Chapter 1

COOPERATION, COMPETITION AND COGNITION IN WIRELESS NETWORKS

From Theory to Implementation

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Abstract Nodes and/or clusters of a wireless network operating on the same frequency can operate using three different paradigms: 1) *Competition*: Traditionally, this is information theoretically casted in the framework of interference channels. 2) *Cooperation*: Silent transmitters/receivers can help active transmitters/receivers in the transmission/reception of their messages, but have to extract this message from the underlying transmission or by other methods, and 3) *Cognitive Radio Transmission*: Some devices extract the message(s) of other transmitter(s) from their signals or by other methods, and use it to minimize interference from/to their own transmitted signals.

Competition has been well-studied in the literature. Cooperation has been less studied and *cognitive radio transmission* has not been studied much. For the cooperative case, we demonstrate that most of the multiple-input multiple-output (MIMO) space-time diversity gain can also be achieved through cooperative communications with single-antenna/multiple-antenna nodes when there is one receiving agent. In particular, for the single antenna case, we consider communication to take place between clusters of nearby nodes. We show the existence of cooperative codes for communications for which the intra-cluster negotiation penalty is in principle small and almost all the diversity gain of traditional space-time codes may be realized. For example, for a single transmitter node with two cooperators and one receiver node, if the cooperators have as little as 10 dB path loss advantage over the receiver, the penalty for cooperation over traditional space-time systems is negligible. Furthermore, we demonstrate and discuss the implementation of this idea in an orthogonal frequency division multiplexing (OFDM) system using a software defined ratio (SDR) platform. On the other hand, cooperative beamforming is an alternative way of realizing cooperative gain, particularly for a wireless sensor network. We analyze the statistical average properties and distribution of the beampattern of cooperative beamforming using the theory of random arrays.

For cognitive radio transmissions, which captures a form of asymmetric cooperation, we define a generalized cognitive radio channel as an n-transmitter, m-receiver interference channel in which sender i obtains (causally or non-causally) the messages of senders 1 through $i-1$. For simplicity, only the two sender, two receiver case is considered. In this scenario, one user, a cognitive radio, obtains (genie assisted, or causally) knowledge of the data to be transmitted by the other user. The cognitive radio may then simultaneously transmit over the same channel, as opposed to waiting for an idle channel as in a traditional cognitive radio channel protocol. Gel'fand-Pinsker coding (and the special case of dirty-paper coding) and ideas from achievable region constructions for the interference channel are used, and an achievable region for the cognitive radio channel is computed. In the Gaussian case, the described achievable region is compared to the upper bound provided by the 2×2 Gaussian MIMO broadcast channel, and an interference-free channel. We then extend the results to provide an achievable region for cognitive multiple access networks.

Keywords: Cognitive multiple access network, cognitive radio channel, cooperative beamforming, cooperative diversity, dirty-paper coding, interference channel, multiple-input multiple-output, orthogonal frequency division multiplexing, space-time codes, wireless networks.

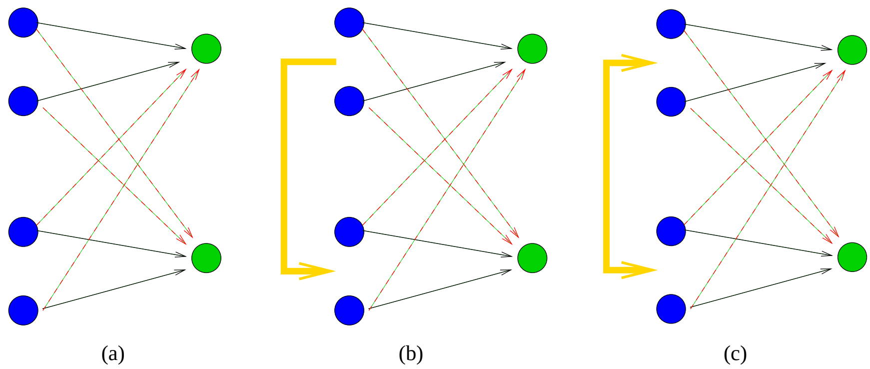


Figure 1.1. Inter-cluster behaviors of wireless networks. (a) Competitive behavior. All messages are independent. (b) Cognitive behavior. The thick solid arrow indicates that the second cluster has knowledge of the messages of the first cluster, but not vice versa. (c) Cooperative behavior. The thick solid two-way arrow indicates that each cluster knows the messages to be sent by the other cluster.

1. Introduction

All networks can be partitioned into clusters of nodes. When a wireless network is partitioned in this fashion, we can speak of *intra-cluster* as well as *inter-cluster behavior.* For *intra-cluster* behavior, we look at the nodes within each cluster, and notice that nodes can compete for resources (*competitive behavior*), can fully cooperate (*cooperative behavior*), or can partially cooperate in what we will call *cognitive behavior*. Similarly, for *inter-cluster* behavior, entire clusters can behave in a *competitive*, *cooperative*, or *cognitive* fashion. Figs. 1.1(a), 1.1(b) and 1.1(c) demonstrate examples of inter-cluster competitive, cognitive and cooperative behaviors respectively. The thick solid lines indicate that the messages of one cluster are known to the other. In Fig. 1.1(a) the two competitive clusters are independent and compete for the channel. In Fig. 1.1(b) the lower cluster has knowledge of the two messages to be sent by the upper cluster, but not vice versa. This is an example of inter-cluster cognitive behavior. In Fig. 1.1(c) the two clusters know each others' messages, an example of inter-cluster cooperative behavior. Within one cluster the messages may remain independent when the cluster is operated in an *intra-cluster competitive fashion*. If messages within one cluster were to be shared, it would constitute *intra-cluster cooperation (full) or cognition (partial)*.

In a full cooperation paradigm, nodes which would have otherwise remained silent in a traditional competition paradigm, cooperate with the source and destination to increase communication capacity and reliability. Arguably, the initial work on cooperative communications stretches as far back as the pioneering papers by van der Meulen [Van der Meulen, 1971] and Cover *et al.* [Cover and El Gamal, 1979] on the relay channel. However, the results obtained there do not appear to be directly applicable to inexpensive relays for wireless networks. This is because in realistic wireless models, it is not practically feasible to transmit and receive on the same antenna simultaneously (half-duplex constraint), since the intensity of the near field of the transmitted signal is much higher than that of the far field of the received signal. In wireless systems, the channel fading coefficients are usually not known to the transmit nodes; only the receive nodes have knowledge of the channel, i.e., realistic wireless channels are *compound channels* [Wolfowitz, 1978, Csiszár and Körner, 1981]. Finally, while the degraded relay channel has been completely solved [Cover and El Gamal, 1979, Reznik et al., 2002], in wireless systems most noise is due to thermal noise in the receiver frontend. While it may be reasonable to assume that the relay has a better signal-to-noise ratio (SNR) than the ultimate receiver, it is unrealistic to assume that the receiver is a degraded version of the relay. Various extensions of the non-compound relay channel may be found in [Schein and Gallager, 2000, Gupta and Kumar, 2003, Gastpar et al., 2002, Gastpar and Vetterli, 2002a, Gastpar and Vetterli, 2002b, Khojastepour et al., 2003a, Khojastepour et al., 2003b, Khojastepour et al., 2003c, Kramer et al., 2004, Wang et al., 2005, Xie and Kumar, 2004].

Some more recent work on cooperative communications with emphasis on treating the wireless channel as a compound channel may be found in [Laneman and Wornell, 2003, Laneman et al., 2004, Hunter and Nostratinia, 2002, Hunter et al., 2003, Sendonaris et al., 2003a, Sendonaris et al., 2003b]. In [Laneman and Wornell, 2003], the authors consider a two-stage protocol (where the source transmits for a fixed amount of time followed by a fixed duration cooperation phase) to solve the half-duplex constraint and consider repetition and space-time coding based cooperative diversity algorithms. This is extended in [Laneman et al., 2004] with the consideration of adaptive protocols such as selection relaying and incremental relaying. In [Hunter and Nostratinia, 2002, Hunter et al., 2003] a similar time-division (TD) approach is employed where the relay is permitted to transmit its own information during the second phase if it is unable to cooperate. In [Sendonaris et al., 2003a, Sendonaris et al., 2003b], the authors assume two dedicated orthogonal subchannels between two mobile nodes, derive an achievable region for communication to a base station and consider code division multiple access (CDMA) implementation aspects. These results are derived by employing coding techniques [Willems, 1982, Willems et al., 1983] similar to those used for multiple access channels with generalized feedback [Carleial, 1982].

