mathbf{u}) &\equiv [u_1, u_2, u_3, \dots, u_{13}, u_{14}, u_{15}, u_0] = [0110011100010110], \\
\sigma^2(\mathbf{u}) &\equiv [u_2, u_3, u_4, \dots, u_{14}, u_{15}, u_0, u_1] = [1100111000101100], \\
\sigma^3(\mathbf{u}) &\equiv [u_3, u_4, u_5, \dots, u_{15}, u_0, u_1, u_2] = [1001110001011001].
\end{aligned}$$

The "$\circ$" symbol represents the concatenation operator, thus the 16 bits in $\mathbf{u}$ is followed by the 16 bits of $\sigma^3(\mathbf{u})$, followed by the 16 bits of $\sigma^6(\mathbf{u})$, etc. The chosen vector $\mathbf{u}$, combined with the fact that 3 and 16 are relatively prime, means that $\sigma^{3m}(\mathbf{u})$ are distinct for $m = 0, 1, \dots, 15$ and $\sigma^{48}(\mathbf{u}) = \mathbf{u}$. Thus this method of symbol scrambler sequence generation has a period of $16 \times 16 = 256$ bits.

Example *PBCC-11*

The encoding of the message bits $[m_1, m_2, m_3, m_4, m_5, m_6, m_7, m_8, \dots]$ = $[1, 0, 1, 1, 1, 0, 1, 0, \dots]$ produces the BCC codeword according to Equation (2.4)

$$\mathbf{c} = [10, 01, 01, 01, 10, 11, 11, 10, \dots].$$

The BCC outputs are translated to the QPSK symbols using Table 2.3

$$[-1 - i, -1 + i, -1 + i, -1 + i, -1 - i, 1 - i, 1 - i, -1 - i, \dots]$$

and selectively rotated according to $s = [00110011\dots]$

$$[-1 - i, -1 + i, -1 - i, -1 - i, -1 - i, 1 - i, 1 + i, 1 - i, \dots]$$

as described in Figure 2.9. ❒

The characteristics and benefits of symbol scrambling are multi-fold—

- ❖ Signal Distance Spectrum
  The distance spectrum of the transmission signal set is invariant to the scram-

bling operation. This is a consequence of distance preserving nature of the $90°$ *rotation [7]. However, unlike a "data scrambling" function (a one-to-one function), symbol scrambling does alter the signal set in beneficial ways.*

- Time Varying Coding
  Typical BCC encoders produce time-invariant codewords. This means that a time shifted version of a valid code sequence is also a valid code sequence. The periodic scrambling, with a long 256 period, makes the code sequences appear aperiodic (actually they are periodic, but with a long period). This effect can be useful.

- Interference Rejection
  When an interfering signal is added to a transmitted packet, it is helpful if the interferer is not a legitimate codeword. This is the case for an aperiodic encoding. Thus, for interferers such as co-channel interference or unmodelled multi-path distortion, the adverse effects of the interfering signal can be significantly reduced.

- Tone Suppression
  Time invariant convolutional coding can generate codewords with unwanted spectral characteristics. For example, the all 0's message will produces an all 0's codeword which, without the symbol scrambler, produce a constant transmission signal. A similar effect will occur if a (small) periodic message is encoded into a periodic codeword. The symbol scrambler removes this signaling possibility, ensuring that signals with poor spectral characteristics are never *transmitted.*

Table 2.4 PBCC-11 Euclidean Weight Distribution

Wt/2E$_s$:	0	9	10	11	12	13	14	15	16	...
Number (PBCC-11):	1	1	6	11	12	45	117	259	629	...
Number (NASA):	1	0	11	0	38	0	193	0	1331	...

The BCC encoder selected for the PBCC-11 code involves a tradeoff between optimal additive white Gaussian noise (AWGN) performance and tolerance to multi-path and other forms of interference. The NASA standard 64 state code (with generator $G = [133, 171]$)[8] is optimized to maximize the Euclidean free distance $\mathrm{d_{free}} = 10$; the Euclidean free distance of the PBCC-11 code is $\mathrm{d_{free}} = 9$. The Euclidean distance spectrum of these two codes is shown in Table 2.4. This data shows that the PBCC-11 code has only one error event of weight 9 and 6 error events at distance 10 while the NASA code has 11 error events of weight 10. These facts explain why the PBCC-11 has only an insignificant loss in SNR, if any, over the AWGN channel (a very small fraction of a dB) as shown in Figure 2.10. The asymptotic coding gain for PBCC-11 is 6.5 dB ($\mathrm{ACG} = 4.5$) over uncoded QPSK. The practical coding gain is about 5.5 dB (as shown in 2.4.2). It is interesting to note that the NASA code has an $\mathrm{ACG} = 5 = 6.9$dB (0.4 dB higher), yet the practical gain is the same 5.5 dB.

Table 2.5 shows a definite advantage for the PBCC-11 code. In this table the symbol or "Hamming" weight distribution of the two codes are compared. It can be seen here that an error event for the PBCC-11 code will span at least 7 QPSK symbols while the NASA code has an error event that spans only 6 QPSK symbols. It was this trade-off between Euclidean distance and symbol distance that led to the selection of the PBCC-11 for the IEEE 802.11b standard.

Table 2.5 PBCC-11 Symbol (Hamming) Weight Distribution

Symbol Weight:	0	6	7	8	9	10	11	12	...
Number (PBCC-11):	1	0	6	8	20	78	204	639	...
Number (NASA):	1	1	4	10	21	66	222	617	...

2.3.4 The "Higher Rate" Standards: *The Future*

The Alantro/TI proposal increases the data rate of the IEEE 802.11b standard in a backword compatible way.

When the engineering team at Alantro started the higher rate project, the following constraints were of main concern —

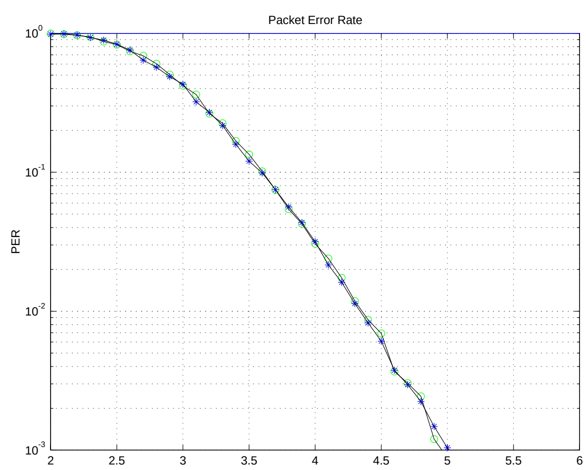


Figure 2.10 Comparison of PBCC-11 and the NASA Code (1000 byte packets)

- ❖ Interoperability with IEEE 802.11b networks
  Introduction of higher rate transmission in an existing network is a prime requirement.

- ❖ Translate coding gain advantage to "double the data rate"
  22 Mbps

- ❖ Compatibility with IEEE 802.11b radios
  8-PSK, 11 MHz symbol rate, short preamble

- ❖ Operate in the same environment as CCK-11
  64 state code $\longrightarrow$ 256 state code; a good engineering solution: cost versus performance

- ❖ Satisfy FCC Requirements
  Same spectral and temporal signal characteristics as IEEE 802.11b; noise and interference tolerance comparable to CCK-11.

2.3.4.1 PBCC 22 Mbps The high rate case (22Mbps) has a 2×3 generator matrix over $\mathcal{Z}_2[D]$:

$$G = \begin{bmatrix} 1 + D^4 & D & D + D^3 \\ D^3 & 1 + D^2 + D^4 & D + D^3 \end{bmatrix}$$

(in octal notation $G = [21, 2, 12; 10, 25, 12]$.)

A 1×3 parity check matrix:

$$H = \begin{matrix} [D + D^2 + D^4 + D^7 & D + D^3 + D^4 + D^5 + D^6 + D^7 \cdots \\ & \cdots\ 1 + D^2 + D^4 + D^6 + D^8] \end{matrix}$$

(In octal notation $H = [226, 372, 525]$.) This BCC encoding function is combined with the "Digital-8PSK" signal mapping shown in Table 2.6 to produce a coded eight level modulation signal.

