Low-Complexity Adaptive Transmission for Cognitive Radios in Dynamic Spectrum Access Networks

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***Abstract*—Cognitive radios that are employed in a network with dynamic frequency assignments must operate efficiently in the presence of uncertainties and variations in the propagation characteristics of the network's communication links. A low-complexity adaptive transmission protocol is described and evaluated for use in cognitive radio networks whose links have unknown and possibly time-varying propagation losses as a result of such phenomena as slow fading or variations in shadowing. The cognitive radios are required to derive only simple statistics in the receivers in order to provide the information that is needed by our protocol; no estimates or measurements of received power or channel gain are used. The protocol's primary mechanism for responding to changes in propagation loss is to adjust the modulation and coding. Because of disruptions that can be caused by higher levels of interference to other radios in the network, the transmitter power is increased only if the most powerful combination of coding and modulation is inadequate. We employ finite-state Markov models for slowly varying channels, and we demonstrate that for such channels our protocol performs nearly as well as an ideal protocol that is told the exact value of the propagation loss for each packet transmission. Thus, the additional complexity that is required to enable cognitive radios to obtain precise channel-gain estimates is not justified and would lead to only negligible improvement in throughput. The throughput of our adaptive transmission protocol is compared with an upper bound that is derived from information theory for a hypothetical ideal protocol that is given perfect channel-state information, and some preliminary results on learning the adaptation decision intervals are included.**

***Index Terms*—Adaptive modulation, adaptive coding, cognitive radio, dynamic spectrum access, packet radio.**

I. Introduction

COGNITIVE radios [1]–[3] are ideally suited for use in dynamic spectrum access networks in which there may be large variations in channel conditions from one session to the next. Such variations are common in networks that operate in a fixed frequency band, but the variations are more severe if the frequency band is changed for consecutive sessions. Each radio in a dynamic spectrum access network must be aware of its communication environment, and it must provide the information that other radios need in order to communicate with it efficiently. This information should be simple, easy to derive, and easy to send to neighboring radios. We focus on the information needed to adapt the error-control coding, modulation, and transmitter power for half-duplex packet transmissions.

A new session begins when one radio, referred to as the *source*, has a collection of packets to send to another radio, the *destination*. At the start of a new session, which may be in a different frequency band than the previous session, the protocol must adjust the transmitter power to provide reliable communications with minimal energy consumption and minimal interference to other radios. As the session progresses, the protocol must adjust the transmissions to compensate for changes in channel conditions. Our results demonstrate the extent to which the adaptive transmission protocol can rely only on code-rate adaptation to offset increased propagation loss. We show that when the cognitive radios must compensate for very large variations in the channel conditions, it is necessary to adapt the modulation also. Power increases are employed by our protocol only if the channel deteriorates so much during a session that changes in coding and modulation cannot provide enough compensation, which occurs very rarely.

The throughput performance of our protocol is compared with theoretical limits that are derived from considering ideal protocols and applying Shannon capacity results. We demonstrate that our protocol performs nearly as well as an ideal protocol that is given perfect channel-state information. The modulation formats that are available to the adaptive transmission system span a range from bandwidth-efficient modulation to power-efficient modulation, which permits the adaptation to compensate for large changes in propagation loss. In order to keep the complexity low, bit-interleaved coded modulation [4] is employed to obtain each combination of coding and modulation that is used by the adaptive transmission protocol. In this approach, it is simple to change the code and modulation independently, because the error-control codes are not tailored to specific modulation formats.

II. Adaptation Statistics

In previous research on adaptive transmission, it is common that the radios are assumed to have full-duplex transmission capability so that channel-state information can be sent on a feedback link at the same time that data transmission is taking place on the forward link. Many of the previously published adaptive transmission protocols rely on the availability of perfect channel-state information (e.g., [5]–[7]). In

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other contributions, pilot or training symbols are added to the transmissions to permit the estimation of the channel state, as mentioned in [7] and employed in several IEEE standards. Many authors either assume the existence of perfect knowledge of the channel gain (i.e., propagation loss) on the forward link or they rely on estimates of the channel gain that are sent to the transmitter on a feedback link, perhaps using a feedback model that incorporates estimation error and delay (see Chapter 9 of [8] and the references cited therein). In contrast, our adaptive transmission protocol does not require channel-state information to be supplied by an external source nor does it require estimation of the channel gain or the received power, insertion of pilot or training symbols, or full-duplex transmission.

We believe that a cognitive radio should rely on the demodulator and decoder to tell the adaptive transmission protocol if changes are needed in coding and modulation and what the changes should be. After all, the acceptability of the code-modulation combination is determined by what happens in the demodulator and decoder, not what happens on the channel. No matter how bad the channel may be, more powerful modulation and coding are unnecessary if an acceptable error probability is provided at the decoder output. The simple statistics used by our protocol are easy to derive in the receiver's demodulator and decoder, they provide reliable assessments of the receiver's performance, and they can be communicated to the transmitter by sending only a few bits in each acknowledgment packet.

For adaptation of coding and modulation, our protocol can use an error count or iteration count, each of which can be obtained easily for each packet that is decoded correctly. A high rate CRC code is used to verify the correctness of the decoded packets. The *error count* for a packet that decodes correctly is the number of binary symbol errors at the output of the demodulator. One way to obtain the error count is to encode the information bits from a correctly decoded packet and compare them with the binary representations of the demodulator hard decisions for the packet. The intent is to count the number of binary symbol errors that would have occurred if there were no decoder in the receiver. The *iteration count* is the average number of decoder iterations per packet among the packets that decode correctly. The decoder can easily report the number of iterations that it made for a packet that has been decoded and verified for correctness. A demodulator statistic is used by our protocol for certain secondary modes of operation, such as the power-adjustment phase at the start of a session that will use a new frequency band. The protocol also uses a demodulator statistic to adapt the power if further adaptation of coding and modulation cannot offset an increase that has occurred in the propagation loss.

The choice of demodulator statistic depends on the modulation format. For QPSK and 16-QAM, we use the *distance statistic*, which is the distance between each received point and its closest point in the signal constellation, averaged over all modulation symbols in the packet. For biorthogonal modulation, we use the *ratio statistic*, which is the second largest correlator output magnitude divided by the largest, averaged over the modulation symbols in the packet. The ratio statistic for an individual symbol was originally introduced for anti-jam communications by Viterbi [9], and it has been employed for many purposes, including soft-decision decoding [10]. The average of the ratio statistics for the symbols in a packet can be used for any modulation format, but for adaptive transmission it typically works best for signal sets of moderate to large size, such as nonbinary orthogonal and biorthogonal modulation or QAM.

Each statistic is used in a simple interval test to decide on the parameters to be used in the next transmission. The test may be applied at the destination or the source; in either case, only a few bits of information are needed in each acknowledgment packet. For half-duplex packet transmission, there is some delay between the determination of the adaptation statistics and the transmission of the packet for which they are applied; however, in an ad hoc network, we believe the primary purpose of adaptive transmission is to compensate for slowly-varying channel conditions, a belief that is supported by the conclusions in [11]. The delay for our protocol is no larger than for any other protocol that accommodates half-duplex packet transmission, including protocols that rely on estimates of the channel gain.

III. Code-Modulation Combinations

The set of modulation formats available to the adaptive transmission protocol is denoted by $\{\mathcal{M}_j : 1 \le j \le n_m\}$. The set of error-control codes is denoted by $\{C_i : 1 \le i \le n_c\}$. Code C_i has rate r_i, and the codes are indexed in order of increasing rate: $r_1 < r_2 < \ldots < r_{n_c}$. The set $\{D_k : 1 \le k \le n\}$ of code-modulation combinations employed by the protocol is indexed with a single subscript whose maximum value is $n \le n_c n_m$. The combinations are indexed in order of increasing information rate. It is common that some of the $n_c n_m$ possible combinations of coding and modulation are not included in the set. For our numerical results, the modulation set includes 16-QAM ($\mathcal{M}_1$), QPSK ($\mathcal{M}_2$), and a third type of modulation ($\mathcal{M}_3$) that uses biorthogonal signals. The *modulation chip* is the basic pulse used for the data-modulated waveform. For QAM or QPSK, each modulation symbol is a modulation chip, so one modulation chip represents four binary digits for 16-QAM and two binary digits for QPSK.