In Section 2, we present a bandwidth efficient decode and forward approach [Mitran et al., 2005] that does not fix phase durations or orthogonal subchannels to resolve the half-duplex constraint: each relay determines based on its own receive channel when to listen and when to transmit. Furthermore, the transmitters are not aware of the channel and we make no assumption of degradedness. In the case of multiple relays assisting the source, the approach permits one relay to assist another in receiving the message. However, more recent work along this line may be found in [Katz and Shamai, 2004, Azarian et al., 2004]. Finally, we briefly outline a practical cooperative system founded upon orthogonal frequency division multiplexing (OFDM) transmission [Shin et al., 2005].

Cooperative beamforming is an alternative cooperation technique to cooperative diversity especially suited for ad hoc sensor networks [Ochiai et al., 2005]. The advantages and applications of traditional beamforming with antenna arrays are well known; in wireless communications, beamforming is a powerful means for interference suppression which enables space division multiple access. Even though each node is equipped with a single antenna, if the nodes in a cluster share the information *a priori* and synchronously transmit data, it may be possible to beamform when transmitting (and also receiving) the data in a distributed manner. The overhead due to intra-cluster information sharing may be relatively small as this can be done by low-cost short-distance broadcasting-type communication among nodes. Thus, with distributed cooperative beamforming, the nodes can send the collected information to the far-end receiver over long distances with high efficiency. Also, only the cluster in the specified target direction receives the data with high signal power and no significant interference occurs for clusters in other directions. In Section 3, the beampattern aspects of cooperative beamforming using random arrays are analyzed in the framework of wireless networks. The average statistical figure of merits are first developed. In particular, it is shown that with N cooperative nodes, one can achieve a directivity of order N asymptotically. The probability distribution of the far-field beampattern is then analyzed.

While the cooperative schemes described thus far have relied on symmetric cooperation between nodes, asymmetric cooperation is also possible, and has been inspired by an explosion of interest in cognitive and software radios, as is evidenced by FCC proceedings [FCC, 2003,FCC, 2005], talks [Mitola, 1999b], and papers [Mitola, 1999a, Mitola, 2000]. *Software Defined Radios* (SDR) [Mitola, 1995] are devices used to communicate over the wireless medium equipped with either a general purpose processor or programmable silicon as hardware base, and enhanced by a flexible software architecture. They are low-cost, can be rapidly upgraded, and may adapt to the environment in real-time. Such devices are able to operate in many frequency bands under multiple transmission protocols and employ a variety of modulation and coding schemes. Taking this one step further, [Mitola, 2000] coined the term *cognitive radio* for software defined radios capable of sensing their environment and making decisions instantaneously, without any user intervention. This allows them to change their modulation schemes or protocols so as to better communicate with the sensed environment. Apart from their low cost and flexibility, another benefit of software radio technology is spectral efficiency. Currently, FCC measurements [FCC Spectrum Policy Task Force, 2002], indicate that at any time roughly 10% of the unlicensed frequency spectrum is actively in use (leaving 90% unused). If a wireless device such as a cognitive radio is able to sense an idle channel at a particular frequency band (or time), then it can shift to that frequency band (or time slot) to transmit its own information, dramatically increasing spectral (or temporal) efficiency.

Although cognitive radios have spurred great interest and excitement in industry, many of the fundamental theoretical questions on the limits of such technology remain unanswered. In current cognitive radio protocol proposals, the device listens to the wireless channel and determines, either in time or frequency, which part of the spectrum is unused [Horne, 2003]. It then adapts its signal to fill this void in the spectrum space. Thus, a device transmits over a certain time or frequency only when no other node does. In Section 4, the cognitive radio behavior is generalized to allow two users to simultaneously transmit over the same time or frequency [Devroye et al., 2004]. According to this approach, a cognitive radio will listen to the channel and, if sensed idle, proceed with the traditional cognitive radio channel model, that is, transmit during the voids. On the other hand, if another sender is sensed, the radio may decide to proceed with simultaneous transmission. The cognitive radio need not wait for an idle channel to start transmission. Specifically, we will prove achievability, in the information theoretic sense, of a certain set of rates at which two senders (cognitive radios, denoted as $\mathcal{S}_1$ and $\mathcal{S}_2$) can transmit simultaneously over a common channel to two independent receivers $\mathcal{R}_1$ and $\mathcal{R}_2$ when $\mathcal{S}_2$ is aware of the message to be sent by $\mathcal{S}_1$. The methods borrow ideas from Gel'fand and Pinsker's coding for channels with known interference at the transmitter [Gel'fand and Pinsker, 1980], Costa's dirty paper coding [Costa, 1983], the interference channel [Carleial, 1978], the Gaussian MIMO broadcast channel [Weingarten et al., 2004], and the achievable region of the interference channel described by Han and Kobayashi [Han and Kobayashi, 1981]. Finally, we discuss extensions of the results to cognitive multiple access networks [Devroye et al., 2005].

2. Cooperative Diversity

Preliminaries

For simplicity, we consider three nodes denoted as source (s), relay (r) and destination (d) as illustrated in Fig. 1.2, each equipped with N_s, N_r and N_d antennas respectively. The results can readily generalize to multiple relay nodes. We assume that while listening to the channel, the relay does not transmit. Hence, the communications protocol is as follows. The source node wishes

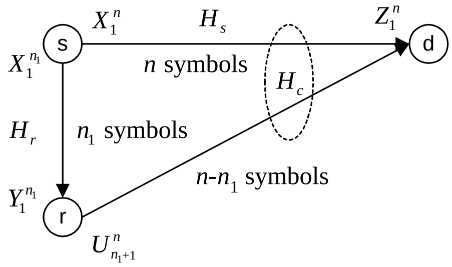


Figure 1.2. The cooperative communications problem for two transmit cooperators and one receiver.

to transmit one of 2^{nR} messages to the destination employing n channel uses. While not transmitting, the relay node listens. Due to the relay node's proximity to the source, after n_1 samples from the channel (a number which the relay determines on its own and for which the source has no knowledge), it may correctly decode the message. After decoding the message, it then proceeds to transmit for the remaining $n-n_1$ transmissions in an effort to improve the reception of the message at the destination. The destination is assumed to be made aware of n_1 before attempting to decode the message. This may be achieved by an explicit low-rate transmission from the relay to the destination. Alternatively, if the value of n_1 is constrained to some integer multiple of a fundamental period n_0 (say $n_0 \sim \sqrt{n}$), then the destination may estimate n_1 accurately using power detection methods. We denote the first phase of the n_1 transmissions as the *listening* phase while the last $n - n_1$ transmissions as the *collaboration* phase.

We assume that all channels are modeled as additive white Gaussian noise (AWGN) with quasi-static fading. In particular, X and U are column vectors representing the transmission from the source and relay nodes respectively and we denote by Y and Z the received messages at the relay and destination respectively. Then during the listening phase we have

$$Z = H_s X + N_Z \tag{1.1}$$
$$Y = H_r X + N_Y, \tag{1.2}$$

where N_Z and N_Y are column vectors of statistically independent complex AWGN with variance 1/2 per row per dimension, H_s is the the fading matrix between the source and destination nodes and likewise, H_r is the fading matrix between the source and relay nodes. During the collaboration phase, we have

$$Z = H_c[X^T, U^T]^T + N_Z, \tag{1.3}$$

where H_c is a channel matrix that contains H_s as a submatrix (see Fig. 1.2).

We further assume that the source has no knowledge of the H_r and H_c matrices (and hence the H_s matrix too). Similarly, the relay has no knowledge of H_c but is assumed to know H_r. Finally, the destination knows H_c. Without loss of generality, we will assume that all transmit antennas have unit average power during their respective transmission phases. Likewise, the receive antennas have unit power Gaussian noise. If this is not the case, the respective H matrices may be appropriately scaled row-wise and column-wise.

Under the above unit transmit power per transmit antenna and unit noise power per receive antenna constraint, it is well known that a MIMO system with Gaussian codebook and with rate R bits/channel use can reliably communicate over any channel with transfer matrix H such that $R < \log_2 \det(I + HH^\dagger) \stackrel{\triangle}{=} C(H)$ [Telatar, 1999, Foschini and Gans, 1998], where I denotes the identity matrix and $H^\dagger$ is the conjugate transpose of H.