Table 2.6 8PSK Mapping

Code Label $c_2c_1c_0$	8PSK Signal y_i	Digital-8PSK x_i	$c_2c_1c_0$	y_i	x_i
000	+1+i	+5+5i	100	$\sqrt{2}$	+7i
001	-1+i	-5+5i	101	$\sqrt{2}$i	-7
010	-1-i	-5-5i	110	-$\sqrt{2}$	-7i
011	+1-i	+5-5i	111	-$\sqrt{2}$i	+7

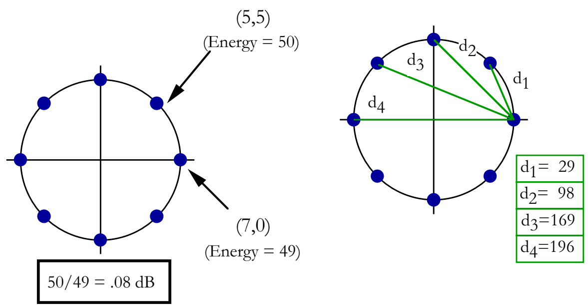


Figure 2.11 Digital-8PSK

This coded modulation was discovered via computer search using a bounding technique illustrated in Figure 2.11 and Table 2.7. The weight values in the table provide a lower bound on the distance between points in the signal constellation. If $(c_2c_1c_0)$ and $(c_2'c_1'c_0')$ are the labels of two points, then

$$||x_i(c_2c_1c_0) - x_i(c_2'c_1'c_0')||^2 \geq w(c_2 \oplus_2 c_2', c_1 \oplus_2 c_1', c_0 \oplus_2 c_0') \tag{2.5}$$

where the operation $\oplus_2$ is modulo 2 addition. Furthermore, this bound is tight (*i.e.*, for each value of w there is a pair of labels that achieve equality in Equation (2.5)). Using this weight function to compare the accumulated distance on a pair of sequences is the basis for the computer search.

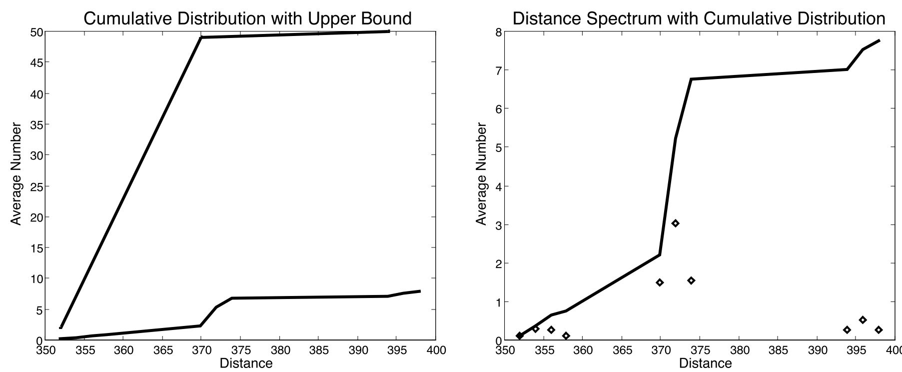
Figure 12(a) shows a plot of the distance spectrum of the PBCC-22 code as well as the bound that was used in the search. One can see that the bound predicts the free distance of the code $\mathrm{d_{free}} = 352$, but overestimates the growth in nearest neighbors. Figure 12(b) shows the average nearest neighbor growth near the free distance of the code, the data for these graphs are presented in Tables 2.8 and 2.9.

Table 2.7 Digital-8PSK Weight Bound

Code Label $c_2c_1c_0$	Weight $w(c_2c_1c_0)$	Code Label $c_2c_1c_0$	Weight $w(c_2c_1c_0)$
000	0	100	29
001	98	101	29
010	196	110	169
011	98	111	29

Table 2.8 PBCC-22 Weight Distribution Bound

Wt/2E_s:	3.56	3.74	3.98	4.14	4.32	4.55	...
99·Wt/2E_s:	352	370	394	410	428	450	...
Number:	2	47	1	53	437	12	...



(a) Large Scale (with Bound)

(b) Fine Scale

Figure 2.12 The Distance Spectrum for PBCC-22

2.4 PERFORMANCE

2.4.1 Spectrum

The power spectrum for all the transmission modes are essentially the same with a small deviation for the original 1 & 2 Mbps modes. To obtain the theoretical power spectral density for a complex waveform of the form

$$x(t) = \sum_{k=-\infty}^{\infty} A_k p(t - k\mathrm{T_s}) \tag{2.6}$$

where A_k is a random symbol sequence, $p(t)$ is the pulse shape and $\mathrm{T_s}$ is the symbol period, the power spectral density is given by the formula

$$S_x(f) = \frac{1}{\mathrm{T_s}}|P(f)|^2 S_A(f\mathrm{T_s}) \tag{2.7}$$

Table 2.9 PBCC-22 Average Weight Distribution

Wt/2E$_s$:	3.56	3.58	3.60	3.62	3.74	3.76	3.78	3.98
99·Wt/2E$_s$:	352	354	356	358	370	372	374	394
Ave. Number:	.0913	.2783	.2677	.0927	1.479	3.017	1.528	.2497
Wt/2E$_s$:	4.00	4.02	4.14	4.16	4.18	4.20	4.32	4.34
99·Wt/2E$_s$:	396	398	410	412	414	416	428	430
Ave. Number:	.5	.2503	1.293	2.786	2.796	1.327	3.843	7.786
Wt/2E$_s$:	4.36	4.55	4.57	4.59	4.61	4.63	...	
99·Wt/2E$_s$:	432	450	452	454	456	458	...	
Ave. Number:	3.933	0.282	1.894	3.267	1.848	.2693	...	

where $P(f)$ is the Fourier transform of the pulse shape and

$$S_A(f) = \sum_l R_A(l) e^{i2\pi l f}$$

is the discrete Fourier transform of the auto-correlation function for the symbol sequence. The fact that Equation (2.7) is the product of two terms shows that the effect of a nontrivial symbol auto-correlation is to modulate the shape of the pulse spectrum. This formula is the basis of the theoretical curves offered in Figure 2.13 and shows very good agreement with experimental results.

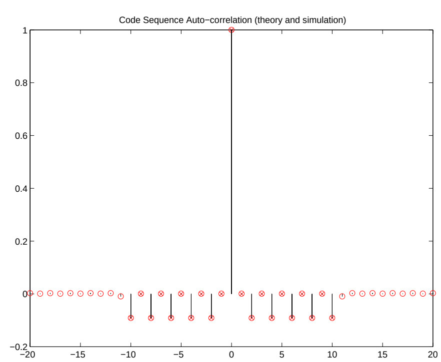
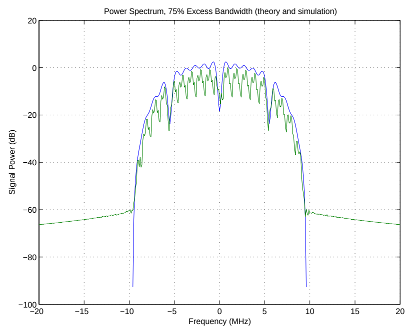


Figure 2.14 Barker Autocorrelation

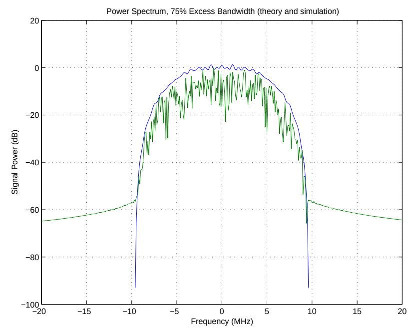
Figure 2.14 shows the auto-correlation for the Barker encoder [4] described in Equation 2.1 on page 41. This nontrivial auto-correlation results in small "ripples" in the power spectral density as observed in both the theoretical and exper-

[4]The symbol sequence A_k is defined with random data encoded according to Equations (2.1) and (2.2), on page 41, and a uniformly distributed phase $A_k = x_{k-N}$, $0 \leq N < 11$. The DFT of the auto-correlation of the Barker sequence

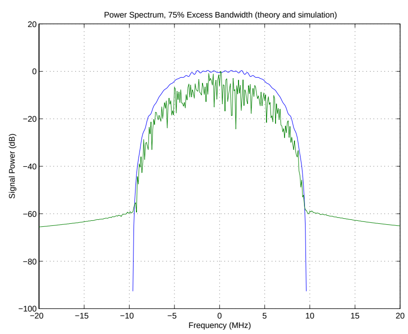
$$S_A(f) = 1 - \frac{2}{11} \sum_{l=1}^{5} \cos(4\pi l f)$$



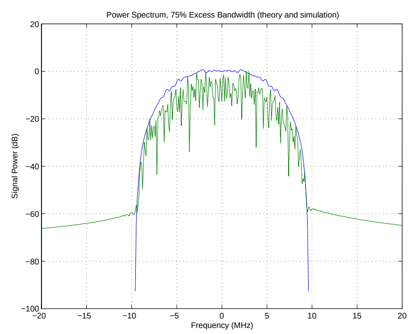
(a) Barker 2 Mbps Spectrum



(b) CCK 11 Mbps Spectrum



(c) PBCC 11 Mbps Spectrum



(d) PBCC 22 Mbps Spectrum

Figure 2.13 The Power Spectrum of Various Codings

imental power spectral results shown in Figure 13(a). Both the CCK code described in Section 2.3.3.1 on page 43 and the PBCC codes described in Section 2.3.3.2 on page 47 and on page 52 offer "white" symbol sequences. This is verified in Figures 13(b) 13(c) and 13(d)