Modulation $\mathcal{M}_3$ is derived from a set of 64 biorthogonal signals. In one construction method, we begin with set of 32 orthogonal signals with 32 chips per signal, which can be obtained from the rows of a 32×32 Hadamard matrix as discussed in [10]. These 32 signals and their negatives give a 64-biorthogonal set [12] which we denote by $\mathcal{B}$. If the inphase and quadrature components of the signals in set $\mathcal{M}_3$ each have alphabet $\mathcal{B}$, then $\mathcal{M}_3 = \mathcal{B} \times \mathcal{B}$ is the signal set for 4096-inphase-quadrature biorthogonal (4096-IQB) modulation, which has 32 quaternary modulation chips per symbol. Although 4096-IQB modulation uses a large signal set, its inphase and quadrature components are modulated and demodulated separately as 64-biorthogonal modulation, which can be demodulated efficiently using matrix multiplication (e.g., see Appendix E of [13]), perhaps with fast transform methods.

For our numerical results, the code set consists of four turbo product codes that can be decoded using available hardware [14]. These turbo product codes are employed in many applications; for example, some are among the optional codes for IEEE 802.16 (e.g., see [15]). The four codes that we chose for illustration of the performance of the adaptive transmission protocol have rates $r_1 = 0.236$, $r_2 = 0.325$, $r_3 = 0.495$, and $r_4 = 0.793$. Each packet represents 4096 binary symbols, which is the block length for codes C_2, C_3, and C_4. The block length of C_1 is 2048, so there are two code words per packet when C_1 is used. For the binary code symbols within a packet, S-random interleaving [16] is employed for QAM, but helical interleaving [17] is used for the other two modulation formats. The log-likelihood-ratio (LLR) metric [18] is used for soft-decision decoding with QPSK and 4096-IQB; however, for QAM, we use a simpler distance metric [19] that gives approximately same performance for QAM as the LLR metric. For comparisons with the performance of the turbo product codes, we also include results for four hypothetical capacity-achieving codes of the same rates. A typical session requires the delivery of several hundred to a few thousand packets. For example, if each packet represents 4096 binary code symbols, then a 1 MB file transfer requires approximately 2500 packets for the high-rate code and approximately 8500 packets for the low-rate rate code.

In our model, once a frequency band has been selected for a session, the bandwidth is fixed for the source's transmissions during the session. As a result, the chip rate is fixed, because the chip rate determines the spectral occupancy of the transmitted signal. This is one reason for using the modulation chip as the fundamental pulse, rather than the modulation symbol or the binary code symbol. Each modulation format requires a different number of chips to represent 4096 binary code symbols: 16-QAM ($\mathcal{M}_1$) requires 1024 chips, QPSK ($\mathcal{M}_2$) requires 2048 chips, and 4096-IQB ($\mathcal{M}_3$) requires $32\lceil 4096/12 \rceil = 10{,}944$ chips. It follows that the number of chips per packet and the packet duration change as the modulation is adapted.

Because the bandwidth is constant, the most useful measure of signal energy is the average energy per chip, denoted by $\mathcal{E}_c$. Other measures include the average energy per modulation symbol ($\mathcal{E}$), the average energy per binary code symbol ($\mathcal{E}_s$), and the average energy per information bit ($\mathcal{E}_b$). An important reason for using the modulation chip as the fundamental pulse is the invariance of the average energy per chip during a change in modulation or coding for which the average power is constant. For such a change, $\mathcal{E}_c$ is constant, but $\mathcal{E}$, $\mathcal{E}_s$, and $\mathcal{E}_b$ may not be. For example, if we change from 16-QAM to QPSK but use the same average power and the same code, then $\mathcal{E}_s$ and $\mathcal{E}_b$ are doubled. If the thermal noise has one-sided power spectral density N_0, then the chip energy to noise density ratio is $\mathcal{E}_c/N_0$ and we define $\mathrm{CENR} = 10\log_{10}(\mathcal{E}_c/N_0)$, which is measured in decibels (dB). For protocols that include adaptive modulation, it is better to give the performance results in terms of $\mathcal{E}_c/N_0$ or CENR rather than $\mathcal{E}_b/N_0$, the bit energy to noise density ratio, or its corresponding performance measure in dB, $\mathrm{ENR} = 10\log_{10}(\mathcal{E}_b/N_0)$.

IV. Shannon Capacity

We employ results from information theory to assess the potential benefits that can be obtained from adaptation of coding and modulation and to determine how efficient our protocols are in responding to changes in the communication environment. It is well known that for reliable communication on the additive white Gaussian noise channel, the minimum value of $\mathcal{E}_b/N_0$ is $\ln 2$, which corresponds to approximately -1.6 dB as the minimum value of ENR. It is also well known that as $K \to \infty$, we can achieve an arbitrarily small error probability with K-biorthogonal signals provided that $\mathcal{E}_b/N_0 > \ln 2$, but the rate and bandwidth efficiency go to zero and the demodulator complexity becomes infinite. Rather than permit the rate to decrease or the modulation to become more complex as we take the limit, we examine the capacity for binary codes of fixed rate r and a fixed modulation format $\mathcal{M}$.

For our capacity analysis, we must determine the minimum value of $\mathcal{E}_c/N_0$ that permits reliable communication with binary codes of rate r for each modulation format that is available to the adaptive transmission system. The capacity equations are given for the two modulation formats that have equal-energy signals, but the numerical values for the capacity limits are given for all three modulation formats. Suppose the code rate is r bits per binary symbol, the modulation format has $M = 2^m$ symbols, the number of modulation chips per modulation symbol is L, and the average energy per modulation symbol is $\mathcal{E}$, then the average energy per chip is

$$\mathcal{E}_c = \mathcal{E}/L = rm\mathcal{E}_b/L. \quad (1)$$

Let $\Lambda_b(r,\mathcal{M})$ be the minimum value of $\mathcal{E}_b/N_0$ that permits reliable communication with binary codes of rate r and modulation format $\mathcal{M}$. We first evaluate $\Lambda_b(r,\mathcal{M})$ and then convert it to $\Lambda_c(r,\mathcal{M})$, the corresponding minimum value of $\mathcal{E}_c/N_0$ that provides reliable communication. For each modulation format $\mathcal{M}$, a *capacity-achieving code* of rate r is an error-control code of rate r that provides a negligibly small probability of error when it is used to transmit packets using modulation $\mathcal{M}$ on a channel for which $\mathcal{E}_c/N_0 > \Lambda_c(r,\mathcal{M})$, but its packet error probability is one if $\mathcal{E}_c/N_0 < \Lambda_c(r,\mathcal{M})$.

For each modulation format $\mathcal{M}$, we first obtain a capacity equation of the form

$$C = \Gamma(\alpha) = \Gamma\left(\sqrt{r\mathcal{E}_b/N_0}\right), \quad (2)$$

where $\alpha = \sqrt{r\mathcal{E}_b/N_0}$. For operation at the capacity limit with a code of rate r using modulation format $\mathcal{M}$, we set $r = C$ and $\mathcal{E}_b/N_0 = \Lambda_b(r,\mathcal{M})$. It follows from (2) that

$$r = \Gamma\left(\sqrt{r\Lambda_b(r,\mathcal{M})}\right). \quad (3)$$

Numerical methods are applied to solve (3) for $\Lambda_b(r,\mathcal{M})$ to find the lower bound on $\mathcal{E}_b/N_0$ for modulation format $\mathcal{M}$ and binary codes of rate r.

The capacity equation for any binary antipodal modulation format (e.g., BPSK) with coherent demodulation is

$$C = 1 - \int_{-\infty}^{\infty} \frac{\exp\{-(u-\alpha)^2\}}{\sqrt{\pi}\ln 2} \ln[1+\exp\{-4\alpha u\}]\, du. \quad (4)$$

From the solution to this equation, we obtain the capacity limit $\Lambda_b(r,\mathcal{M}_2)$ for both BPSK and QPSK, because their capacities are identical when expressed in terms of $\mathcal{E}_b/N_0$ or ENR. This is not true for the capacity limit on $\mathcal{E}_c/N_0$, however, because

TABLE I
CAPACITY LIMITS Λ_c IN dB FOR THREE MODULATION FORMATS AND FOUR CODE RATES.