Intuition for the above problem then suggests the following. During the listening phase, the relay knowing H_r listens for an amount of time n_1 such that $nR < n_1 C(H_r)$. During this time, the relay receives at least nR bits of information and may reliably decode the message. The destination, on the other hand, receives information at the rate of $C(H_s)$ bits/channel use during the listening phase and at the rate of $C(H_c)$ bits/channel use during the cooperative phase. It may reliably decode the message provided that $nR < n_1 C(H_s) + (n - n_1) C(H_c)$. In the limit as $n \to \infty$, the ratio n_1/n approaches a fraction f and we may conjecture that there exists a "good" code of rate R for the set of channels (H_r, H_c) which satisfies

$$R \leq fC(H_s) + (1 - f)C(H_c) \text{ and } R \leq fC(H_r), \tag{1.4}$$

for some $f \in [0, 1]$. We note that if the channel between the source and the relay is particularly poor, we may fall back on the traditional point-to-point communications paradigm and add the following region to that given in (1.4)

$$R \leq C(H_s). \tag{1.5}$$

Achievable Rate of Cooperative Diversity

In this section, we define and state a theorem on the achievability for compound synchronous relay channels. The codebook for the source will be denoted by $C_s^{(n)}$ and consists of $K2^{nR}$ codewords for some constant $K > 0$. The wth codeword of the source node codebook will be denoted by $x_1^n(w)$. If the source node has N_s transmit antennas, then each codeletter consists of a column vector with dimension N_s and each codeword is in fact an $N_s \times n$ matrix. For the relay, we will denote by $C_r^{(n)}$ a family of n codebooks $C_r^{(n,n_1)}$ indexed by $1 \leq n_1 \leq n$ where $C_r^{(n,n_1)}$ is a codebook with $K2^{nR}$ codewords of length $n - n_1$. The wth codeword of $C_r^{(n,n_1)}$ will be denoted by $u_{n_1+1}^n(w)$. Finally, we will denote by $C^{(n)} = \{C_s^{(n)}, C_r^{(n)}\}$.

Before explaining the encoding procedure, it will help to explain the decoder. With each message $W = w$, pair of channels (H_r, H_c) and value of n_1, we associate some disjoint (over w) subsets of $\mathbb{C}^N$ as follows: $S_{w,H_c,n_1} \subset \mathbb{C}^{N_d \times n}$ and $R_{w,H_r,n_1} \subset \mathbb{C}^{N_r \times n_1}$. We shall refer to $C^{(n)}$ as the encoder or codebook and the sets S_{w,H_c,n_1} and R_{w,H_r,n_1} as the decoder.

Encoding and decoding: The source wishes to transmit message $W = w$ to the destination. To that end, the source looks up the wth codeword in its codebook and proceeds to transmit it to the destination and the relay. The relay, knowing the channel H_r, decides upon the smallest value of n_1 for which $nR + \delta < n_1 C(H_r)$ (for some fixed $\delta > 0$) and for which $\delta < n_1/n < 1 - \delta$. If no such n_1 exists, the relay takes $n_1 = n$, makes no attempt to decode the message and remains silent. If $n_1 < n$, the relay listens to the channel for this duration and lists all the $\hat{w}$ for which $Y_1^{n_1} \in R_{\hat{w},H_r,n_1}$. If $\hat{w}$ exists (and is hence unique), the relay looks up the $\hat{w}$th codeword in the $C_r^{(n,n_1)}$ codebook and proceeds to transmit it for the remaining $n - n_1$ channel uses. Otherwise, the relay declares an error.

After the last transmission, the destination has now received Z_1^n where

$$Z_i = \begin{cases} H_s X_i + N_{Z,i} & i \le n_1 \\ H_c [X_i^T, U_i^T]^T + N_{Z,i} & i > n_1, \end{cases} \tag{1.6}$$

and is informed of the value of n_1. The destination then proceeds to list all $\hat{w}$ such that $Z_1^n \in S_{\hat{w},H_c,n_1}$. If $\hat{w}$ exists (and is hence unique), the destination declares the transmitted message as $\hat{W} = \hat{w}$. Otherwise an error is declared. We shall abuse notation and denote by the event $\hat{W} \neq W$ the case when either the relay or the destination declares an error or decodes an incorrect message (if the relay makes no attempt at decoding the message, it cannot produce an error).

Since the source, relay and destination nodes each have N_s, N_r and N_d antennas respectively, we note that $H_r \in \mathbb{C}^{N_r \times N_s}$ and $H_c \in \mathbb{C}^{N_d \times (N_s + N_r)}$. We denote by $\mathcal{H}$ a subset of compound relay channels, i.e., $\mathcal{H} \subset \mathbb{C}^{N_r \times N_s} \times \mathbb{C}^{N_d \times (N_s + N_r)}$. Also, for a codebook $C^{(n)}$, we denote by λ_n^s (where the superscript s denotes synchronism)

$$\lambda_n^s = \max_w \sup_{(H_r,H_c) \in \mathcal{H}} P[\hat{W} \neq W | W = w, H_r, H_c]. \tag{1.7}$$

DEFINITION 1.1 *(Achievability for a compound relay channel) A rate R is said to be achievable for a set of pairs $(H_r, H_c) \in \mathcal{H}$ if for any $\epsilon > 0$, there exists a sequence of encoders and decoders $C^{(n)}$, S_{w,H_c,n_1} and R_{w,H_r,n_1} in n*

such that $\lambda_n^s \to 0$ *as* $n \to \infty$ *and each codeword in each sub-codebook of* $C^{(n)}$ *has average power at most* $1 + \epsilon$. □

Before stating a theorem on the existence of good codes, we define a norm on a complex matrix H with entries $H_{i,j}$ as $||H|| \stackrel{\triangle}{=} \max_{i,j}\{|H_{i,j}|\}$.

THEOREM 1.2 *Consider the set* $\mathcal{H}_{\delta,L}(R)$ *of matrices* (H_r, H_c) *such that* $||H_r|| \leq L$ *and* $||H_c|| \leq L$ *and which satisfy either both*

$$R + \delta \leq fC(H_s) + (1-f)C(H_c) \text{ and } R + \delta \leq fC(H_r) \tag{1.8}$$

for some $\delta \leq f \leq 1 - \delta$ *(each* f *may depend on* H_r*), or*

$$R + \delta \leq C(H_s). \tag{1.9}$$

Then, the rate R *is achievable for the compound relay channel* $\mathcal{H}_{\delta,L}(R)$ *for any* $\delta > 0$ *and* $L > 0$. □

This theorem essentially states that the region in equations (1.4) – (1.5) may be arbitrarily approximated by taking $\delta > 0$ sufficiently small and $L > 0$ sufficiently large. The proof is given in [Mitran et al., 2005], where the result was also extended to the case of bounded asynchronism, similar to that in [Cover et al., 1981], between nodes.

Performance Analysis

We analyze the theoretical performance of a code that achieves the compound channels in Theorem 1 for single antenna nodes when the fading is quasi-static Rayleigh distributed. In particular, since L was an arbitrarily large constant, we shall take $L = \infty$. Similarly, since δ was an arbitrarily small positive number, we take $\delta = 0$. Furthermore, we will relax our restrictions on unit power per transmit antenna (as stated earlier, this was allowable since the respective H matrices could be appropriately scaled to compensate). In this section, it will be more convenient to keep the H matrices fixed and show the explicit dependence of the outage probability on the receive signal power at the destination node (during the listening phase) per transmit antenna at the source node, E_s, and the noise power at each receive antenna, σ^2. Under these conditions, we have

$$C(H, \gamma) \stackrel{\triangle}{=} \log_2 \det(I + \gamma H H^\dagger), \tag{1.10}$$

where $\gamma \stackrel{\triangle}{=} \frac{E_s}{\sigma^2}$ and the expression holds regardless of the number of transmit antennas (as E_s is defined as the normalized receive power per transmit antenna). We model the proximity of the relay node to the source node by a reduction $G \in \mathbb{R}$ in path loss, or equivalently, an increase in the achievable rate between cooperator nodes as expressed by $C(H, G\gamma)$. With these conventions, we will assume that the code successfully transmits the message from the source to the destination in a two cooperator scenario provided that either

$$R \leq fC(H_r, G\gamma) \text{ and } R \leq fC(H_s, \gamma) + (1-f)C(H_c, \gamma), \tag{1.11}$$

for some $0 < f < 1$, or

$$R \leq C(H_s, \gamma) \tag{1.12}$$

holds.