2.4.2 AWGN Performance

The performance of the various combinations of modeling and modulation is presented in Figures 2.15–2.18. In Figure 2.15, the *bit error rate* (BER) of the various choices is shown as a function of the received signal to noise ratio $\mathrm{E_s/N_o}$. Figure 2.16 shows the *packet error rate* (PER), for 1000 byte (8000 bits) packets, as a function of the received signal to noise ratio $\mathrm{E_s/N_o}$. Figure 2.17 shows the PER as a function of the energy per bit to noise ratio $\mathrm{E_b/N_o}$; these curves are useful for computing and comparing the practical coding gains of the systems. Finally,

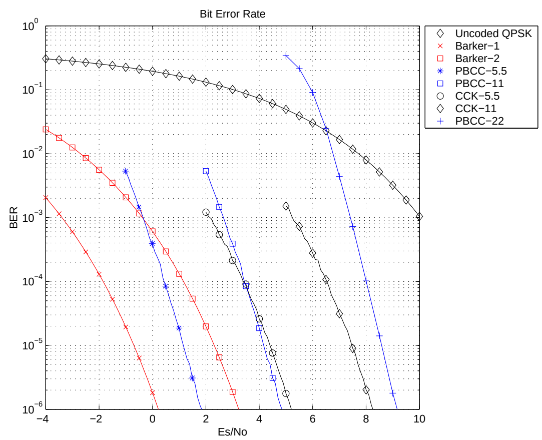


Figure 2.15 Bit Error Rate vs Channel SNR (E_s/N_o)

Figure 2.18 shows the *packet error rate* (PER) as a function of the received signal to noise ratio E_s/N_o for the 22 Mbps system with the multipath receiver that is the basis of the Alantro/TI baseband receiver product. The multipath is modeled using a method developed by the IEEE 802.11 committee and indexes the multipath by a factor known as the "delay spread" [9]. In this model, an increase in delay spread corresponds to a more severe multipath environment.

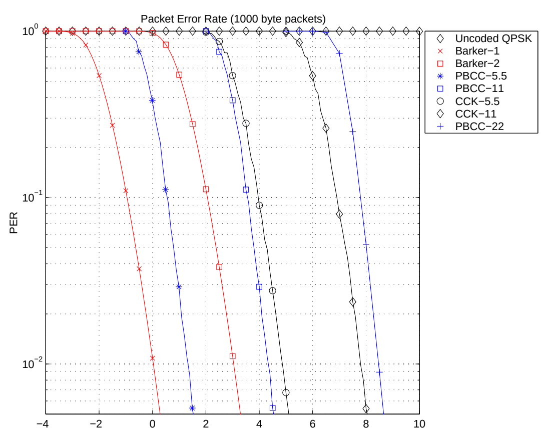


Figure 2.16 Packet Error Rate vs Channel SNR (E_s/N_o)

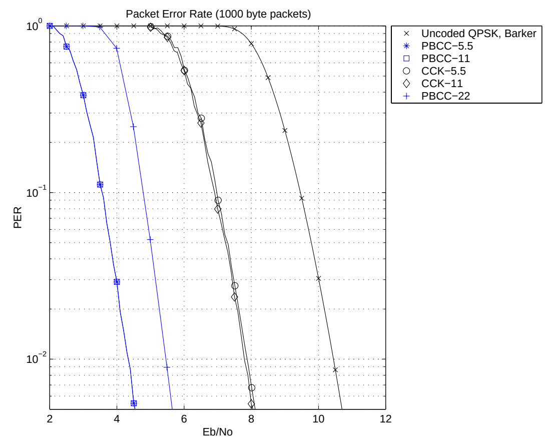


Figure 2.17 Packet Error Rate for Coding Gain $(\mathrm{E_b}/\mathrm{N_o})$

2.4.3 Computational Complexity

A comparison of the computational requirements to decode all the high rate modes is given in Table 2.10. This table shows the number of basic computations required to perform an optimal decoding in AWGN using the Viterbi algorithm [8]. Note that these results does not consider the cost of dealing with the prevailing issue in wireless Ethernet, multipath. Thus these results, which are useful for a raw comparison of the various coding schemes, does not give a complete picture of complexity required to implement a wireless Ethernet baseband processor.

Table 2.10 Trellis Complexity with Viterbi Decoding Compared

Code	Branches per Information Bit	Mega-Branches per Second
CCK5.5	14	154
CCK11	37	407
PBCC5.5	128	704
PBCC11	128	1,408
PBCC22	1024	11,264

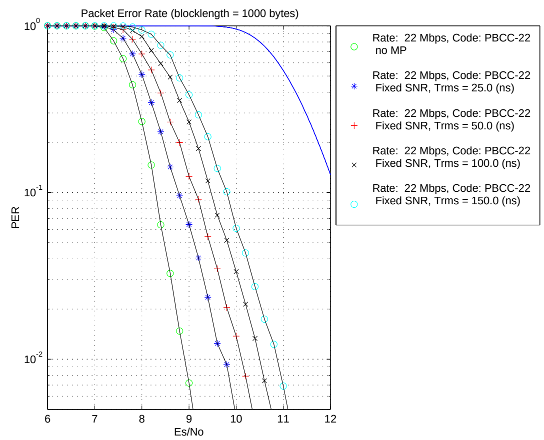


Figure 2.18 Packet Error Rate in Multipath (22 Mbps)

2.5 SPREAD SPECTRUM TRANSMISSION

In wireless communications, such as IEEE 802.11b and other shared media systems, information is often encoded using spread spectrum signaling methods. The *spectral efficiency* of a digital transmission system is defined as the ratio of the user *data rate* (in bits/second) to the *bandwidth* (in Hertz) of the power spectral density (suitably defined) of the ensemble of transmission signals. As argued in the very thought provoking chapter, [10], Jim Massey considered a information theoretic definition of spread spectrum, and studied some of the consequences of his view. This definition is described in some detail in this section beginning on page 61. For example, Massey's definition imples that in spread spectrum signaling systems, the spectral efficiency will be low. He also shows, via examples, that the converse is not true: a low spectral efficiency does not imply a spread spectrum signal set.

Massey demonstrated that in systems with a low spectral efficiency, the use of spread spectrum is a reasonable means of communications that has only a modest, acceptable, loss in Shannon capacity. He also showed that in high spectral efficiency systems, mathematically precise notions of spread spectrum imply a very significant, un-economic, loss in capacity. In the Massey framework, if the spectral efficiency is not a small fraction of 1, spread spectrum is not practical.

This view is in contrast to the view of the U. S. Federal Communications Commission (FCC) which uses a more pragmatic definition of spread spectrum. The FCC defines direct sequence spread spectrum in a much less restrictive way. According to the FCC:

Direct Sequence Systems *"A spread spectrum system in which the carrier has been modulated by a high speed spreading code and an information data stream. The high speed code sequence dominates the "modulating function" and is the direct cause of the wide spreading of the transmitted signal."*

and

Spread Spectrum Systems *"A spread spectrum system is an information bearing communications system in which: (1) Information is conveyed by modulation of a carrier by some conventional means, (2) the bandwidth is deliberately widened by means of a spreading function over that which would be needed to transmit the information alone. (In some spread spectrum systems, a portion of the information being conveyed by the system may be contained in the spreading function.)"*

It is interesting to observe that it is the last parenthetical element that differentiates the strict requirements of Massey's definition and that the FCC rules that have opened the door to higher spectral efficiencies and user data rates in the "ISM" 2.4 GHz band[5]. Without flexibility and pragmatism on the part of the FCC, a more technically strict definition such as Massey's would have prevented the wide spread success of the IEEE 802.11b standard. It has been indicated by the FCC that as the standardization process continues to make progress in the development of higher performing wireless Ethernets, regulators will continue to support the needs of the industry and consumers.

In the process of significantly increasing the data rate, the spread spectrum nature of the signal, in the narrow sense of Massey, is sacrificed. However, the flexible FCC definition allowed the FCC to certify the existing IEEE 802.11b 11Mbps systems under direct sequence spread spectrum rules. This practical approach to regulation is based upon the fact that as an interferer, the high rate IEEE 802.11b signals are the same as the classical low rate Barker signals. This is true both in the frequency characteristics, as shown in the power spectral density (Figure 2.13), as well as in the time domain or the temporal characteristics of the transmitted signals.

Furthermore, the IEEE 802.11 specifies three disjoint frequency bands for wireless Ethernet systems This means that the legacy 2 Mbps systems send a total of 6 Mbps in the entire ISM band while the 11 Mbps supply 33 Mbps in the band; the 22 Mbps system double the total capacity to 66 Mbps.

Radio spectrum is a rare and valuable resource and it is the responsibility of the FCC to insure that the resource is used for the public good and in an efficient way. One compelling issue is the demands from the public for higher performance data transmission. Another important issue is the need to avoid the introduction of new of signals with spectral and temporal characteristics that were formerly disallowed under the existing rules. Such a change threatens the large base of current products that were built under existing rules with interference that was not previously allowed or anticipated; from a fairness position, this is unjust.