Code Rate	4096-IQB	QPSK	16-QAM
0.2	−8.83	−4.95	−0.51
0.4	−6.78	−1.22	3.76
0.6	−5.30	1.46	6.95
0.8	−3.79	4.04	10.21

BPSK has one binary digit per chip and QPSK has two. In the conversion of the result of (3) to the capacity limit for $\mathcal{E}_c/N_0$, we apply (1) with $L=1$ for BPSK and QPSK, $m=1$ for BPSK, and $m=2$ for QPSK.

The capacity equation for K-biorthogonal modulation with $L=K/2$ is

$$C = 1-\int_{\mathbb{R}^L} \frac{L^{-1}}{\pi^{L/2}} \exp\left\{-\beta^2-\sum_{j=0}^{L-1} u_j^2\right\} \times \sum_{i=0}^{L-1} \exp(2\beta u_i) \log_2\left[1+\frac{\sum_{j=0}^{L-1}\exp(-2\beta u_j)}{\sum_{j=0}^{L-1}\exp(2\beta u_j)}\right] d\mathbf{u}, \quad (5)$$

where $\beta=\sqrt{\mathcal{E}/N_0}=\alpha\sqrt{\log_2 K}$ and $\mathbf{u}=(u_0,u_1,\ldots,u_{L-1})$. The solution of the capacity equation gives the capacity limit $\Lambda_b(r,\mathcal{B})$, where $\mathcal{B}$ is K-biorthogonal modulation. If $M=K^2$, then $\Lambda_b(r,\mathcal{B})$ is also the capacity limit for M-IQB modulation, and if $K=64$ and $\mathcal{B}$ is the 64-biorthogonal modulation format defined in Section III, then $\mathcal{M}_3=\mathcal{B}\times\mathcal{B}$ is 4096-IQB modulation and $\Lambda_b(r,\mathcal{M}_3)=\Lambda_b(r,\mathcal{B})$.

V. POTENTIAL GAINS FROM CODE AND MODULATION ADAPTATION

Before we examine a specific protocol for adaptive modulation and coding, it is worthwhile to determine what we can gain by applying such a protocol. Suppose that a session is in progress and the shadow loss increases by several dB. If the transmitter power is increased by an equal amount to offset the additional propagation loss, then the interference to unintended receivers is increased by several dB, which may cause some ongoing sessions to be disrupted or even terminated. On the other hand, if we hold the transmitter power constant and change the modulation or coding, then the interference to unintended receivers is not increased. How much increase in propagation loss can be offset by changing only the coding compared with changing the combination of coding and modulation? For our first answer to this question, we turn to information theory.

In Table I we list the capacity limits in dB as a function of code rate for 4096-IQB modulation, QPSK, and 16-QAM. Suppose that an adaptive transmission protocol can use four capacity-achieving codes that have the rates listed in Table I, and suppose it can use any of the three modulation formats listed there. The protocol cannot compensate for an increase in propagation loss of more than 11 dB by changing only the code rate. However, if the source is using 16-QAM with code rate 0.8, then the protocol can compensate for more than 19 dB by having the source change to 4096-IQB modulation with code rate 0.2, yet there is no increase in the interference caused to other sessions. On the other hand, if the protocol were to compensate by increasing the source's transmitter power by 19 dB, the additional interference is likely to disrupt several ongoing sessions in the network.

TABLE II
CAPACITY AND PERFORMANCE FOR TEN COMBINATIONS OF TURBO PRODUCT CODES AND MODULATION FORMATS.

Combination	Modulation	Code, r	Λ_c	CENR
D_1	4096-IQB	C_1, 0.236	−8.2 dB	−6.8 dB
D_2	4096-IQB	C_2, 0.325	−7.5 dB	−6.3 dB
D_3	4096-IQB	C_3, 0.495	−6.1 dB	−5.0 dB
D_4	4096-IQB	C_4, 0.793	−3.9 dB	−3.1 dB
D_5	QPSK	C_1, 0.236	−4.1 dB	−1.7 dB
D_6	QPSK	C_2, 0.325	−2.4 dB	−0.8 dB
D_7	QPSK	C_3, 0.495	0.1 dB	1.7 dB
D_8	QPSK	C_4, 0.793	4.0 dB	4.9 dB
D_9	16-QAM	C_3, 0.495	5.3 dB	7.2 dB
D_{10}	16-QAM	C_4, 0.793	10.0 dB	11.2 dB

As far as the protocol is concerned, it is straightforward to obtain even more gain from adaptive coding and modulation by including additional modulation formats among the choices. For example, if we add 64-biorthogonal modulation to the set of modulation formats, then the protocol can compensate for more than 22 dB of increased propagation loss. If both 64-biorthogonal modulation and 64-QAM are added to the set, then it can compensate for more than 27 dB. Although each of these maximum compensations is for a set of four capacity-achieving codes, the maximum compensation for the four turbo product codes is almost as large. From Table II we see that the maximum compensation for the four turbo product codes is 18 dB, which is only 1 dB less than for the capacity-achieving codes in Table I. If we use capacity-achieving codes of the four rates in Table II instead of the four rates in Table I, then the maximum compensation is 18.2 dB, which is only 0.2 dB more than is available with the turbo product codes. Of course, the values of CENR are higher for the turbo product codes than for the capacity-achieving codes, but only by 1–2 dB for most rates and modulation formats.

The code-modulation combinations $\{D_k : 1 \le k \le 10\}$ in Table II are listed in order of increasing information rate. These ten combinations are the only ones that are used by our adaptive transmission protocol. Notice that two of the available code-modulation combinations are not listed in Table II. We omitted 16-QAM with codes C_1 and C_2 because these combinations provide little if any benefit to the adaptive transmission protocol. Additional discussion of the selection of the code-modulation combinations for use by the protocol is given in Section VII.

VI. INITIAL POWER ADJUSTMENT

When a new session begins, the source's transmitter power may be much higher than necessary or much lower than required for reliable communication, especially if the session's frequency band has not been used recently by the source and destination. One problem is that the propagation loss is very difficult to predict accurately. This is especially true for shadow loss, which can vary by more than 20 dB in urban environments [20]. The empirical formulas for outdoor propagation (e.g., see [8] and [21]) typically include modifications and correction factors that depend on the terrain and other features that are unknown to the source. Variations in correction factors of several dB are not uncommon [22] for different environments. A typical model for indoor communications

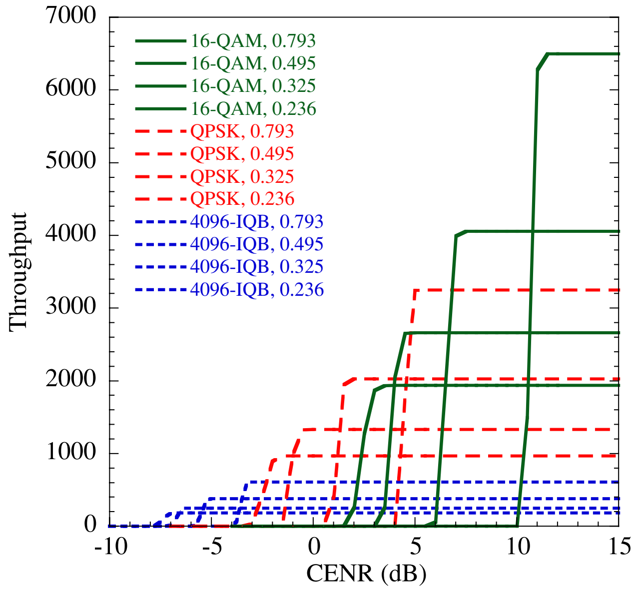

Fig. 1. Throughput for each of 12 code-modulation combinations.