We note that the fraction f is determined by the relay and depends only on the realization of H_r according to

$$f^* \triangleq \min\{1, R/C(H_r, G\gamma)\}. \tag{1.13}$$

Since $C(H_c, \gamma) \geq C(H_s, \gamma)$, this is the optimal choice of f to minimize the outage probability of our scheme. Even if the relay knew H_c, it could not do better. Furthermore, given f, the effective receive power at the destination is $(2-f)E_s$ as the relay was only transmitting for a fraction $1-f$ of the total transmission time. The effective receive SNR for the duration of the transmission is then $(2-f)E_s/\sigma^2$. We may thus rewrite the outage probability P_{out} *for our proposed scheme* as

$$P_{out} = P\left[R > f^* C(H_s, \gamma) + (1-f^*)C(H_c, \gamma)\right]. \tag{1.14}$$

Under the assumption that the relay transmits the same instantaneous energy as the source node and suffers the same amount of path loss, (1.14) can be rewritten using the definition in (1.10) as

$$\begin{aligned} P_{\text{out}} = P[R >& f^* \log_2\left(1+\gamma|H_s|^2\right) \\ &+ (1-f^*)\log_2\left(1+\gamma\left(|H_s|^2+|H_{r,d}|^2\right)\right)]. \end{aligned} \tag{1.15}$$

Evaluation of (1.15) is difficult to carry out even numerically, since exact analysis typically yields double integrals. Using the Jensen's inequality, a tight lower bound for (1.15) can be derived as [Mitran et al., 2005]

$$\begin{aligned} P_{\text{out}} &\geq P\left[R > \log_2\left(1+\gamma\left(|H_s|^2 + (1-f^*)|H_{r,d}|^2\right)\right)\right] \\ &= P\left[|H_s|^2 + (1-f^*)|H_{r,d}|^2 < \frac{2^R-1}{\gamma}\right]. \end{aligned} \tag{1.16}$$

Note that the right hand side of (1.16) is the cumulative distribution function (CDF) of the random variable $|H_s|^2 + (1-f^*)|H_{r,d}|^2$, which can be found

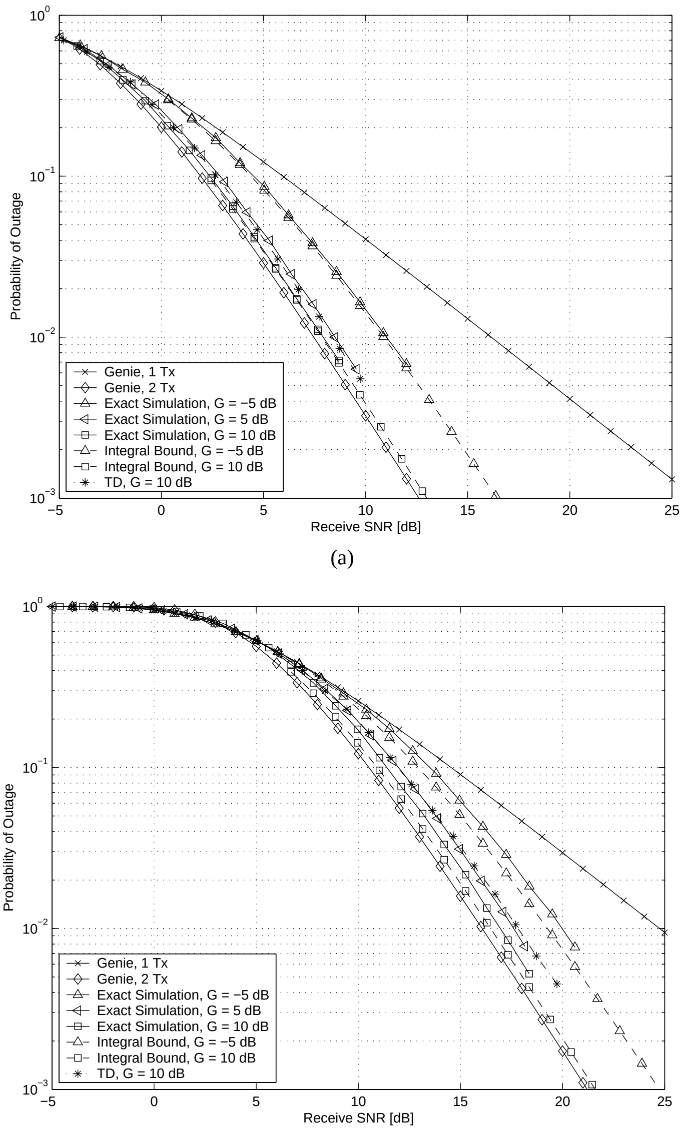


Figure 1.3. Outage probability of two transmit collaborators and one receiver for various geometric gain factors G. (a) $R = 0.5$. (b) $R = 2$.

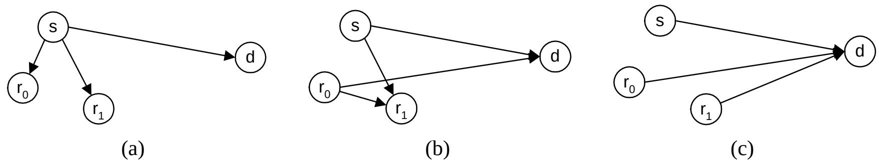


Figure 1.4. The three cooperators problem with one receiver. (a) Nodes r_0, r_1, and d are listening. (b) Node r_0 has stopped listening and started cooperating. It transmits to nodes d *and* r_1. (c) Node r_1 has stopped listening and started cooperating.

by standard algebra of random variables. The explicit expression of (1.16) is given in [Mitran et al., 2005] for the case that all fading channel coefficients $H_r, H_s, H_{r,d}$ are i.i.d. complex Gaussian random variables.

Fig. 1.3 illustrates the simulated outage probability (for $R = 0.5$ and $R = 2$) and the corresponding lower bound (1.16) for various values of G versus the averaged receive SNR for quasi-static Rayleigh fading channels, i.e., channels where each of the H matrices have independent circularly symmetric Gaussian distributed entries with total variance 1. Exact results were obtained by Monte-Carlo simulation of equations (1.13) and (1.14). These simulation results are confirmed by the tightness of the lower bound (1.16). Also illustrated in Fig. 1.3 is the outage probability of a traditional 1×1 and 2×1 space-time system, which is referred to as a genie bound. We see that even with as little gain as $G = -5$ dB, for an outage probability of 10^{-2}, the loss in performance is only 3.5 dB compared to the genie 2×1 bound. With $G = 10$ dB, the genie 2×1 bound is closely approached by the proposed cooperative scheme.

Finally, it is instructive to compare the performance of this scheme (where the relay listens for the smallest fraction of time f that is necessary) to a scheme where f is not allowed such flexibility. In Fig. 1.3, the performance of a scheme where f is constrained to 0.5 or 1.0 is also illustrated. (We use the notation TD for this scheme. Whereas $f = 0.5$ corresponds to a half listening/half cooperation protocol, $f = 1.0$ is equivalent to no cooperation.) For $G = 10$ dB and $R = 0.5$, we see that this TD scheme performs as well as the proposed scheme with $G = 5$ dB at an outage probability of 10^{-2}. Hence, in this case, the penalty for employing a predetermined TD scheme is equivalent to a 5 dB penalty in geometric gain. For higher rates such as $R = 2$, the penalty increases.

The result in Theorem 1 generalizes in a straightforward manner to multiple transmit collaborators. In particular, an extension to the case of three transmit cooperators and one receiver, as illustrated in Fig. 1.4, is presented in [Mitran et al., 2005]. We note a remarkable feature of the scheme in Fig. 1.4. If the relay r_0 has a better channel from the source than the relay r_1, relay r_1 may receive information not only from the source node s, but from the relay node r_0

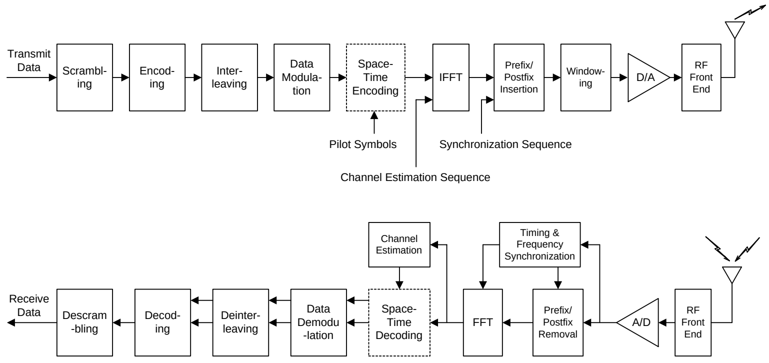


Figure 1.5. Block diagram of the CO-OFDM transmitter and receiver (dotted blocks are used only in the cooperation phase).

as soon as r_0 has finished listening. By symmetry, a similar situation is possible if relay r_1 has a better channel than r_0. If the number of cooperative nodes were further increased to m, we would see a cascade effect by which the relays would quickly share among themselves the message by way of $m(m-1)/2$ possible paths. More recent literature that allows for this sort of information sharing strategy may be found in [Katz and Shamai, 2004,Azarian et al., 2004].