With the huge success of the IEEE 802.11b standard, one can see the wisdom of the FCC. It is anticipated that the future regulations will continue to satisfy the demands for higher performance while maintaining a level playing field. The beauty of the PBCC-22 modulation approach is that the data rate is doubled while

[5]The "ISM" band is 83.5 MHz wide using the range 2.4000-2.4835 GHz.

maintaining backword compatibility with existing networks using a signal with the same interference characteristic as the existing signal sets. The noise immunity or "processing gain" of the system is the same as the CCK-11 system. The spread spectrum nature of the new signals, in the sense of Massey, is the same as the existing systems; this is discussed in the following section of this chapter. It is demonstrated that from an information theoretic viewpoint, the spread spectrum nature of the new signals is identical to the existing signal sets used in currently deployed networks. Thus, in under any reasonable definition, the PBCC-22 and the CCK-11 systems are equally spread spectrum.

2.5.1 Massey's Definition of Spread Spectrum

Massey defined two notions of bandwidth and argued that the indication of spectrum spreading was related to the size of the ratio of the two. The first definition of bandwidth relates to the spectral occupancy of a given signal or a collection of signals. This form of bandwidth, B_F, is known as the "Fourier bandwidth" and relates to the span of frequencies occupies by the signal(s). As is often the case in Communication Theory, the exact numerical value of the Fourier bandwidth for a given signal or set of signals depends on a measurement criteria such as "3 dB" bandwidth or 95% power bandwidth, etc. Such required criteria are often needed to define other quantities of interest in Communications theory; examples include the definition of signal to noise ratio (SNR) and power spectral density. The Fourier bandwidth is directly related to the "Nyquist bandwidth" [11] which relates to periodic sampling of a signal (or sets of signal) and is of fundamental importance in the study of digital signal processing (DSP).

Massey's second notion of bandwidth is related to the fundamental problem of information transmission and is meaningful to define only for a collection or a set of signals. Fundamentally, the problem of information transmission is one of signal design and signal detection. Massey logically argues that the definition of spread spectrum should only involve the signal design issue and not signal detection (*i.e.*, the determination of spread spectrum character of a transmission scheme should not change with a change in the receiver).

Signal design involves the creation of a collection of signals used by a transmitter to represent the multitude of messages that the transmitter is trying to convey. In the signal design problem, various parameters are considered in order to optimize the transmission systems. Such parameters include transmission power, Fourier bandwidth, power spectrum and data rate and a host of others including the dimensionality of the signal set.

The *data rate* parameter of a signal set relates to the size of the collection or *number* of signals in the signal set; a system transmits at a rate of R bits per second if, over a time interval of length T seconds, the designed signal set defines 2^{RT} distinct signals. With such a collection of signals, $k = RT$ bits of information can be transmitted by assigning a correspondence between the list of signals in the signal set and the 2^k possible values for a k bit message.

The *dimensionality* of a signal set involves the standard notion of basis as defined in the area of linear algebra. Roughly speaking, the *dimension* of a signal set relates to the *minimum* number of *independent* parameters (*i.e.*, numbers) required to describe the collection of signals.

The second definition of bandwidth, B_S, relates to the dimensionality of a signal set and describes the linear complexity of the scheme; a system transmits using a bandwidth of B_S Hz if over a time interval of length T seconds, the designed signal set has a basis with $\mathrm{B}_S T$ elements. Due to the strong relationship between this notion of bandwidth and Information Theory, Massey called this second definition the "Shannon bandwidth."

Note that the Fourier bandwidth, the Shannon bandwidth and the data rate are distinct ideas that all describe attributes of a signal set. For example, the spectral efficiency of a system is the ratio of the data rate to the Fourier bandwidth R/B_F. Another important parameter is the *spreading ratio* $\rho = \mathrm{B}_F/\mathrm{B}_S$ which relates the two notions of bandwidth.

The first observation that Massey noted was the Theorem that says that the Fourier bandwidth is never less than the Shannon bandwidth, $\mathrm{B}_F \geq \mathrm{B}_S$. This means that the spreading ratio satisfies the inequality

$$\rho = \frac{\mathrm{B}_F}{\mathrm{B}_S} \geq 1.$$

Furthermore, Massey argued that the spreading ratio is the logical measure of the degree in which a communications system spreads the spectrum. If a given system has a large value for ρ, say 10 or 100, then it should be considered a spread spectrum system, and conversely, a system with a spreading ratio ρ near the minimum of 1 would not be labeled a spread spectrum system. It would be debatable if a system with a spreading ratio of say $\rho = 4$ is spread spectrum or not, this is the "gray" area.

In Shannon's original 1948 paper [12], a famous formula for the capacity of a bandlimited additive white Gaussian channel was presented

$$C(P/\mathrm{N_o}, \mathrm{B}_F) = \mathrm{B}_F \log_2(1 + \frac{P}{\mathrm{N_o}\mathrm{B}_F}) \quad \text{bits/second} \tag{2.8}$$

where P is the signal power, $\mathrm{N_o}$ is the white Gaussian noise level and B_F is the permissible Fourier bandwidth. The interpretation of the Shannon capacity is that reliable transmission is possible, for a given signal to noise ratio (SNR) $P/\mathrm{N_o}$ and Fourier bandwidth B_F, *if and only if* the rate of transmission is no more than the Shannon capacity C. In practical terms, the Shannon limit defines an objective data rate goal for a given signalling environment. For the past 53 years, communications engineering have been striving to approach this goal.

If one is to impose the requirement that the transmission system operate with a required spreading ratio of ρ, then the formula is modified to be

$$C(P/\mathrm{N_o}, \mathrm{B}_F, \rho) = \frac{\mathrm{B}_F}{\rho}\log_2(1 + \frac{P\rho}{\mathrm{N_o B}_F}) \quad \text{bits/second.} \tag{2.9}$$

To understand the limitations imposed on the Shannon capacity when spreading is introduced, it is helpful to interpret Equation (2.9).

First it is to be noted that spreading in this sense incurs a loss in capacity; for a fixed SNR and bandwidth, the Shannon capacity monotonically decreases with increasing spreading ρ; if $\rho > 1$, $C(P/\mathrm{N_o}, \mathrm{B}_F) \equiv C(P/\mathrm{N_o}, \mathrm{B}_F, 1) > C(P/\mathrm{N_o}, \mathrm{B}_F, \rho)$. However, as noted in Massey's paper, there are often situations where the loss is small and spreading is reasonable. The modified Shannon formula, Equation (2.9), involves the product of two terms, the symbol frequency $\left(\frac{\mathrm{B}_F}{\rho}\right)$ measured in "symbols per second" and the normalized data rate $\left(\log_2(1 + \frac{P\rho}{\mathrm{N_o B}_F})\right)$ measured in "bits per symbol." The spectral efficiency of a system, which is the data rate divided by the Fourier bandwidth, is the normalized rate divided by the spreading ratio and a capacity given by the expression $\left((1/\rho) \cdot \log_2(1 + \frac{P\rho}{\mathrm{N_o B}_F})\right)$.

Spreading is useful only when the normalized rate or the spectral efficiency is very small[6]. Since the normalized rate grows with the spreading ratio ρ, for a given situation (*i.e.*, SNR and bandwidth), there will be a practical limit on the spreading ratio. For example, in order to obtain 90% of the Shannon capacity, with a very modest spreading ratio of $\rho = 2$, requires that the normalized rate be less than about .3 bits per symbol and a spectral efficiency of less than .15 bits-per-second per Hz. Similarly, a system with a spreading ratio of $\rho = 10$ operating with a tiny spectral efficiency of .01 bits-per-second per Hz will incur a greater than 20% loss in Shannon capacity from the spreading.

2.5.2 Spread Spectrum in Wireless Ethernet

It is interesting to see how Massey's notion of spreading relate to the DSSS wireless Ethernet standard and the higher rate extensions. In terms of the coding level, the Barker systems introduce a nontrivial spreading ratio of $\rho = 11$ (2 Mbps) and $\rho = 22$ (1 Mbps). All the high rate (> 2 Mbps) cases have $\rho = 1$, with the exception of PBCC-5.5 which has $\rho = 2$. In practice, the wireless Ethernet signals use a nontrival excess bandwidth pulse shape so that the occupied bandwidth is larger than the 11 MHz symbol rate. A comparison of the spreading ratio for the various choices are given in Table 2.11. It is important to note, that in terms of Massey's spread ratio, all the high rate systems have the same value (with the exception of PBCC-5.5).

[6] A small loss occurs when the approximation $\log_2(1 + x) \approx \log_2(e) \cdot x$ is close; this occurs only for small x.