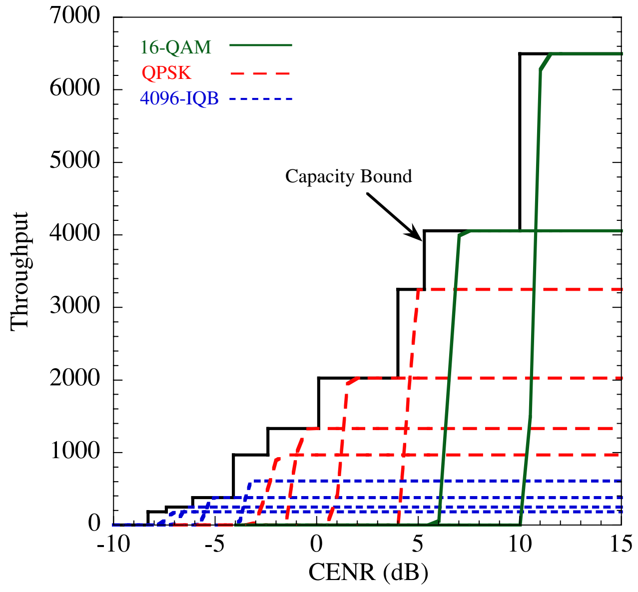

Fig. 2. Throughput for each of ten combinations and the corresponding capacity limit for the adaptive transmission protocol.

includes a Gaussian random variable with a standard deviation as large as 10 dB for some buildings [21].

It is necessary to adjust the transmitter power as quickly as possible after the session is underway to obtain an acceptable packet error probability without causing interference to other sessions in the network. For our protocol, each session begins with the code-modulation combination that has the highest information rate. For the combinations in Table II, this is 16-QAM with code C_4. The power-adjustment protocol is typically needed for only the first six to eight packets of a session. Recall that for a 1 MB file transfer, 2500–8500 packets must be delivered to the destination, so the power-adjustment phase of a session has negligible effect on the throughput or completion time of the session. During the power-adjustment phase, a receiver statistic is included in each acknowledgment packet. A simple interval test is performed on the statistic to determine the power for the next packet. If the initial power is too low, it may be that the destination is not even aware that a packet was sent. Thus, for each unacknowledged packet that is sent during the power-adjustment period, the source automatically increases the power by a fixed amount (e.g., 5 dB). The termination of the power-adjustment phase is determined by a stopping condition that is applied to the demodulator statistics for the initial sequence of packets in the session. A full description of the power-adjustment protocol is given in the Appendix. Once the power-adjustment phase is completed, the adaptive transmission protocol takes over and compensates for changes that might occur in the channel during the remainder of the session. In general, the response to a deterioration in channel conditions is to switch to a more robust, lower-rate, code-modulation combination. An improvement in channel conditions is exploited by switching to a code-modulation combination that provides a higher information rate.

VII. Selection of the Code-Modulation Combinations

Because the chip rate is held constant as the modulation is changed, the packet durations are different for different modulation formats. Therefore, it is necessary to have a standard unit of time for comparisons of the throughput for different modulation formats. We define the standard *time unit* to be the packet duration for QPSK, which is the time required to transmit 2048 chips at the fixed chip rate. The *throughput* is defined to be the number of information bits delivered to the destination per unit of transmission time. An information bit is said to be *delivered* to the destination if it is part of a packet that is decoded correctly by the destination. Even if a bit is decoded correctly, it does not contribute to the throughput unless all the information bits in its packet are decoded correctly. If one or more information bits are not decoded correctly, the entire packet is retransmitted. Although the information bits in an erroneous packet do not contribute to the throughput, the time required to transmit such a packet is included in the total transmission time for a session. Thus, the system is penalized for unsuccessful packets.

In Fig. 1, the throughput is given as a function of CENR for each of the 12 code-modulation combinations that can be obtained from the four turbo product codes and the three modulation formats. Notice that the throughput curve for 16-QAM and $r=0.236$ (C_1) is dominated by the throughput curve for QPSK and $r=0.495$ (C_3). If the goal is to maximize throughput, one would never choose the code-modulation combination (16-QAM,C_1), because, for each value of CENR, at least one other combination gives a higher throughput. Although the throughput for 16-QAM and $r=0.325$ (C_2) is not dominated by a single combination, its throughput curve is the highest of the 12 curves for only a very small range of CENR, so it would be of almost no benefit to the system. As a result, we do not include the combinations (16-QAM,C_1) and (16-QAM,C_2) in the code-modulation set $\{D_k : 1 \le k \le 10\}$ that is

TABLE III
ENDPOINTS FOR THE DESIGN INTERVALS USED TO OBTAIN THE PERFORMANCE RESULTS IN SECTIONS IX AND X.

Combination	γ_1(EC)	γ_2(EC)	γ_1(IC)	γ_2(IC)
D_1	746	4096	4	32
D_2	493	744	3	12
D_3	165	491	3	15
D_4	61	163	3	10
D_5	746	899	4	15
D_6	493	744	3	12
D_7	165	491	3	18
D_8	59	163	3	15
D_9	165	499	3	20
D_{10}	0	163	0	20

used by the adaptive transmission protocol. The throughput for each of the ten combinations is shown in Fig. 2 along with the capacity bound. For each value of CENR, the capacity bound represents the maximum throughput that can be obtained from any of the ten code-modulation combinations that use capacity-achieving codes with the same rates as the turbo product codes. In the capacity curve, we have omitted 16-QAM for both $r = 0.236$ and $r = 0.325$, just as we did for the turbo product codes.

VIII. THE ADAPTIVE TRANSMISSION PROTOCOL

After the initial power adjustment is complete, the adaptive transmission protocol governs the choice of modulation and error-control coding. The receiver statistic that is employed for adaptive coding and modulation is either the error count or the iteration count; the two are approximately equally effective. For each packet that decodes correctly, the protocol performs an interval test to determine which code-modulation combination to use for the next packet. For each packet that does not decode correctly, the next packet is sent with the next lower-rate code-modulation combination than was used for the failed packet. If the failed packet used the lowest-rate code-modulation combination, then the next packet uses the same combination again. After several consecutive failures with the combination of lowest rate, the protocol may be required to increase the transmitter power if that is permitted by the spectrum access protocol.

For each adaptation statistic, each code-modulation combination is associated with three *decision intervals*, $\mathcal{I}_{-1} = (-\infty, \gamma_1)$, $\mathcal{I}_0 = [\gamma_1, \gamma_2]$, and $\mathcal{I}_1 = (\gamma_2, \infty)$. The *decision interval endpoints*, γ_1 and γ_2, define completely the decision intervals. The endpoints can be adjusted during the session by a learning process within the adaptive transmission protocol, so we distinguish between the *design intervals*, which are the nominal intervals selected by the system designer, and the *adjusted intervals*, which are derived from what the cognitive radio learns from its past decisions as the session progresses. In Table III we list the endpoints for the design intervals for the modulation formats, turbo product codes, and decoders that are used for the performance results presented in the next two sections. The entries in the kth row of the table are the values of γ_i(EC) and γ_i(IC) that are employed with the error count and iteration count, respectively, to select the code-modulation combination for packet $i+1$ when combination D_k was used for packet i. The maximum number of iterations is 32 for the decoders used for the four turbo product codes [14], and of

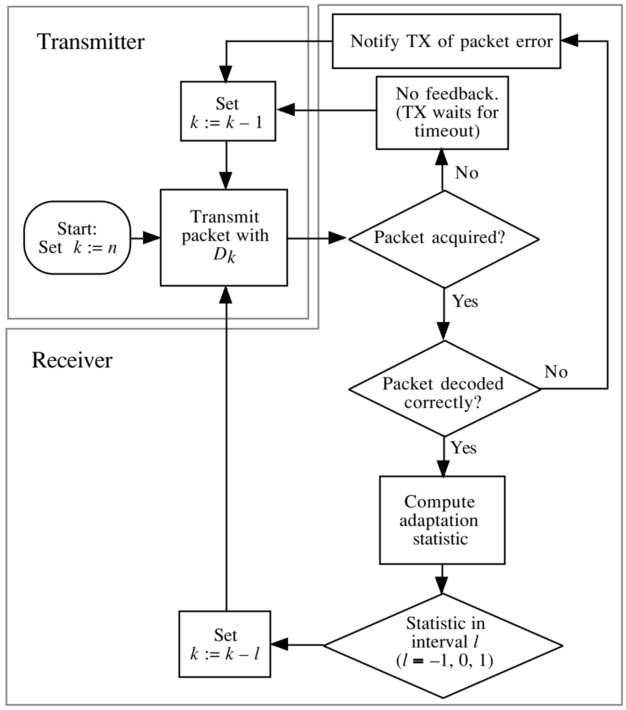

Fig. 3. Flow chart for the adaptive transmission protocol.

course the maximum error count is 4096, the number of binary symbols per packet.