Implementation: Cooperative OFDM System

A space-time cooperative system based on orthogonal frequency division multiplexing (OFDM), which is referred to as a cooperative (CO)-OFDM system, has been designed in [Shin et al., 2005]. The system will be implemented on a software radio platform. We briefly outline the main features of the CO-OFDM system and some performance results. Details can be found in [Shin et al., 2005].

Fig. 1.5 illustrates a block diagram of the CO-OFDM transmitter and receiver. The structure is similar to that of the IEEE 802.11a standard [IEEE 802.11a-1999, 1999] except for the use of space-time cooperation. Note that transmit symbols are encoded according to a form of time-division cooperative diversity protocol discussed in Section 2.0. The transmission of each frame involves two subsequent phases with fixed duration: the *listening phase* and the *cooperation phase*. In the *listening phase*, the source broadcasts a listening subframe to the relays and destination. Space-time coding is not employed in this phase, since the source is equipped with only one transmit antenna. If the destination succeeds in decoding the listening subframe, the following cooperation phase is ignored at the destination. Otherwise, the destination attempts to decode the succeeding cooperation subframe. Note that the relays and des-

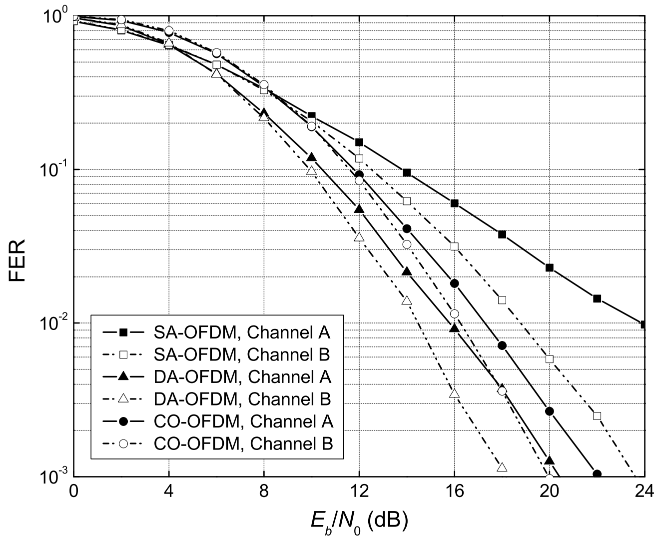


Figure 1.6. The overall FER performance of the CO-OFDM system.

tination can realize whether decoding of each subframe is successful or not by computing the checksum of the frame check sequence.

In the *cooperation phase*, the source constructs and transmits a cooperation subframe, which corresponds to a portion of the space-time coded version of the listening subframe. The behavior of the relay depends on whether it has succeeded or not in decoding the preceding listening subframe. If a relay has succeeded in decoding, the relay also constructs and transmits a cooperation subframe, which corresponds to another portion of space-time coded signal. Then the destination may receive the complete space-time coded signal from the source and relay, enabling the reliable decoding of the cooperation subframe. Otherwise, if the relay has failed to decode the listening subframe, it is silent in the cooperation phase. The listening and cooperation subframes are allowed to be transmitted at different transmission rates. For the case of a single relay node, [Shin et al., 2005] has also devised a frame structure including preamble sequences, and provided simple and effective timing and frequency synchronization algorithms and a channel estimation algorithm.

Fig. 1.6 shows the overall frame error rate (FER) performance of the CO-OFDM system, when the synchronization and channel estimation algorithms proposed in [Shin et al., 2005] are adopted. The performance of a single-antenna OFDM (SA-OFDM) system and a double-antenna OFDM (DA-OFDM) system without cooperation is also presented for comparison. The geometric gain G is assumed to be 10 dB. Other details of simulation conditions are given in [Shin et al., 2005]. From the figure, we can observe significant performance

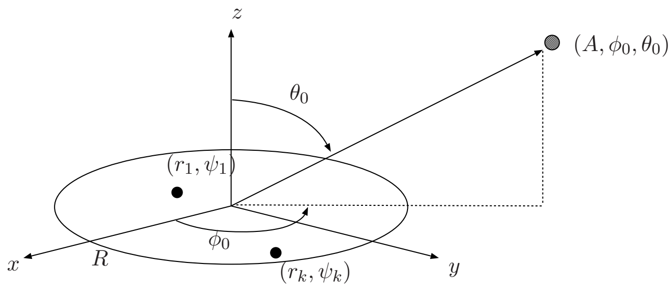


Figure 1.7. Definitions of notation for cooperative beamforming.

improvement of the CO-OFDM system over the SA-OFDM system. At a FER of 10^{-2}, for example, the energy gain of the CO-OFDM system over the SA-OFDM system is found to be as much as 6.7 dB for channel A, and 2.5 dB for channel B, where the channel models are given in [European Telecommunications Standards Institute, 1998, Shin et al., 2005]. From the slopes of the FER curves, we notice that the CO-OFDM system achieves a diversity order comparable to that of the DA-OFDM system, as predicted by the theory.

3. Cooperative Beamforming

System Model and Beampattern

The geometrical configuration of the distributed nodes and destination (or target) is illustrated in Fig. 1.7 where, without loss of generality, all the cooperative nodes are assumed to be located on the x-y plane. The kth node location is thus denoted in polar coordinates by (r_k, ψ_k). The location of the destination is given in spherical coordinates by (A, ϕ_0, θ_0). Following the standard notation in antenna theory [Balanis, 1997], the angle $\theta \in [0, \pi]$ denotes the elevation direction, whereas the angle $\phi \in [-\pi, \pi]$ represents the azimuth direction. In order to simplify the analysis, the following assumptions are made:

- The location of each node is chosen randomly, following a uniform distribution within a disk of radius R.
- Each node is equipped with a single ideal isotropic antenna.
- All nodes transmit identical energies, and the path losses of all nodes are also identical. Thus the underlying model falls within the framework of phased arrays.
- There is no reflection or scattering of the signal. Thus, there is no multipath fading or shadowing.
- The nodes are sufficiently separated that any mutual coupling effects [Balanis, 1997] among the antennas of different sensor nodes are negligible.

- All the nodes are perfectly synchronized so that no frequency offset or phase jitter occurs.

Let a realization of node locations $\boldsymbol{r} = [r_1, r_2, \ldots, r_N] \in [0, R]^N$ and $\boldsymbol{\psi} = [\psi_1, \psi_2, \ldots, \psi_N] \in [-\pi, \pi]^N$ be given, where N denotes the number of nodes. We are interested in the radiation pattern in the far-field region, and we assume that the far-field condition $A \gg r_k$ holds. Then under the above assumptions, the array factor in the far field can be approximated as [Ochiai et al., 2005]

$$F(\phi, \theta | \boldsymbol{r}, \boldsymbol{\psi}) \approx \frac{1}{N} \sum_{k=1}^{N} e^{j \frac{2\pi}{\lambda} r_k [\sin\theta_0 \cos(\phi_0 - \psi_k) - \sin\theta \cos(\phi - \psi_k)]} \tag{1.17}$$

Of particular interest in practice is the case where $\theta_0 = \frac{\pi}{2}$, i.e., the destination node is in the same plane as the cooperative transmit nodes. Without loss of generality, we also assume that $\phi_0 = 0$, since the choice of ϕ_0 does not change the results. Under this assumption and the assumption on the distribution of node location, the array factor in (1.17) is simplified to [Ochiai et al., 2005]

$$F(\phi | \boldsymbol{z}) \triangleq F(\phi, \theta = \pi/2 | \boldsymbol{r}, \boldsymbol{\psi}) = \frac{1}{N} \sum_{k=1}^{N} e^{-j4\pi \tilde{R} \sin\left(\frac{\phi}{2}\right) z_k}, \tag{1.18}$$

where each element z_k of $\boldsymbol{z}$ is given as $z_k \triangleq \frac{r_k}{R} \sin(\psi_k - \phi/2)$, and $\tilde{R} \triangleq \frac{R}{\lambda}$ is the radius of the disk normalized by the wavelength λ. Finally, the far-field beampattern can be defined as

$$P(\phi | \boldsymbol{z}) \triangleq |F(\phi | \boldsymbol{z})|^2 = \frac{1}{N} + \frac{1}{N^2} \sum_{k=1}^{N} e^{-j\alpha(\phi) z_k} \sum_{\substack{l=1 \\ l \neq k}}^{N} e^{j\alpha(\phi) z_l}, \tag{1.19}$$

where $\alpha(\phi) \triangleq 4\pi \tilde{R} \sin(\phi/2)$.