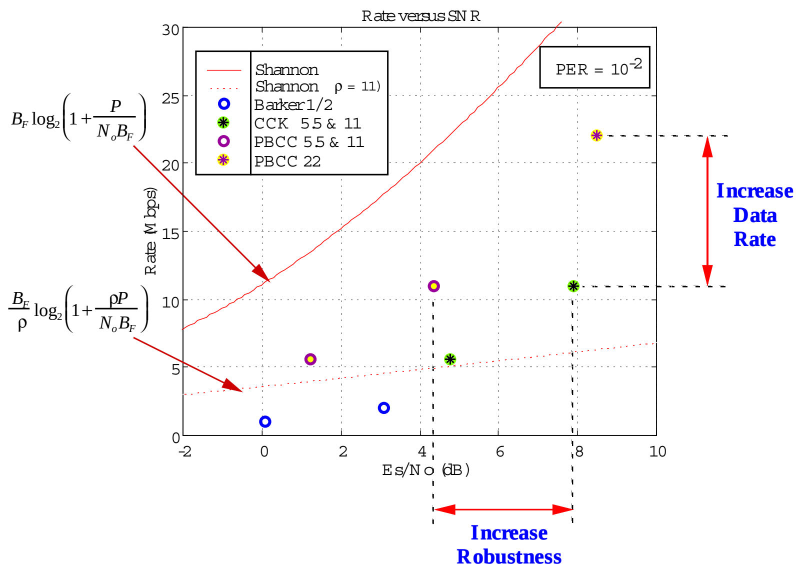


Figure 2.19 Performance Wireless Ethernet Relative to the Shannon Limit

Table 2.11 Wireless Ethernt Spreading Ratios

Scheme	Code Level	Waveform Level (75% excess bandwidth)
Barker-1	22	40.00
Barker-2	11	20.00
CCK-5.5	1	1.82
CCK-11	1	1.82
PBCC-5.5	2	3.64
PBCC-11	1	1.82
PBCC-22	1	1.82

Thus, for example, from the viewpoint of information theory, the CCK-11 and the PBCC-22 signals show the same degree of of siganl spreading.

In Figure 2.19 the offered data rate and signal to noise ratio requirements for the IEEE 802.11b standard and the Alantro 22 Mbps extension are displayed. On the "x-axis" is the signal to noise ratio defined as the symbol energy to noise ratio E_s/N_o[7] while the "y-axis" is the data rate of the system assuming the common 11 MHz symbol frequency that is common to the standard. The upper solid curve is the Shannon limit

[7] $E_s = P/B_S = P \cdot \rho/B_F$.

as described by Equation (2.8), on page 62,. The dotted curve shows the Shannon limit assuming a spreading ratio of $\rho = 11$ in Equation (2.9) (this is the spreading ratio of the 2 Mbps Barker system). The individual points on the graph describe the various data rates and SNR requirements of the host of systems. Note that the SNR requirement is defined as the SNR required to maintain a packet error rate (PER) of 10^{-2} with a 1000 byte (8000 bit) packet; this 1% PER threshold is a standard measure of "robustness" used by the IEEE 802.11 committee in deliberations leading to the selection of standards.

This graph in Figure 2.19 shows how the superior error control properties of the PBCC method of signal generation can be used to improve robustness (*i.e.*, SNR requirements) or user data rate. It is also interesting to see that the existing IEEE 802.11b standard, which is widely deployed in FCC certified products, violate the Massey spread spectrum result in terms of Shannon theory. The reason for this discrepancy is explained by the pragmatism of the FCC regulatory body, the FCC's broader definition of spread spectrum as well as the strictness of Massey's theoretical result. Without such flexibility on the part of the FCC, there would be no high performance wireless Ethernets.

2.6 RANGE VERSUS RATE WIRELESS LOCAL AREA NETWORKS

To determine the effectiveness of a family of modulation and coding options for wireless Ethernet applications, it is useful to understand how data throughput and distance are traded. In this section, a mathematical model is presented that allows for a rational comparison of IEEE 802.11g proposals. In this study the legacy CCK systems are compared to the PBCC, CCK/OFDM and 11a/OFDM. The comparison demonstrates that while PBCC and 11a/OFDM follow similar rate versus range curves in the 2.4 GHz band, the additional overhead required for 802.11b backword compatibility of the CCK/OFDM has a severe rate versus range penalty.

This section is organized in two parts. In the first part, the background information required compute the range and throughput of a wireless Ethernet system is described. In the following section, a comparison of various alternatives considered by the Task Group G is presented. The analysis shows the superiority of the PBCC based systems over the CCK/OFDM ones.

The IEEE 802.11a standard is defined only for the 5 GHz band, however, the Task Group G has been considering a proposal for an "11a-like" scheme known as CCK/OFDM. Thus it is prudent to consider the capabilities of 11a/OFDM if it were transmitted in the 2.4 GHz band. Of course, if the 802.11a signal is transmitted in the 2.4 GHz band it would not be backward compatible with the existing base of IEEE 802.11b networks since the preamble structure of the two signaling sets are incompatible. The introduction of pure 11a/OFDM signals in the 2.4 GHz band could be quite disruptive since the two networks would cause mutual interference if used in an overlapping band of frequencies.

The idea of CCK/OFDM was introduced by Intersil as a method of combining 802.11a signals in a backward compatible way with the single-tone modulations (*i.e.*, Barker, CCK and PBCC) that are the basis of IEEE 802.11b. The scheme involves transmitting an IEEE 802.11b preamble followed by a transition to the OFDM blocks defined in the IEEE 802.11a standard. Unfortunately, the backwards compatible requirement makes the overhead of the CCK/OFDM solution excessive. For the highest mandatory rate, in additive white Gaussian noise (AWGN), PBCC-22 achieves a throughput of 12.8 Mbps at a range that is 95% of the CCK-11 system while the CCK/OFDM-24 achieves 13.0 Mbps (a trival improvement in rate) at a range that is only 76% as far. (The notation XXX-NN denotes an "XXX" signal with a maximum instantaneous rate of "NN" Mbps.) In terms of area, these factors are 90% and 58% coverage, respectively. With 100 ns of multipath, the range numbers become 92% and 74%.

It is interesting to note that 11a/OFDM signals, which does not suffer from the large overhead required to be backword compatible with the 11b preamble, has the same ranges as CCK/OFDM in the 2.4 GHz band but much higher throughput. For example with 11a/OFDM-24 the throughput is 18.5 Mbps. The curves for PBCC and an 11a/OFDM system (used in the 2.4 GHz band) shown in Figure 2.23 demonstrate that for ranges up to 60% of the CCK-11 range, the two schemes are very competitive, while the CCK/OFDM system significantly lags both solutions in all cases.

Furthermore, IEEE 802.11a systems are designed for the 5 GHz bands; these higher frequencies experience a penalty due to the higher frequency. This factor predicts that IEEE 802.11a systems will have range problems compared to 2.4 GHz systems at the same power levels and throughput.

2.6.1 Background Development

The calculation of user data rate or throughput versus distance involves several components that include:

- Calculation of symbol signal-to-noise ratio (E_s/N_o) required for maximal operational packet-error-rate (PER)
- Translation of waveform signal power to symbol energy
- Determination of receiver noise floor power spectral density (N_o) and receiver sensitivity
- Formulation of propagation loss model that relates receiver signal power to distance
- Determination of the maximum throughput of the system including effects of preambles and acknowledgments
- Understanding of effects of multipath distortion on receiver performance

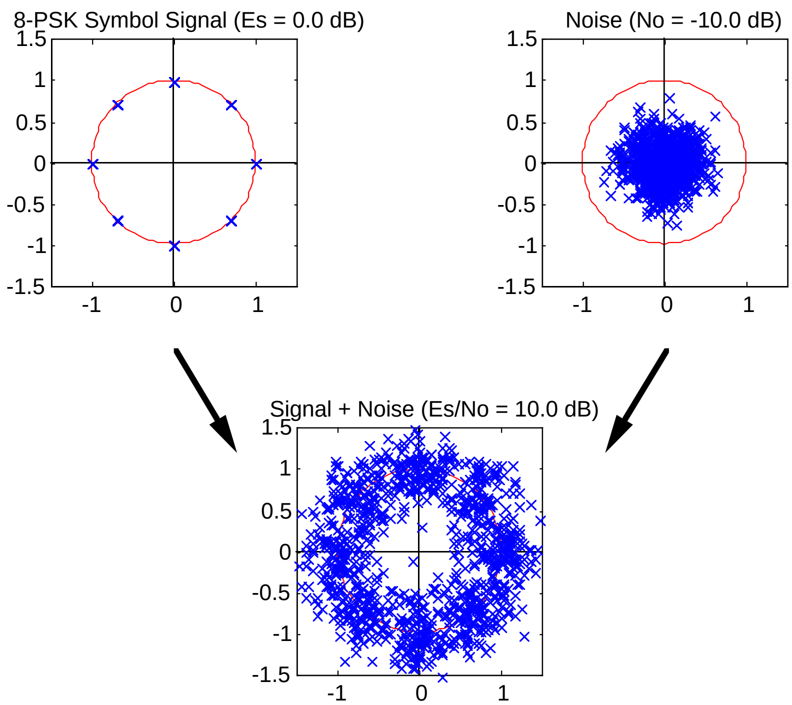


Figure 2.20 Signal Plus Noise in 8-PSK

2.6.1.1 Symbol SNR and PER In bandpass digital transmission, a basic concept is the discrete time, 2-dimensional symbol. In wireless Ethernet applications for example, phase shift keying (PSK) and quadrature amplitude modulation (QAM) symbols are transmitted by the sender to convey the intended message. At the receiver, a detection process is used to process the corrupted symbol to determine the message that was transmitted. The corruption of the symbol can include both noise and signal distortion. The noise in the receiver is typically a function of how well the receiver radio can amplify the very small receive signal to bring it to a level that is required by the detection process. The bulk of the noise is modeled as additive white Gaussian noise since the source of the noise is wide band (relative to the signal). The dominant form of signal distortion is can be attributed to multipath distortion which arises from multiple refections of the signal during propagation. The symbol signal-to-noise ratio (SNR) relates the average symbol signal power $\mathrm{E_s}$ to the variance of the symbol noise $\mathrm{N_o}$ (*i.e.*, the noise in 2 dimensions). For a PSK signal, the symbol energy is constant, $\mathrm{E_s} = A^2$, where A is the radius of the circle. For QAM, the symbol energy is generally not constant; the average symbol energy $\mathrm{E_s} = A^2$ for

4-QAM (which is the same as QPSK) and $\mathrm{E_s} = 5A^2$ for 16-QAM. In Figure 2.20, an 8-PSK symbol with $\mathrm{E_s/N_o}$ = 10 dB is shown.