Because the adaptation statistics for the iteration count and the error count decrease as the channel quality increases, the description of the adaptation decision process is the same for the two. Let z_i denote the statistic obtained from the receiver as a result of the reception of the ith packet; that is, z_i is either the error count or the iteration count for the ith packet, depending on which receiver statistic is used for the session. Suppose that code-modulation combination D_k was used for packet i. The code-modulation combination for packet $i+1$ is D_k if $z_i \in \mathcal{I}_0$, D_{k+1} if $z_i \in \mathcal{I}_{-1}$, and D_{k-1} if $z_i \in \mathcal{I}_1$. The interval tests are equivalent to testing $z_i < \gamma_1$ and $z_i > \gamma_2$ (if neither, then $z_i \in \mathcal{I}_0$). The complete operation of the adaptive transmission protocol is illustrated in Fig. 3.

IX. PERFORMANCE FOR STATIC CHANNELS

The decoders for the turbo product codes use iterative decoding methods for which there are no performance analyses, approximations, or bounds that are sufficiently accurate for our needs; therefore, the underlying performance results for the individual turbo product codes are obtained from simulations. These results are combined with analysis in Section X to evaluate the performance of ideal protocols that use the turbo product codes. These analytical results provide bounds on the performance of our protocols. The first performance results for adaptive modulation and coding are given in Fig. 4 for a channel with fixed but unknown propagation loss, so the value of CENR is unknown to the protocol. The throughput curves for the adaptive protocol with the iteration count, the adaptive protocol with the error count, and the ten individual code-modulation combinations are shown in Fig. 4. The upper envelope of the ten curves for

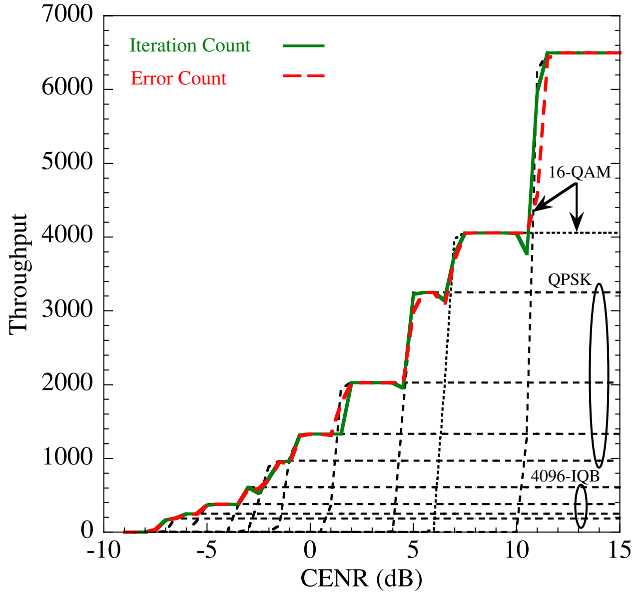

Fig. 4. Throughput for the adaptive transmission protocol on a static channel with unknown propagation loss.

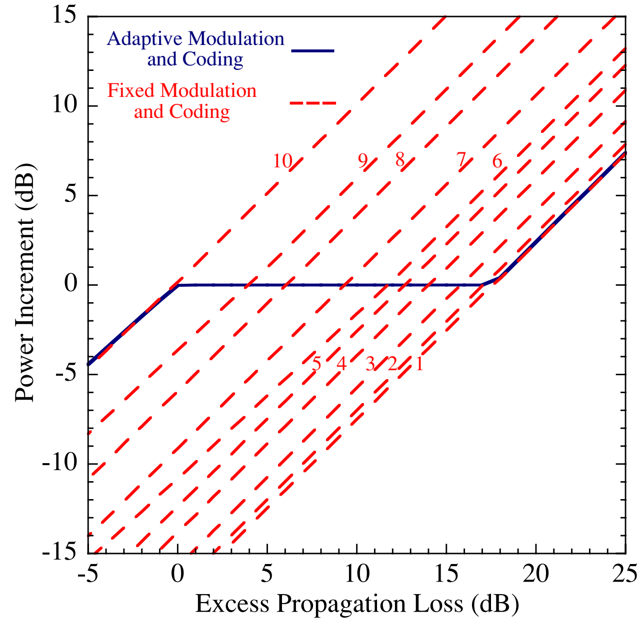

Fig. 5. Response of the adaptive protocol to changes in propagation loss.

the fixed combinations represents the performance of an ideal protocol that is told the exact value of CENR and uses the combination that maximizes the throughput for that value. The performance of our adaptive protocol with either the iteration count or the error count is almost as good as the performance of the ideal protocol that is provided with perfect channel-state information.

Results such as those in Figs. 2 and 4 are very useful in the design and evaluation of adaptive systems for dynamic channels, even though they are for static channels. First, the throughput results in Fig. 4 show that the cognitive radio would gain very little if more complex channel estimates were made, because there is very little gap between the throughput curve for the adaptive protocol and the upper envelope of the throughput curves for the individual code-modulation combinations. Second, the results such as those in Fig. 4 give us the steady-state response of the protocol to a change in the channel conditions. Third, such results tell us how much could be gained with more powerful error-correcting codes. The gain that could be obtained by stronger (and therefore more complex) error-control codes is the gap between the upper envelope and the capacity bound in Fig. 2.

The next results demonstrate the performance of our adaptive coding and modulation protocol when used in conjunction with a protocol that adapts the transmitter power only when further changes in modulation and coding cannot accomplish the desired objective. The power is increased only if the most robust code-modulation combination is inadequate; it is decreased only if the code-modulation combination with the highest information rate is stronger than necessary. We employ a session model in which the source must deliver 500 KB of information to the destination over a channel with a fixed but unknown propagation loss, which requires that the destination must decode correctly approximately 1250 – 4250 packets, depending on the code-modulation combination that is used by the source. The *session duration* is defined as the number of time units required to complete the session. The results in Fig. 5 illustrate the power increment as a function of the excess propagation loss for the adaptive coding and modulation protocol. The *power increment* is the amount in dB by which the transmitter's power is increased (or decreased, if it is negative) by the protocol. For comparison, the power increments for ten fixed code-modulation combinations are also shown. After the initial power adjustment has been completed, the adaptive transmission protocol begins operation at the reference point in the curve in Fig. 5, which corresponds to a power increment of 0 dB and an excess propagation loss of 0 dB. These reference values are relative to the transmitter power level and propagation loss at the time that the power-adjustment protocol hits its stopping condition. After that, as the excess power varies between 0 dB and 18 dB, the adaptive transmission protocol follows the horizontal line by changing the coding and modulation but not changing the power (the power increment remains constant at 0 dB). As long as the excess propagation loss does not exceed 18 dB during the session, the adaptive transmission protocol does not increase the transmitter power. As discussed in Section V, the adaptive transmission protocol can compensate for propagation loss increases that are much greater than 18 dB without increasing the transmitter power if 64-biorthogonal modulation and 64-QAM are added to the set of modulation formats. In contrast, we see from Fig. 5 that each fixed code-modulation requires that the power be increased to match each increase in the propagation loss that occurs during the session. It is likely that the amount of power increase that is permitted will be limited in a dynamic spectrum access network because of the disruption that higher-power transmissions may cause to other sessions. If power increases are desired and permitted, the protocol can automatically adapt the power if the most robust code-modulation cannot handle the increase in the propagation loss.