Average Far-Field Beampattern

By taking the average of (1.19) over all realizations of $\boldsymbol{z}$, we obtain the average beampattern as

$$P_{\text{av}}(\phi) \triangleq E_{\boldsymbol{z}} \{P(\phi | \boldsymbol{z})\} = \frac{1}{N} + \left(1 - \frac{1}{N}\right) \left| 2 \cdot \frac{J_1(\alpha(\phi))}{\alpha(\phi)} \right|^2, \tag{1.20}$$

where $J_n(x)$ is the nth order Bessel function of the first kind. Using this formula, several statistical properties have been investigated in [Ochiai et al., 2005], including positions of peaks and zeros, 3 dB beamwidth, 3 dB sidelobe region, and average directivity. The most important amongst them is the average directivity, which characterizes how much radiated energy is concentrated on in the desired direction relative to a single isotropic antenna. The average directivity D_{av} is defined as

$$D_{\text{av}} \triangleq E_{\boldsymbol{z}}\left\{D(\boldsymbol{z})\right\} = E_{\boldsymbol{z}}\left\{\frac{2\pi}{\int_{-\pi}^{\pi} P(\phi|\boldsymbol{z})d\phi}\right\}. \tag{1.21}$$

In [Ochiai et al., 2005], the following Theorem on a lower bound of the normalized directivity was proved.

THEOREM 1.3 (NORMALIZED DIRECTIVITY LOWER BOUND) *For large $\tilde{R}$ and N, D_{av}/N is lower bounded by $\frac{1}{1+\mu N/\tilde{R}}$, where μ is a positive constant independent of N and $\tilde{R}$ ($\mu \approx 0.09332$).*

Note that the factor $N/\tilde{R}$ in the lower bound can be seen as a *one-dimensional node density*. Theorem 1.3 indicates that the node density almost uniquely determines the normalized directivity D_{av}/N. It is important to note that in order to achieve a certain normalized directivity with a large number of nodes N, the node density should be maintained to the desired value by spreading the nodes as sparsely as possible. The above theorem also indicates that in order to achieve high directivity (i.e. directivity close to N), the distribution of nodes should be as sparse as possible.

Distribution of Far-Field Beampattern of Cooperative Beamforming

For random arrays, the above average behavior does not necessarily approximate a beampattern of any given realization unless $N \to \infty$. In fact, even though the average beampattern has a sharp mainbeam and sidelobes always close to $1/N$, there is a large dynamic range of sidelobes among randomly generated beampatterns. Therefore, in practice, the statistical *distribution* of beampatterns and sidelobes in particular, is of interest. By approximating the beampattern sidelobes as a complex Gaussian process, Lo [Lo, 1964] has derived the distribution of the beampattern in the case of linear random arrays. In the context of cooperative beamforming, we briefly summarize the exact complementary CDF (CCDF) of the beampattern and a Gaussian approximation of it similar to [Lo, 1964].

- Exact distribution:

$$\Pr\left[P(\phi) > P_0\right] = \iint_{x^2+y^2>N^2P_0} f_{\tilde{X},\tilde{Y}}(x,y)\,dx\,dy, \tag{1.22}$$

  where kth entries of $\tilde{X}$ and $\tilde{Y}$ are, respectively, given as $\tilde{x}_k \triangleq \cos(z_k\alpha(\phi))$ and $\tilde{y}_k \triangleq \sin(z_k\alpha(\phi))$ ($k = 1, 2, \cdots, N$).

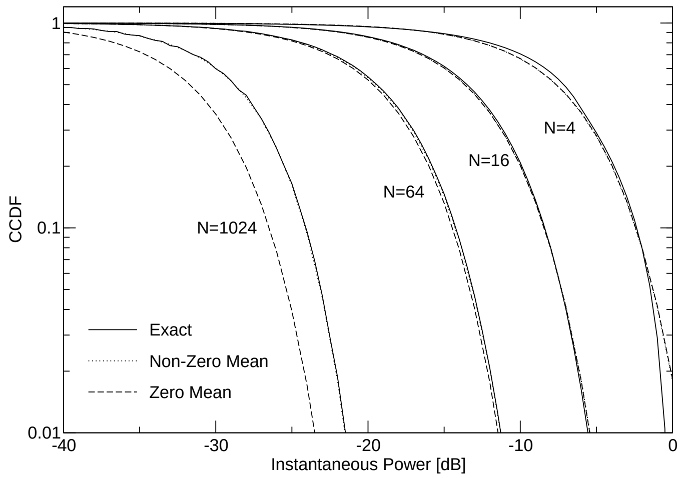


Figure 1.8. CCDF of beampattern with $\tilde{R} = 2$ and $\phi = \pi/4$.

- Gaussian approximation:

$$\Pr\left[P(\phi) > P_0\right] \approx Q\left(2\sqrt{2N}\frac{J_1\left(\alpha(\phi)\right)}{\alpha(\phi)}, \sqrt{2NP_0}\right), \tag{1.23}$$

where $Q(\cdot, \cdot)$ denotes the Marcum-Q function.

- Zero-mean Gaussian approximation:

$$\Pr\left[P(\phi) > P_0\right] \approx e^{-NP_0}. \tag{1.24}$$

In Fig. 1.8, the CCDFs computed with various formulae are shown with $\tilde{R} = 2$ and $\phi = \pi/4$, which corresponds to the sidelobe region. The exact formula of (1.22), the Gaussian approximation of (1.23), and the zero-mean Gaussian approximation of (1.24) are shown in the figure. As observed from Fig. 1.8, even the zero-mean Gaussian approximation may be valid for this sidelobe region, but for $N = 1024$ the Gaussian approximation will have some noticeable discrepancy with the exact value. This is due to the fact that the zero-mean approximation does not hold for this case. In [Ochiai et al., 2005], the distribution of the beampattern was discussed in more detail, and an approximate upper bound on the distribution of the peak sidelobes was derived. Furthermore, [Ochiai et al., 2005] also considered both open-loop and closed-loop scenarios and investigated the effects of phase ambiguities and location estimation errors on the resultant average beampatterns.

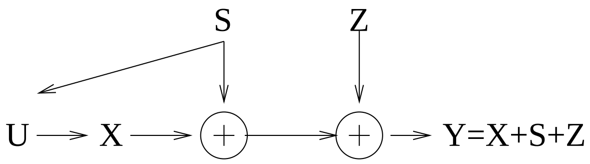


Figure 1.9. Dirty paper coding channel with input X, auxiliary random variable U, interference S known non-causally to the transmitter, additive noise Z and output Y.

4. Cognitive Radio

Preliminaries

A key idea behind achieving high data rates in an environment where two senders share a common channel is interference cancellation or mitigation. When side-information is known at the transmitter only, the channel capacity is given by the well-known formula obtained by Gel'fand and Pinsker [Gel'fand and Pinsker, 1980] as

$$C = \max_{p(u,x|s)} \left[I(U;Y) - I(U;S) \right], \tag{1.25}$$

where X is the input to the channel, Y is the output, S is the interference, and U is an auxiliary random variable chosen to make the channel $U \to Y$ appear causal. The channel model and variables are shown in Fig. 1.9 for additive interference and noise. We refer to the coding technique used in [Gel'fand and Pinsker, 1980] as Gel'fand-Pinkser coding or binning. In the Gaussian noise and interference case, Costa achieves the capacity of an interference-free channel by assuming the input X to the channel is Gaussian, and then considering an auxiliary variable U of the form $U = X + \alpha S$ for some parameter α whose optimal value is equal to the ratio of the signal power to the signal plus noise power. Since the rate thus obtained is equal to the capacity of an interference-free channel, which provides an upper bound, optimality is achieved by the assumed Gaussian input X. *Dirty paper coding* is the term first used by Costa [Costa, 1983] to describe a technique which completely mitigates *a-priori* known interference over an input power constrained additive white Gaussian noise channel. We will make use of the coding techniques of Costa [Costa, 1983], Gel'fand and Pinsker [Gel'fand and Pinsker, 1980], as well as Cover and Chiang [Cover and Chiang, 2002] in Section 4.0.