The effect of the $\mathrm{E_s/N_o}$ value on system performance is reflected in the packet error rate (PER) of the detector. In the IEEE 802.11 working groups, a threshold PER of 10^{-2} (one packet error in 100 packet transmissions) is considered the maximum acceptable value. Notice that due to the incorporation of a reliable error detection code within the body of the packet, it can be assumed that an error corrupted packet will be detected and rejected (and typically retransmitted). When the PER rises above the threshold, the system typically backs down to a more reliable, albeit slower transmission mode. The PER is also a function of packet length, for small BER (bit error rate) the PER is approximately N*BER where N is the length of the packet in bits. Thus, for a packet with 1000 bytes of data and a PER less than 10^{-2} requires a BER of less than 1.25×10^{-6}.

The detector performance is affected by the choice of transmission signal constellation set and the form of forward error control (FEC) designed into the transmission system as well as the detection algorithm used at the receiver. In Table 2.12, the value of $\mathrm{E_s/N_o}$ required to achieve a PER of 10^{-2} in AWGN is given. For example, the table shows that the CCK-11 system requires at least 7.8 dB of $\mathrm{E_s/N_o}$ for an acceptable PER while the PBCC-11 system requires 4.3 dB of SNR. This 3.5 dB improvement in SNR is a direct consequence of the 64 state binary convolutional code (BCC) [8] specified in the IEEE 802.11b standard for PBCC transmission [1]. Notice that the OFDM-12 systems, which incorporate a similar 64 state code, has the same coding gain advantage over CCK-11. All of these systems use a QPSK signal set and transmit at a rate of 1 bit-per-symbol due to the presence of a rate 1/2 FEC encoder.

The higher rate systems incorporate various signal sets and FEC codes. Consider the systems that transmit 2 bits-per-symbol. As a reference, uncoded QPSK requires a threshold $\mathrm{E_s/N_o}$ of 13.5 dB. The PBCC-22 system combines 8-PSK modulation with a 256 state BCC with a 2/3 code rate. The threshold for PBCC-22 is 8.5 dB, an improvement of 5 dB over uncoded QPSK; this 5 dB improvement is known as the coding gain. The OFDM-24 systems use 16-QAM symbols with the same 64 state BCC as OFDM-12; the threshold $\mathrm{E_s/N_o}$ is 10.0 dB, a 3.5 dB coding gain over uncoded QPSK.

2.6.1.2 Signal Power to Symbol Energy, Receiver Noise and Sensitivity The signal and noise energy collected at the radio and baseband processor is a function of several factors. With the proper design of transmit signal and receiver structures, incorporating such concepts as “matched filtering”, the symbol signal-to-noise ratio will satisfy the equation

$$\mathrm{E_s/N_o} = \frac{\mathrm{P_R T_s}}{\mathrm{N_o}}$$

where $\mathrm{P_R}$ is the receive signal waveform power, $\mathrm{T_s}$ is the symbol period and $\mathrm{N_o}$ is the noise floor power spectral level.

Intuitively, the symbol energy is derived from the product of signal power (energy per second) and a symbol period (seconds). Notice that such factors as "excess bandwidth", which are important in system design, do not play a role in the equation that matched signal power to symbol energy.

The noise level $\mathrm{N_o}$ of the receiver is difficult to estimate analytically since many factors are needed. Such factors include the "noise figure" of the receiver amplifiers and other physical quantities. The fact that the noise floor level (*i.e.*, the power spectral density height) and the symbol noise variance (*i.e.*, the 2 dimensional noise variance) are the same is the fact that white noise has the interesting property that the amount of noise is "the same in all directions". If white noise with a power spectral density level of $\mathrm{N_o}$ is passed through a filter with impulse response h(t) or transfer function H(f), then the output power is equal to where

$$\|h\|^2 = \int_{-\infty}^{\infty} |h(t)|^2 \, dt = \int_{-\infty}^{\infty} |H(f)|^2 \, df$$

independent of the shape of $h(t)$ or $H(f)$. (In fact, one could use this as a definition of "white" noise.) For this reason, and other related calibration problems, rather than attempt to find an absolute value for the noise floor and the range, we prefer a relative analysis. This approach allows one to compare systems with respect to a known solution, the basis IEEE 802.11b, 11 MHz system.

In our analysis, we take CCK-11 as the base system that is used to set a "stake in the ground" from which other systems are compared. We define a new quantity $\mathrm{E_o}$ that will account for factors such as symbol rate and power overhead. The CCK-11 system has a symbol period $\mathrm{T_s} = 91$ (nsec) (*i.e.*, the symbol frequency is 11 MHz). When the various systems are compared in terms of range, the ratio of the symbol period to the period of CCK must be considered; for CCK-11, we take $\mathrm{E_o}$ = $\mathrm{E_s}$ or $\mathrm{E_o}/\mathrm{E_s}$ = 1 (= 0 dB). For PBCC-22, which uses the same symbol rate, $\mathrm{E_o}/\mathrm{E_s}$ = 1 (= 0 dB) also. However other PBCC modes, such as PBCC-33 , use a faster symbol rate of 16.5 MHz, $\mathrm{T_s} = 61$ (nsec), to increase the data rate. In these modes the bandwidth is preserved by decreasing the excess bandwidth to about 20% from the 80% of typical CCK-11 and PBCC-11 systems. In this case, the non-trivial ratio of symbol periods makes $\mathrm{E_o}/\mathrm{E_s}$ = 3/2 (= 1.76 dB).

In the case of OFDM systems the equivalent symbol period is based on a 12 MHz, $\mathrm{T_s} = 83$ (nsec) period. This accounts for a factor of 12/11 (= .38 dB) in the calculation of $\mathrm{E_o}$. The reasoning for the 12 MHz value can be seen in many ways. For example, the OFDM systems use 48 tones to convey data. Each of the tones is allocated an equal fraction of the transmit power (ideally each tone would receive 1/48 th of the power, in fact each tone gets 1/52 of the power, more on this later) and uses a long symbol period. The symbol period for each tone is 4 usec. This period is obtained via a 64 point FFT that is cyclically extended by 25% (16 terms) to 80 points and clocked using a 20 MHz clock, resulting in a 250 kHz symbol frequency. The 12 MHz follows from the fact that 48 independent tones generating 250k symbols per second will generate 12M symbols per second in total.

There is another factor that must be considered in the calculation of $\mathrm{E_o}$ for OFDM systems. This factor is the OFDM signal power overhead that results from 2 sources. The first source is the fact that 52 equal power tones are transmitted, but 4 of the tones are used for modem tracking functions and do not carry user information; this results in a factor of 52/48 (= .348 dB). The other source is a consequence of the cyclic extension technique for mitigating the effects of multipath to minimize the occurrence of inter-symbol interference (ISI). The transmitted tones are orthogonal (the "O" in OFDM) over the 64 points (not the 80) or 3.2 usec (not the full symbol period of 4 usec). The receiver uses this subinterval of 3.2 usec in the detection process and thus sacrifices 5/4 (= .969 dB) of the received signal power.

Thus, for OFDM systems, the calculation of $\mathrm{E_o}/\mathrm{E_s}$ = 65/44 (= 1.695 dB); this includes both the symbol rate difference and the signal power overhead.