Our protocol's ability to mitigate large variations in propagation loss without increasing power results in variable session

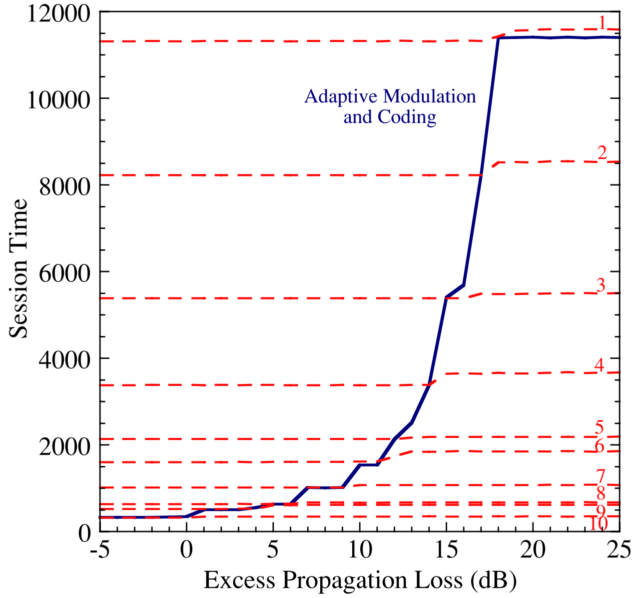

Fig. 6. Session duration for the transmission of 500KB of data.

durations. The results for the session duration are illustrated in Fig. 6. If a fixed combination of modulation and coding is used, then the session durations are constant regardless of the propagation loss. However, in a dynamic spectrum access network, a radio may not be permitted to increase its power by as much as required for a fixed combination, which may require termination of the session if the increase in the propagation loss exceeds the allowed increase in transmitter power. The session durations for our protocol increase if the propagation loss increases, but it is far better to endure an increase in the session duration than to be forced to terminate the session and reschedule it for a later time or in a different frequency band. Furthermore, the largest values for the session duration shown in Fig. 6 for our protocol would occur only if the excess propagation loss increases by more than 18 dB immediately following the power-adjustment period and remains above 18 dB for the entire session, which is an extremely unlikely event.

X. Performance for Dynamic Channels

Next, we turn attention to dynamic channels. The goal of the adaptive transmission protocol is to choose the combination of modulation and coding that provides the largest throughput for each channel state, but the protocol does not know the state of the channel. Instead, the protocol must base its choice on the receiver statistic only. One of our goals is to examine how much improvement might be obtained from a more complex cognitive radio that provides more channel-state information than a cognitive radio that provides only the receiver statistics described in this paper. The tradeoff is that more channel-state information is obtained at a cost of decreased information rate if pilot signals or training symbols are used and a cost of increased complexity in the receiver for the channel measurement and estimation systems. We explore this tradeoff by introducing ideal protocols that have *perfect previous-state information (PPSI)* or *perfect next-state information (PNSI)*. For a given set of codes, the former provides bounds on the best possible performance of protocols that rely only on feedback from the previous packet, and the latter provides bounds on the best possible performance of any protocol with any degree of complexity in the cognitive radios, including hypothetical cognitive radios that can predict the future perfectly. The results for these ideal protocols are compared with the performance of our protocol that derives all its information from the iteration count or the error count, each of which has very low complexity. For static channels, the comparison between our protocol and the ideal PNSI protocol is just the comparison shown in Fig. 4. The performance the PNSI protocol is the upper envelope of the throughput curves for the individual code-modulation combinations.

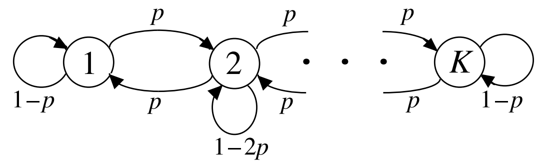

Fig. 7. K-state Markov chain for modeling propagation losses.

The dynamic channels are modeled with finite-state Markov chains of the type shown in Fig. 7. The use of finite-state Markov chains to model slow fading is described in [23]. We believe that Markov chains are useful models to test adaptive protocols that are designed to respond to any type of slow variations in the channel. In other investigations (e.g., [24] and [25]), we have used finite-state Markov models for time-varying multipath interference, time-varying partial-band interference, and Rician fading with a time-varying fading parameter.

For modeling slow fading or changes in shadow loss, the states in the Markov chain correspond to different propagation losses. In our approach, there is a nominal propagation loss that is fixed, but an excess propagation loss is added to the nominal value, and the excess loss is different for different states. To be more precise, we define *excess propagation loss* to be the amount in dB by which the actual propagation loss exceeds some nominal value. For simplicity, we consider excess propagation losses in dB that are multiples of a propagation-loss increment Δ dB. State i of the Markov chain corresponds to excess propagation loss $L_i = (i-1)\Delta$ dB, so the excess propagation loss for state 1 is 0 dB. That is, the propagation loss when the channel is in state 1 is the nominal loss. For most of our results, the state is fixed for the duration of a packet, but it can change from one packet to the next. The value of CENR for state $i = 0$ is denoted by CENR_0, so the value of CENR for a packet that is transmitted when the channel is in state i is $\text{CENR} = \text{CENR}_0 - L_i$. The state transition probability is $p = 0.1$ for all results. We also evaluated the throughput performance of adaptive transmission protocols for $p = 0.2$, and we found the larger transition probability has no effect on any of our conclusions.

From simulation results for the individual code-modulation combinations, such as those in Figs. 1 and 2, we can determine analytically the average throughput for ideal protocols that are given perfect channel-state information. The throughput achieved by combination D_i when the channel is in state k

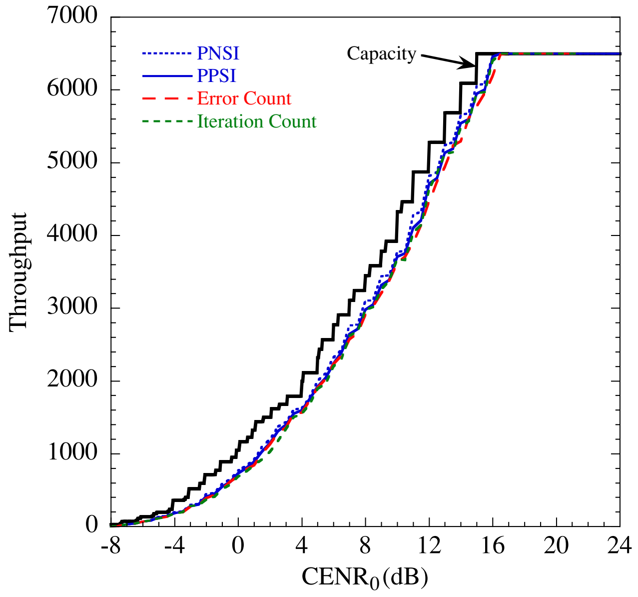

Fig. 8. Throughput for the adaptive transmission protocol with time-varying propagation loss modeled by a six-state Markov chain ($\Delta = 1$ dB).

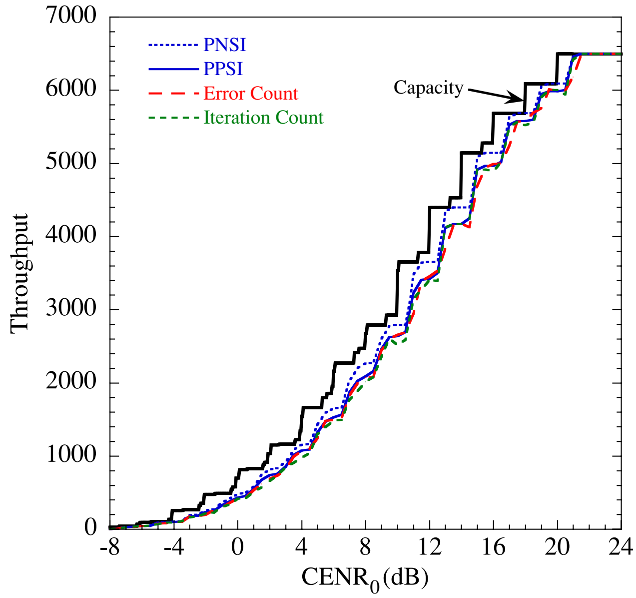

Fig. 9. Throughput for the adaptive transmission protocol with time-varying propagation loss modeled by a six-state Markov chain ($\Delta = 2$ dB).