The cognitive radio channel is also closely related to the interference channel, which is briefly described next. Consider a discrete memoryless *interference channel* [Carleial, 1978], with random variables $X_1 \in \mathcal{X}_1$, $X_2 \in \mathcal{X}_2$ as inputs to the channel characterized by the conditional probabilities $p(y_1|x_1, x_2)$, $p(y_2|x_1, x_2)$ with resulting channel output random variables $Y_1 \in \mathcal{Y}_1$, $Y_2 \in \mathcal{Y}_2$. The interference channel corresponds to two independent senders $\mathcal{S}_1$, $\mathcal{S}_2$,

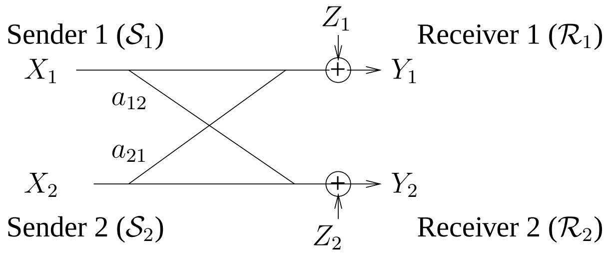


Figure 1.10. The additive interference channel with inputs X_1, X_2, outputs Y_1, Y_2, additive noise Z_1, Z_2 and interference coefficients a_{12}, a_{21}.

with independent non-cooperating receivers $\mathcal{R}_1, \mathcal{R}_2$, transmitting over the same channel. The additive interference channel is shown in Fig. 1.10, where the parameters a_{12}, a_{21} capture the effects of the interference. In addition to the additive interference from the other sender, each output is affected by independent additive noise Z_1, Z_2.

The interference channel capacity, in the most general case, is still an open problem. In the case of strong interference, as defined in [Han and Kobayashi, 1981, Sato, 1981], and very strong interference, as defined in [Carleial, 1978], the capacity is known. Achievable regions of the interference channel have been calculated in [Han and Kobayashi, 1981], and recently in [Sason, 2004]. We will make use of techniques as in [Han and Kobayashi, 1981], merged with Gel'fand-Pinsker coding [Gel'fand and Pinsker, 1980] to provide an achievable region for the cognitive radio channel.

Genie-Aided Cognitive Radio Channel

We define a *cognitive radio channel* to be an interference channel in which $\mathcal{S}_2$ has knowledge of the message to be transmitted by $\mathcal{S}_1$. This is either obtained causally, or could possibly be given to the sender non-causally by a *genie*. We first focus on the non-causal scenario, i.e., *genie-aided cognitive radio channel* C_{COG}. $\mathcal{S}_2$ can exploit the knowledge of $\mathcal{S}_1$'s message, and potentially improve the transmission rate. It can do so using a dirty paper coding technique [Costa, 1983] and an achievable region construction for the interference channel [Han and Kobayashi, 1981]. Intuitively, the achievable region in [Han and Kobayashi, 1981] should lie entirely within the achievable region of C_{COG}, since senders are permitted to at least partially cooperate. An upper bound for our region in the Gaussian case is provided by the 2×2 MIMO broadcast channel whose capacity has recently been calculated in [Weingarten et al., 2004]. In [Weingarten et al., 2004], dirty paper coding techniques are shown to be optimal for non-degraded vector broadcast channels. The cognitive radio channel model resembles that of [Weingarten et al., 2004], with

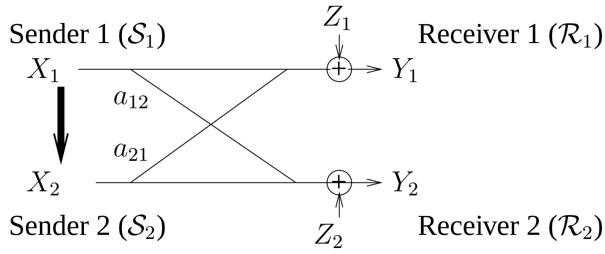


Figure 1.11. The additive interference genie-aided cognitive radio channel with inputs X_1, X_2, outputs Y_1, Y_2, additive noise Z_1, Z_2 and interference coefficients a_{12}, a_{21}. S_1's input X_1 is given to S_2 (indicated by the arrow), but not the other way around.

one important difference: the relation between the two senders is asymmetric. The rate of $\mathcal{S}_2$ is also bounded by the rate achievable in an interference-free channel, with $a_{12} = 0$.

An (n, K_1, K_2, ϵ) *code* for the *genie-aided cognitive radio channel* consists of K_1 codewords $x_1^n(i) \in \mathcal{X}_1^n$ for $\mathcal{S}_1$, and $K_1 \cdot K_2$ codewords $x_2^n(i,j) \in \mathcal{X}_2^n$ for $\mathcal{S}_2$, $i \in \{1, 2, \ldots, K_1\}$, $j \in \{1, 2, \ldots, K_2\}$, which together form the *codebook*, revealed to both senders and receivers such that the average error probabilities under some decoding scheme are less than λ.

DEFINITION 1.4 *(Achievable Rate and Region) A rate pair* (R_1, R_2) *is said to be* achievable *for the genie-aided cognitive radio channel if there exists a sequence of* $(n, 2^{\lceil nR_1 \rceil}, 2^{\lceil nR_2 \rceil}, \epsilon_n)$ *codes such that* $\epsilon_n \to 0$ *as* $n \to \infty$. *An* achievable region *is a closed subset of the positive quadrant of* $\mathbb{R}^2$ *of achievable rate pairs.*

The Modified Genie-Aided Cognitive Channel C_{COG}^m

As in [Han and Kobayashi, 1981], we introduce a modified genie-aided cognitive radio channel, C_{COG}^m, (m for modified) and demonstrate an achievable region for C_{COG}^m. Then, a relation between achievable rates for C_{COG}^m and C_{COG} is used to establish an achievable region for the latter. The modified genie-aided cognitive radio channel C_{COG}^m is defined in Fig. 1.12, and we reuse the notation of the interference channel.

The modified genie-aided cognitive radio channel introduces two pairs of auxiliary random variables: (M_1, N_1) and (M_2, N_2). The random variables $M_1 \in \mathcal{M}_1$ and $M_2 \in \mathcal{M}_2$ represent, as in [Han and Kobayashi, 1981], the private information to be sent from $\mathcal{S}_1 \to \mathcal{R}_1$ and $\mathcal{S}_2 \to \mathcal{R}_2$ respectively. In contrast, the random variables $N_1 \in \mathcal{N}_1$ and $N_2 \in \mathcal{N}_2$ represent the public information to be sent from $\mathcal{S}_1 \to (\mathcal{R}_1, \mathcal{R}_2)$ and $\mathcal{S}_2 \to (\mathcal{R}_1, \mathcal{R}_2)$ respectively. The function of these M_1, N_1, M_2, N_2 is as in [Han and Kobayashi, 1981]:

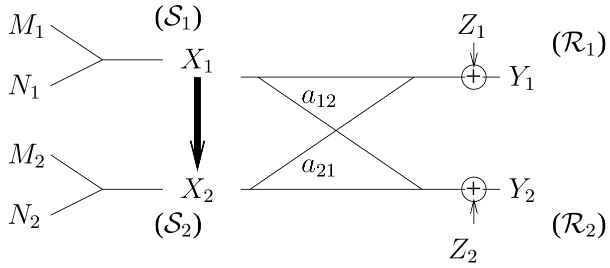


Figure 1.12. The modified cognitive radio channel with auxiliary random variables M_1, M_2, N_1, N_2, inputs X_1, X_2, additive noise Z_1, Z_2, outputs Y_1, Y_2 and interference coefficients a_{12}, a_{21}.

to decompose or define *explicitly* the information to be transmitted between various input and output pairs.