2.6.1.3 Propagation Loss The signal power observed at the input to the receiver radio is a function of several factors including transmit signal power, antenna gain and propagation loss from the channel. A common model for propagation loss as a function of distance d takes the form

$$L(d) = c \cdot d^{\nu}$$

where the exponent ν is the critical parameter of the loss model. In free space, with a spherical radiation of transmit power, the exponent ν = 2 since the area of the surface of a sphere grows with the square of the radius. In less ideal situations, such as in a building with walls and such, a larger value for the exponent ν would be observed. In the IEEE 802.15 committee, a model for propagation loss in Bluetooth systems assume a free space model up to 8 meters and a ν = 3.3 exponent for larger distances

$$L(d) = \begin{cases} \left(\frac{4d_1\pi}{\lambda}\right)^2 \left(\frac{d}{d_1}\right)^2, & d \leq d_1, \\ \left(\frac{4d_1\pi}{\lambda}\right)^2 \left(\frac{d}{d_1}\right)^{\nu}, & d \geq d_1, \end{cases} \tag{2.10}$$

where the crossover distance is taken to be$d_1 = 8$ and where the wavelength at 2.4 GHz is λ = .1224 meters. Note that the loss function is a continuous in the distance parameter d [13]. This model is derived in [14], and supported by equation 3.1, page 71, [15]; it has been adopted by the IEEE 802.11 committee as well.

In this section, the IEEE 802.15/802.11 model at large distance is assumed, *i.e.*, ν = 3.3. To normalize relative to CCK-11, the waveform signal to noise ratio

Table 2.12 Range versus Rate Data, AWGN

Item	Mod	Max Rate	Max Throughput*	Es/No (PER:10e-2)	Eo/No (PER: 10e-2)	Eo/Es	Range (ν = 3.3)
		Mbps	Mbps	dB	dB	dB	
1	CCK-5.5	5.50	4.7	4.8	4.8	0.0	123
2	CCK-11	11.00	8.1	7.8	7.8	0.0	100**
3	Uncoded QPSK	22.00	12.8	13.5	13.5	0.0	67
4	PBCC-5.5	5.50	4.7	1.3	1.3	0.0	157
5	PBCC-8.25	8.25	6.3	1.3	3.1	1.8	142
6	PBCC-11	11.00	8.1	4.3	4.3	0.0	128
7	PBCC-16.5	16.50	10.7	4.3	6.1	1.8	113
8	PBCC-22	22.00	12.8	8.5	8.5	0.0	95
9	PBCC-33	33.00	15.9	8.4	10.2	1.8	85
10	PBCC-49.5	49.50	18.9	11.4	13.2	1.8	69
11	PBCC-66	66.00	20.9	14.4	16.2	1.8	56
12	CCK/OFDM-6	6.00	5.0	1.2	2.9	1.7	141
13	CCK/OFDM-12	12.00	8.4	4.3	6.0	1.7	113
14	CCK/OFDM-24	24.00	13.0	10.0	11.7	1.7	76
15	CCK/OFDM-36	36.00	15.9	13.2	14.9	1.7	61
16	CCK/OFDM-48	48.00	17.9	17.6	19.3	1.7	45
17	CCK/OFDM-54	54.00	18.5	18.9	20.6	1.7	41
18	11a/OFDM-6	6.00	5.6	1.2	2.9	1.7	141
19	11a/OFDM-12	12.00	10.5	4.3	6.0	1.7	113
20	11a/OFDM-24	24.00	18.6	10.0	11.7	1.7	76
21	11a/OFDM-36	36.00	25.2	13.2	14.9	1.7	61
22	11a/OFDM-48	48.00	30.5	17.6	19.3	1.7	45
23	11a/OFDM-54	54.00	32.5	18.9	20.6	1.7	41

* 1000 Byte Packets with Preamble, 1 SIFS, 1 CCK-11 ACK with Preamble, 1 DIFS
** Reference range = 100

$$\mathrm{P_R/N_o} = \frac{c_o}{d^{3.3}}$$

where the constant

$$c_o = \frac{d_o^{3.3}(\mathrm{E_s/N_o})}{\mathrm{T_s}}$$

is determined by setting $d_o = 100$, $\mathrm{E_s/N_o}$ is equal to the SNR for CCK-11 that has a PER of 10^{-2} (*i.e.*, 7.8 dB) and $\mathrm{T_s} = 91$ nsec.

Note that choosing $d_o = 100$ forces the range of CCK-11 to be the normalized range of 100. This can be used to estimate the range of other systems once the absolute range of CCK-11 is known. For example, if a realized system has a CCK-11 range of 40 meters, then the absolute range for other systems, such as PBCC-11 can be estimated. In this case, Table 2.12 indicates a normalized range of 128 (*i.e.*, 28% more); this translates into an absolute range of 51.2 meters. Similarly, a PBCC-22 system will reach 38 meters, an X/OFDM-12 system will have a range of 45.2 meters and X/OFDM-24 will have 30.4 meters reach.

Table 2.13 Range versus Rate Data, AWGN plus Multipath Distortion (100 ns)

	Mod	Max Rate	Max Throughput*	Es/No (PER:10e-2)	Eo/No (PER: 10e-2)	Eo/Es	Range (ν = 3.3)
Item		Mbps	Mbps	dB	dB	dB	
1	CCK-11	11.00	8.1	11.1	11.1	0.0	100**
2	PBCC-11	11.00	8.1	7.0	7.0	0.0	133
3	PBCC-22	22.00	12.8	12.3	12.3	0.0	92
4	CCK/OFDM-12	12.00	8.4	8.2	9.9	1.7	109
5	CCK/OFDM-24	24.00	13.0	13.8	15.5	1.7	74
6	11a/OFDM-12	12.00	10.5	8.2	9.9	1.7	109
7	11a/OFDM-24	24.00	18.6	13.8	15.5	1.7	74

* 1000 Byte Packets with Preamble, 1 SIFS, 1 CCK-11 ACK with Preamble, 1 DIFS
** Reference range = 100

2.6.1.4 Rate and Throughput It is well known that in packet systems such as IEEE 802.3 and 802.11, the user data rate is smaller than the maximum instantaneous data rate of the transmission system. In the IEEE 802.11 media access control (MAC) protocol, a successful data packet transmission is followed by an acknowledgment packet. This overhead is in addition to the other factors such as guard intervals (so called SIFS and DIFS) and packet preambles and postambles. For reasons of clarity, it is assumed that the acknowledge packets are fixed length at all rates according to Table 2.14.

The throughput of a system is a function of the transmission system, instantaneous rate and packet length. In this section, packets are assumed to be long, 1000 bytes in length; this is an optimistic assumption. In addition, this analysis does not account for other forms of MAC overhead such a the MAC header, data error detection and security such as required for WEP.

Table 2.14 Packet Overhead

Mod	Preamble	Postamble	DIF	ACK*	Total
	μsec	μsec	μsec	μsec	μsec
CCK & PBCC	96	0	50	116	262
CCK/OFDM	108	6	50	116	280
11a/OFDM	20	0	34	40	94

* ACK: Preamble, Data, SIFS

In Table 2.12, the throughput for the various choices are listed. As an example calculation, consider the transmission of 1000 bytes (8000 bits) of data using CCK-11 or PBCC-11. The total transmission time will be $\mathrm{T}_{\mathrm{total}}$ = 262+8000/11 = 989.27 usec yielding a throughput of $R = 8000/\mathrm{T}_{\mathrm{total}}$ = 8.0867 Mbps.

2.6.2 Calculation of Rate versus Range

2.6.2.1 Rate and Range Data The signal to noise ratio calculation can be summarized by the equations that relate transmit power to receive power

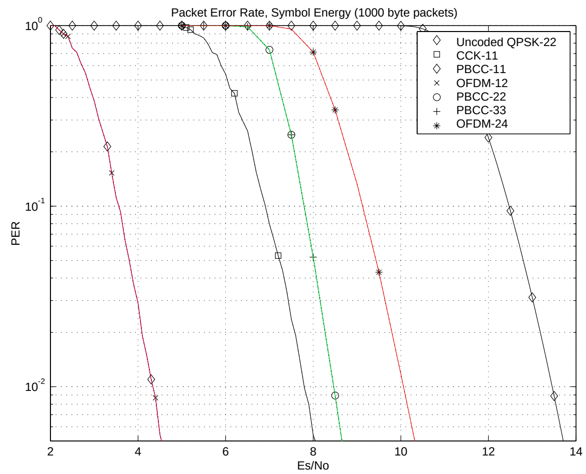


Figure 2.21 Packet Error Rate vs Channel SNR $(\mathrm{E_s/N_o})$

$$\mathrm{P_R} = \frac{\mathrm{PE_oT}}{L(d)}$$

and symbol energy to receive power

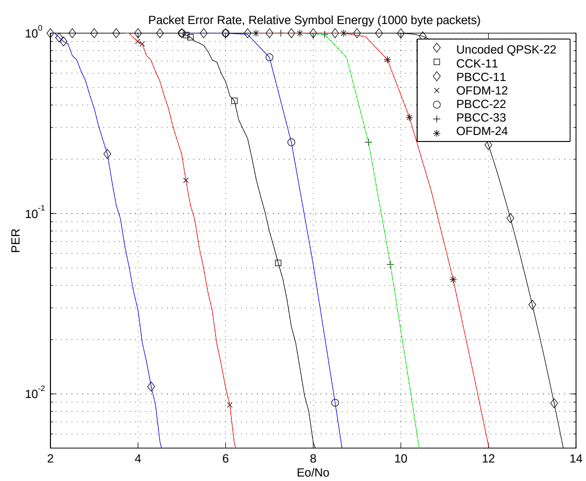


Figure 2.22 Packet Error Rate vs Normalized SNR $(\mathrm{E_o/N_o})$

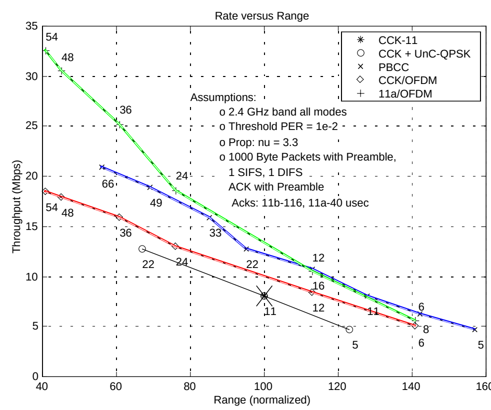


Figure 2.23 Rate versus Range, AWGN

$$\frac{\mathrm{E_s}}{\mathrm{N_o}} = \frac{\mathrm{P_R}\delta\mathrm{E_o}\mathrm{PT_s}\delta\mathrm{E_o}\mathrm{T}}{\mathrm{N_o}}$$