is denoted by $s(i|k)$. The transition probability $p(k|j)$ is the probability that the next state is k given that the previous state is j. The conditional expected throughput for the ideal protocol with PPSI is

$$\bar{s}(i|j) = \sum_{k=1}^{K} s(i|k)\, p(k|j). \tag{6}$$

When the previous state is j, combination D_{i_j} is selected for the next transmission if

$$\bar{s}(i_j|j) = \max\{\bar{s}(i|j) : 1 \le i \le n\}. \tag{7}$$

If π_j denotes the steady-state probability for state j in the Markov chain, then the average throughput for the ideal protocol with PPSI is

$$\bar{S}_1 = \sum_{j=1}^{K} \pi_j\, \bar{s}(i_j|j). \tag{8}$$

Now, consider the ideal protocol with PNSI. The conditional expected throughput for combination D_i given that the next state is k is $s(i|k)$. When the next state is k, combination D_{i_k} is selected for the next transmission if

$$s(i_k|k) = \max\{s(i|k) : 1 \le i \le n\}, \tag{9}$$

and the average throughput for the ideal protocol with PNSI is

$$\bar{S}_2 = \sum_{k=1}^{K} \pi_k\, s(i_k|k). \tag{10}$$

For the ideal protocol with PNSI, we employ two sets of codes. One is the set of four turbo product codes whose rates are listed in Table II and the other is a set of four capacity-achieving codes that have the same four rates. The analytical performance results for the ideal protocol with PNSI and capacity-achieving codes provide an upper bound on the throughput for any protocol that uses any four codes with those rates. The four turbo product codes are also employed with the ideal protocol that has PPSI to give a more realistic benchmark for our protocols. It is unrealistic to assume that any protocol will have perfect knowledge of future channel states, so the best we can hope for in practice is to use statistics that give an accurate representation of the previous channel state.

Throughput results are given for a six-state Markov channel model in Figs. 8 and 9 for $\Delta = 1$ dB and $\Delta = 2$ dB, respectively. The performance of our protocol with either the error count or the iteration count is nearly as good as the performance of the ideal protocols that are provided with perfect channel-state information; in fact, for $\Delta = 1$ dB, the curves are almost indistinguishable. For a channels with larger variations in propagation loss (e.g., $\Delta = 2$ dB), there are larger differences between the PNSI protocol and the PPSI protocol, but the performance difference between our protocol and the PPSI protocol is very small.

The Markov model for Fig. 10 has smaller variations in propagation loss, which has a greater potential to produce adaptation statistics that are near the endpoints of the decision intervals, so there might be a risk of confusing the protocol. However, no such confusion occurs and the adaptive protocol gives performance that is near the theoretical limits.

Figs. 4 and 8–10 also indicate how much improvement could be obtained with more complex encoding and decoding equipment. For the specified rates and modulation formats, the curve for capacity-achieving codes and PNSI is an upper bound on the performance of any adaptive transmission protocol with the code rates and modulation formats in Table II. In each figure, the margin between this curve and the curve for the turbo-product codes with PNSI is dependent only on capabilities of the error-control code. The difference between the curve for the turbo product codes with PNSI and the corresponding curve for PPSI is the result of not knowing the channel state for the next packet. The small differences between the curve for PPSI and the curves for the adaptive transmission protocol is a consequence of imperfect channel-state information from the previous packet.

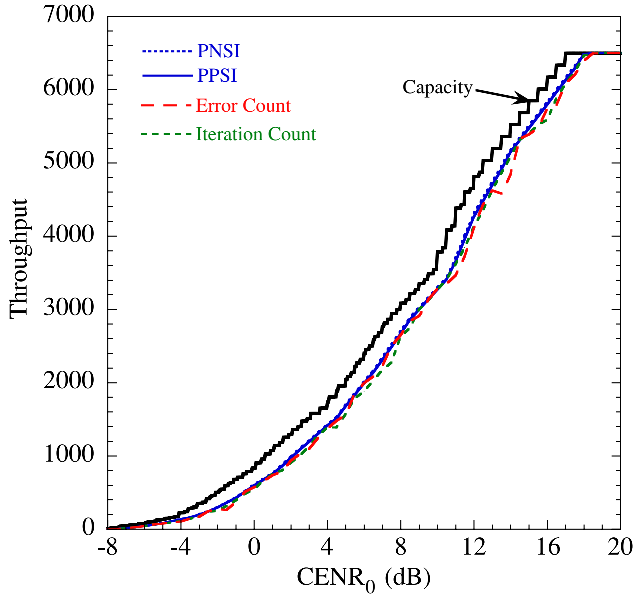

Fig. 10. Throughput for the adaptive transmission protocol with time-varying propagation loss modeled by a fifteen-state Markov chain ($\Delta = 0.5$ dB).

For all the results presented thus far, the Markov model for the channel has a single transition time in each interval between consecutive packets. In some situations, there might be a few packet intervals between some consecutive pairs of packets in a session, in which case the channel might change a few times between the determination of the adaptation statistic from one packet and its application to the choice of code-modulation combination for the next packet. Because of this possibility, we investigated the performance of the adaptive transmission protocol for a six-state Markov model with N_T transition times between consecutive packets. The results in Table IV are for the protocol that uses the error count. From Table IV, we see that if there are three or fewer transition times between consecutive packets, then the throughput is within approximately 6.1% of the throughput for a single transition time. For five or fewer transition times, the throughput is within approximately 11.1% of the throughput for a single transition time. For some values of CENR$_0$, these percentages are much lower; for example, for 4 dB they are 1.6% and 4.7%, respectively, and for 12 dB they are 0.8% and 2.3%, respectively. We also observed that the percentages are approximately the same for the PPSI protocol.

Although the design intervals defined by the endpoints in Table III give excellent performance over a wide range of channel parameters, we investigated the possibility of having the cognitive radio adjust the intervals in response to the information that it learns from past transmissions. In particular, we verified that a simple learning algorithm is effective in correcting poor choices for the endpoints of the design intervals. For this investigation, we intentionally selected endpoints that would at times cause the protocol to use a code-modulation combination whose rate is too high, which increases the packet error probability and reduces the throughput. We describe the algorithm for an adaptive transmission protocol that uses the error count together with adjusted decision intervals to choose among code-modulation combinations D_k, $1 \leq k \leq n$.

TABLE IV
THROUGHPUT FOR A SIX-STATE MARKOV CHAIN WITH N_T TRANSITION TIMES BETWEEN CONSECUTIVE PACKETS ($\Delta = 1$ dB).

CENR$_0$ (dB)	$N_T = 1$	$N_T = 2$	$N_T = 3$	$N_T = 4$	$N_T = 5$
0	715.48	693.65	672.07	650.47	636.29
4	1576.00	1559.51	1551.02	1517.80	1501.61
8	2927.57	2839.65	2782.74	2730.94	2685.61
12	4448.85	4441.45	4414.57	4381.15	4346.03

If combination D_k is used for packet i and the resulting error count is z_i, then $\gamma_{1,k,i}$ and $\gamma_{2,k,i}$ denote the endpoints of the decision intervals that are used with z_i to decide which code-modulation combination to select for packet $i+1$.

We begin the description of the learning algorithm with the reception of packet $i-1$. Suppose combination D_m is used for packet $i-1$, which is decoded correctly and produces error count z_{i-1}. As a result of using z_{i-1} in an interval test with endpoints $\gamma_{1,m,i-1}$ and $\gamma_{2,m,i-1}$, the protocol decides to use code-modulation combination D_ℓ for packet i. Of course, according to our protocol, ℓ must be $m-1$, m, or $m+1$. Because packet $i-1$ decoded correctly, all endpoints for packet i are the same as the endpoints for packet $i-1$; that is, $\gamma_{j,k,i} = \gamma_{j,k,i-1}$ for $1 \leq j \leq 2$ and $1 \leq k \leq n$. If packet i is not decoded correctly at the destination, then an error count cannot be obtained for packet i. In this event, the learning algorithm updates the endpoints to be used for packet $i+1$ according to $\gamma_{1,\ell-1,i+1} = \alpha\gamma_{1,\ell-1,i} + (1-\alpha)z_{i-1}$ and $\gamma_{2,\ell,i+1} = \alpha\gamma_{2,\ell,i} + (1-\alpha)z_{i-1}$ for $0 \leq \alpha \leq 1$. All other endpoints are unchanged. The adjustments in the endpoints make it less likely that combination D_ℓ will be used for future packets that experience the same channel conditions. The endpoints $\gamma_{1,\ell-1,i+1}$ and $\gamma_{2,\ell,i+1}$ can be rounded to integer values, if desired.