where $\delta\mathrm{E_o}\mathrm{P}$ reflects the power overhead and $\delta\mathrm{E_o}\mathrm{T}$ accounts for symbol clock change relative to the reference (in this section, 11 MHz for CCK-11). For the various systems, Table 2.15 gives the power factors which are the basis of the equation

$$\mathrm{E_s} = \mathrm{E_o}\delta_\mathrm{P}\delta_\mathrm{T},\ \mathrm{E_o} = \mathrm{P_R}\mathrm{T_s}.$$

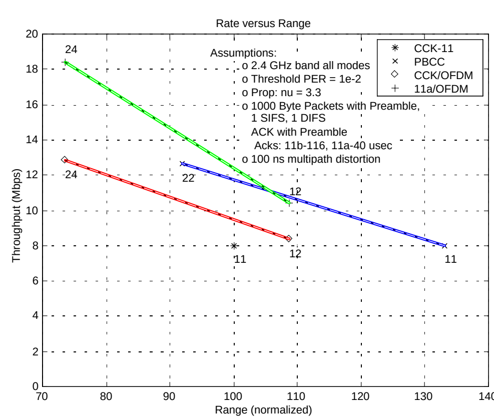


Figure 2.24 Rate versus Range, AWGN + Multipath Distortion

Table 2.15 $\mathrm{E_o}$ **to** $\mathrm{E_s}$ **Translation**

Mod	Rates	δP	δT
1 CCK	all	1 (0 dB)	1 (0 dB)
2 PBCC	{5.5,11,22}	1 (0 dB)	1 (0 dB)
3 PBCC	{8.25,16.5,33,49.5,66}	1 (0 dB)	22/33 (-1.76 dB)
4 OFDM	all	48/65 (-1.32 dB)	11/12 (-.38 dB)

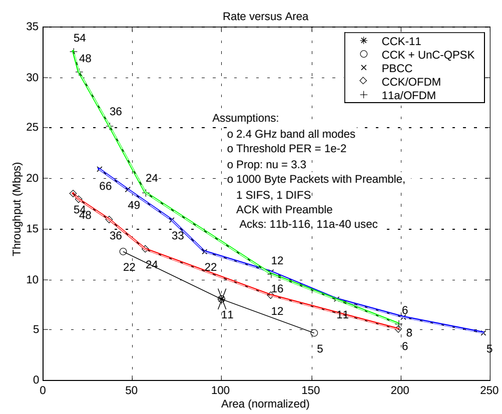


Figure 2.25 Rate versus Area, AWGN

The power overhead $\delta_{\mathrm{P}} \leq 1$, always bounded by 1, has the effect of reducing the symbol energy available for detection from the power received by the radio. The symbol clock parameter, δ_{T} is the ratio of the symbol periods (or symbol frequencies) relative to the base, in this case 11 MHz (*i.e.*, the symbol rate of CCK-11). In this section, $\delta_{\mathrm{T}} \leq 1$ since the symbol rates considered are 11 MHz, 12 MHz and 16.5

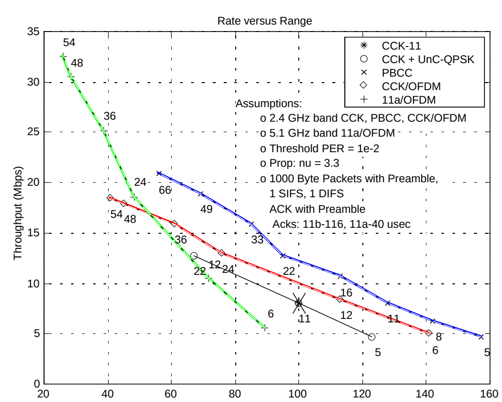


Figure 2.26 Rate versus Range, AWGN w/ 5GHz data

MHz. In Figure 2.21, selected E_s/N_o curves are displayed. These curves show that with this notion of SNR, the PBCC-11 and OFDM-12 systems follow the same curve and have a significant coding gain, about 3.5 dB at a PER of 10^{-2}, when compared to CCK-11. Similarly, the PBCC-22 and PBCC-33 curves are identical on this graph, requiring a fraction of a dB of additional SNR when compared to CCK-11. When one accounts for power overhead and clocking rate differences, one obtains the graph shown in Figure 2.22. On this scale, PBCC-33 moves 1.76 dB to the right due to the higher symbol clock frequency, OFDM-12 and OFDM-24 move 1.70 dB to the right due to the power overhead and clock difference.

The rate and range data for all modes considered in this section is presented in Table 2.12 for AWGN. In Table 2.13, data for channels with 100 nsec multipath distortion, generated via the IEEE 802.11 multipath model [9], is presented. This data is displayed in Figure 2.23 and Figure 2.24. In Figure 2.25, the throughput versus area coverage is shown. Figure 2.26, the throughput versus range in comparison to 5 GHz 802.11a is shown.

These graphs show the superiority of the PBCC based systems over the CCK/OFDM ones. For the highest mandatory rate, PBCC-22 achieves a throughput of 12.8 Mbps at a range that is 95% of the CCK-11 system while the CCK/OFDM-24 achieves 13.0 Mbps at a range that is 76% in AWGN. In terms of area, these factors are 90% and 58% coverage, respectively. With 100 ns of multipath, the range numbers become 92% and 74%.

It is interesting to consider the case of an 11a/OFDM signal if it were used in the 2.4 GHz band; such a hypothetical system would not be backward compatible with 802.11b. However, this modulation would not suffer from the large overhead required to be backword compatible with the 11b preamble; it has the same ranges as CCK/OFDM but much higher throughput. For example, for 11a/OFDM-24, the throughput is 18.6 Mbps while CCK/OFDM-24 has only a 13.0 Mbps throughput. The curves for PBCC and 2.4 GHz 11a/OFDM, shown in Figure 2.23, demonstrate that for ranges up to 60% of the CCK-11 range, the two schemes are very competitive, while the CCK/OFDM system significantly lags both solutions in all cases.

Furthermore, IEEE 802.11a systems are designed for the 5.2 GHz (and higher) U-NII bands (these bands include the two indoor bands 5.15-5.25 GHz band and 5.25-5.35 GHz band and an outdoor band 5.725-5.825 GHz in the USA). The loss model given in Equation (2.10) shows a penalty of over 4.3 (= 6.3 dB) in received signal power due to the higher frequency (*i.e.*, shorter wavelength). It is this factor that moves the 11a curves in Figure 2.23 to the left in Figure 2.26. This shows why 5.2 GHz systems have range problems compared to 2.4 GHz systems at the same power levels and throughput.

2.7 CONCLUSIONS

This chapter considers the history, development and future of high speed wireless Ethernet in the 2.4 GHz ISM band. Networks that allow users to connect to networks without wires and with high throughput have recently become popular and show the potential for exponential growth in the coming years.

The birth of wireless Ethernet began over a decade ago with the work of the IEEE 802.11 wireless networking standards body. This group developed the technology behind the very successful IEEE 802.11b standard that has shown explosive growth over the last couple of years.

This chapter considers the origins of the “11b” standard and includes an introduction to the media access control technology including a description of the MAC header structure. The chapter describes the physical layer technology specified in the 11b standard including the CCK and PBCC modes. An extension of the 11b technology developed by Alantro Communications (now a part of Texas Instruments) is described; this extension provides a “double the data rate” (22 Mbps) mode that is fully backward compatible with existing 11b networks.

The chapter also discusses the role and limitations of spread spectrum communications in wireless Ethernet.

A comparison in terms of range versus rate is given. The comparison includes Intersil’s CCK/OFDM modulation as well as the 802.11a standard in the 5 GHz bands.

Currently, Texas Instruments is shipping wireless Ethernet chips that fully implement the 11b standard, with both CCK and PBCC modes, and include the PBCC-22 extension.

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