We compared the results of 100 sessions with 256 KB per session for each of two adaptive transmission protocols, one that employs the learning algorithm with $\alpha = 0.75$ and one that does not use learning to adjust the adaptation intervals. Both protocols begin each session with $\gamma_{1,3,1} = 265$ and $\gamma_{2,4,1} = 263$. According to Table III, each of these is larger by 100 than the best value. We simulated the two protocols for a channel with a time-varying excess propagation loss that is modeled by a ten-state Markov chain with $p = 0.1$, $\Delta = 0.2$ dB, and CENR$_0$ $= -3.3$ dB. We found that the protocol that uses the learning algorithm provides a larger throughput (by 5.5%), but more importantly it provides a much lower packet error probability (8×10^{-3} compared with 1.57×10^{-1}). These preliminary results suggest that a cognitive radio can learn while packets are being received, and it can use what it learns to correct mistakes that might have been made in the initial choices for the parameters of the adaptive transmission protocol.

XI. CONCLUSION

Cognitive radios equipped with adaptive transmission protocols can make intelligent use of a time-varying communication channel. For responding to slow variations in the channel, such as changes in shadow loss, we have shown that low-complexity protocols give nearly optimal performance. Additional complexity in the channel measurement and estimation systems would not improve the performance noticeably. Shannon capacity limits provide bounds on the throughput that

TABLE V
ENDPOINTS FOR THE INITIAL POWER ADJUSTMENT PROTOCOL.

μ_i	QPSK	16-QAM	4096-IQB
μ_1	0.051041	0.002199	0.504
μ_2	0.041794	0.001852	0.533
μ_3	0.034308	0.001550	0.560
μ_4	0.027949	0.001298	0.587
μ_5	0.022743	0.001070	0.611
μ_6	0.018231	0.000884	0.634
μ_7	0.014663	0.000714	0.655
μ_8	0.011675	0.000573	0.676
μ_9	0.009398	0.000462	0.694
μ_{10}	0.007382	0.000375	0.712
μ_{11}	0.005956	0.000293	0.729
μ_{12}	0.004734	0.000230	0.744
μ_{13}	0.003742	0.000187	0.759
μ_{14}	0.002955	0.000149	0.773
μ_{15}	0.002365	0.000118	0.786
μ_{16}	0.001889	0.0000937	0.798
μ_{17}	0.001500	0.0000748	0.809
μ_{18}	0.001184	0.0000597	0.820
μ_{19}	0.000947	0.0000473	0.830
μ_{20}	0.000751	0.0000368	0.840

can be achieved for any cognitive radio, no matter how much complexity is included for estimation of the channel gain.

We have described two low-complexity protocols for cognitive radios that are especially useful in dynamic spectrum access networks. The power-adjustment protocol enables successful communication in a new frequency band with an unknown propagation loss while avoiding the transmission of excessive power that would cause interference in unintended receivers and disrupt other sessions in the network. The adaptive transmission protocol chooses the combination of modulation and error-control code that is best suited to the channel state without having to estimate the state. Our protocol gives performance that is as good as hypothetical ideal protocols, yet it requires only very simple adaptation statistics that are derived easily in the receiver. The cognitive radios can learn from past receptions of packets and correct errors that might have been made in the choices for the adaptation parameters.

APPENDIX

During the power-adjustment phase, each acknowledgment packet includes a demodulator statistic. The range of the demodulator statistic is partitioned into intervals $[\mu_i, \mu_{i+1})$ with endpoints $\mu_1, \mu_2, \ldots, \mu_N$. The interval test for the demodulator statistic consists of determining which interval contains the statistic. As a result of the interval test, the power for the next packet may be increased or decreased, or the power-adjustment phase may be stopped. The minimum step size for transmitter power adjustments is denoted by β_0 and expressed in dB.

At the start of the power adjustment, it is possible that the destination will not even be able to detect the presence of the packet, in which case an acknowledgment will not be sent. Until the first time that the power is decreased, it is increased automatically by β_1 dB each time a packet is sent but not acknowledged. Each time that an acknowledgment is received for a packet transmission, then an interval test is applied to the receiver statistic. At the start of a session, a counter c_s is set to zero. This counter and a corresponding threshold S^* track the protocol and define the stopping condition. For the following description of the protocol, assume that the statistic increases as the power increases; if it decreases as the power increases, then the inequalities involving γ should be reversed.

Let γ denote the value of the demodulator statistic from the most recent packet reception. For convenience, we define $\mu_{N+1} = \infty$. If the power has not yet been decreased and if $\gamma < \mu_1$, then increase the power by β_1 dB. If the power has been decreased previously and either $\gamma < \mu_1$ or an acknowledgment is not received, then increase the power by β_2 dB. If the receiver statistic is γ and i is such that $\mu_i \leq \gamma < \mu_{i+1}$ then decrease the power by $i\beta_0$ dB. If none of the previous conditions is met, then increment c_s. If $c_s = S^*$, then the protocol is stopped; otherwise, another packet is sent and the process repeats. For our numerical results, $\beta_0 = 0.5$ dB, $\beta_1 = 5$ dB, $\beta_2 = 2$ dB, and $N = 20$. The endpoints for the intervals are listed in Table V. For QPSK and QAM, the adaptation statistics decrease as the power increases, so the inequalities in the interval test should be reversed and we set $\mu_{N+1} = 0$.

TABLE VI
INITIAL POWER ADJUSTMENT FOR $S^* = 1$.

Modulation	2δ	$N_{\max}$	Target	Δ_1	Δ_2	FR
4096-IQB	20 dB	7	2.5 dB	0.2 dB	1.2 dB	1.0 dB
4096-IQB	30 dB	8	2.5 dB	0.3 dB	1.1 dB	0.8 dB
QPSK	20 dB	7	3.5 dB	0.1 dB	1.2 dB	1.1 dB
QPSK	30 dB	8	3.5 dB	0.1 dB	1.2 dB	1.1 dB
16-QAM	20 dB	7	6.5 dB	0.3 dB	1.7 dB	1.4 dB
16-QAM	30 dB	8	6.5 dB	0.3 dB	1.8 dB	1.5 dB

For each modulation format, we simulated the power-adjustment protocol for 10,000 sessions. The sessions are started with random, independent propagation losses, so that the protocol has no information regarding the propagation loss for the first packet. The target power level P is 0.5 dB above the minimum power $P_{\min}$ that is required to provide a packet error probability of 10^{-2}. All power levels are expressed in dB. The margin of 0.5 dB is used to make it very unlikely that the final power level will be below $P_{\min}$. The initial power levels for the sessions are independent random variables, uniformly distributed on the interval from $P-\delta$ to $P+\delta$, where δ is also in dB. The initial range is defined as 2δ. Results are reported in Table VI for two values of the initial range: $2\delta = 20$ dB and $2\delta = 30$ dB. The goal is for the protocol to reduce the 20 dB and 30 dB initial uncertainty range to a final uncertainty range less than 1–2 dB within a few packets.

When the protocol stops in the ith session, the power level $\hat{P}_i$ and the packet number N_i are recorded. Results are presented in Table VI for the two initial uncertainty ranges. In these tables, $\Delta_1 = \min_i\{\hat{P}_i - P\}$, $\Delta_2 = \max_i\{\hat{P}_i - P\}$, and $N_{\max} = \max_i\{N_i\}$. The final range (FR) is $\Delta_2 - \Delta_1$. Note that each of the results represents the "worst session" out of 10,000 sessions, yet only seven or eight packets were required for convergence. Most sessions required a smaller number of packets before the stopping condition was met and most also had a final power level closer to the target. For QPSK or 16-QAM, a smaller final range can obtained by increasing the value of S^* in the stopping condition, but $S^* = 1$ gives the smallest final range for 4096-IQB.

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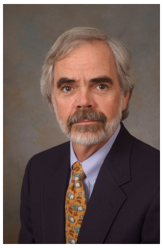

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