In C_{COG}^m, M_2 and N_2 also serve a dual purpose: these auxiliary random variables are analogous to the auxiliary random variables of Gel'fand and Pinsker [Gel'fand and Pinsker, 1980] or Cover and Chiang [Cover and Chiang, 2002]. They serve as fictitious inputs to the channel, so that after $\mathcal{S}_2$ is informed of the encoded message of $\mathcal{S}_1$ non-causally, the channel still behaves like a Discrete Memoryless Channel (DMC) from $(M_1, N_1, M_2, N_2) \rightarrow (Y_1, Y_2)$. As in [Cover and Chiang, 2002, Gel'fand and Pinsker, 1980], there is a penalty in using this approach which will be reflected by a reduction in achievable rates (compared to the fictitious DMC from (M_1, N_1, M_2, N_2) to (Y_1, Y_2)) for the links which use the non-causal information.

A code and an achievable region for C_{COG}^m can be defined similarly to those in C_{COG}. As mentioned in [Han and Kobayashi, 1981], the introduction of a time-sharing random variable W is thought to strictly extend the achievable region obtained using a convex hull operation. Thus, let $W \in \mathcal{W}$ be a time-sharing random variable whose n-sequences $w^n \stackrel{\triangle}{=} (w^{(1)}, w^{(2)}, \ldots, w^{(n)})$ are generated independently of the messages, according to $\prod_{t=1}^n p(w^{(t)})$. The n-sequence w^n is given to both senders and both receivers. The following theorem and corollary on achievable rates for C_{COG}^m were proved in [Devroye et al., 2004].

THEOREM 1.5 *Let $Z \stackrel{\triangle}{=} (Y_1, Y_2, X_1, X_2, M_1, N_1, M_2, N_2, W)$, and let $\mathcal{P}$ be the set of distributions on Z that can be decomposed into the form*

$$p(w)p(m_1|w)p(n_1|w)p(x_1|m_1, n_1, w)p(m_2|x_1, w)p(n_2|x_1, w)$$
$$\times\, p(x_2|m_2, n_2, w)p(y_1|x_1, x_2)p(y_2|x_1, x_2).$$

For any $Z \in \mathcal{P}$, let $S(Z)$ be the set of all quadruples $(R_{11}, R_{12}, R_{21}, R_{22})$ of non-negative real numbers such that there exist non-negative real (L_{21}, L_{22})

satisfying:

$$
\begin{aligned}
R_{11} &\leq I(M_1; X_1 | N_1, W) \\
R_{12} &\leq I(N_1; X_1 | M_1, W) \\
R_{11} + R_{12} &\leq I(M_1, N_1; X_1 | W)
\end{aligned}
$$

$$
\begin{aligned}
R_{21} &\leq L_{21} - I(N_2; M_1, N_1 | W) \\
R_{22} &\leq L_{22} - I(M_2; M_1, N_1 | W)
\end{aligned}
$$

$$
\begin{aligned}
R_{11} &\leq I(Y_1, N_1, N_2; M_1 | W) \\
R_{12} &\leq I(Y_1, M_1, N_2; N_1 | W) \\
L_{21} &\leq I(Y_1, M_1, N_1; N_2 | W) \\
R_{11} + R_{12} &\leq I(Y_1, N_2; M_1, N_1 | W) \\
R_{11} + L_{21} &\leq I(Y_1, N_1; M_1, N_2 | W) \\
R_{12} + L_{21} &\leq I(Y_1, M_1; N_1, N_2 | W) \\
R_{11} + R_{12} + L_{21} &\leq I(Y_1; M_1, N_1, N_2 | W)
\end{aligned}
$$

$$
\begin{aligned}
L_{22} &\leq I(Y_2, N_1, N_2; M_2 | W) \\
R_{12} &\leq I(Y_2, N_2, M_2; N_1 | W) \\
L_{21} &\leq I(Y_2, N_1, M_2; N_2 | W) \\
L_{22} + L_{21} &\leq I(Y_2, N_1; M_2, N_2 | W) \\
L_{22} + R_{12} &\leq I(Y_2, N_2; M_2, N_1 | W) \\
R_{12} + L_{21} &\leq I(Y_2, M_2; N_1, N_2 | W) \\
L_{22} + R_{21} + L_{12} &\leq I(Y_2; M_2, N_1, N_2 | W).
\end{aligned}
$$

Let S be the closure of $\cup_{Z \in \mathcal{P}} S(Z)$. Then any element of S is achievable for the modified genie-aided cognitive radio channel C_{COG}^m. □

Another important rate pair for C_{COG}^m is achievable: that in which $\mathcal{S}_2$ transmits no information of its own to $\mathcal{R}_2$, and simply aids $\mathcal{S}_1$ in sending its message to $\mathcal{R}_1$. When this is the case, the rate pair $(R_1^*, 0)$ is achievable, where R_1^* is the capacity of the vector channel $(\mathcal{S}_1, \mathcal{S}_2) \to \mathcal{R}_1$. Note however, that the analogous rate pair $(0, R_2^*)$ is not achievable, since that would involve $\mathcal{S}_1$ aiding $\mathcal{S}_2$ in sending its message, which cannot happen under our assumptions: $\mathcal{S}_2$ knows $\mathcal{S}_1$'s message, but not vice versa. The overall achievable region is given by the following Corollary.

COROLLARY 1.6 *Let $\mathcal{C}_0$ be the set of all points $(R_{11}+R_{12}, R_{21}+R_{22})$ where $(R_{11}, R_{12}, R_{21}, R_{22})$ is an achievable rate tuple of Theorem 1.5. Consider*

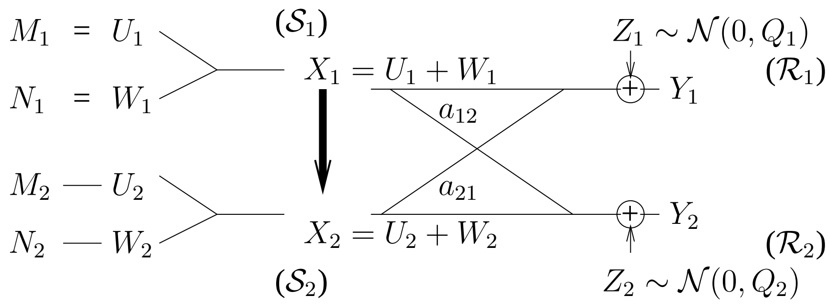


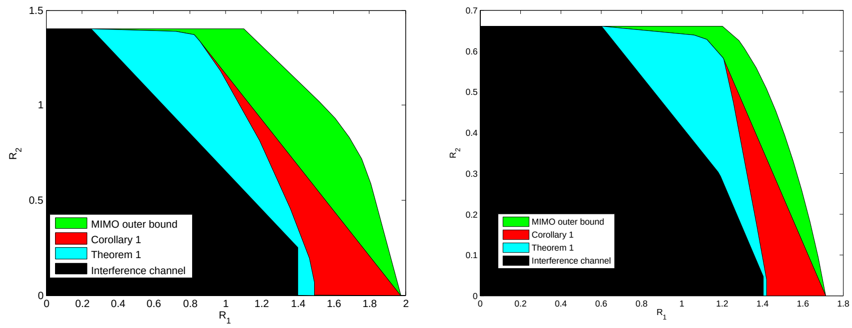
Figure 1.13. The modified Gaussian genie-aided cognitive radio channel with interference coefficients a_{12}, a_{21}.

the vector channel $(\mathcal{S}_1, \mathcal{S}_2) \rightarrow \mathcal{R}_1$ *described by the conditional probability density* $p(y_1|x_1, x_2)$ *for all* $y_1 \in \mathcal{Y}_1$, $x_1 \in \mathcal{X}_1$, $x_2 \in \mathcal{X}_2$, *and define* $R_1^* \triangleq \max_{p(x_1,x_2)} I(X_1, X_2; Y_1)$. *Then the convex hull of the region* $\mathcal{C}_0$ *with the point* $(R_1^*, 0)$ *is achievable for the genie-aided cognitive radio channel* C_{COG}^m. $\square$

The Gaussian Cognitive Radio Channel

Consider the genie-aided cognitive radio channel, depicted in Fig. 1.13 with independent additive noise $Z_1 \sim \mathcal{N}(0, Q_1)$ and $Z_2 \sim \mathcal{N}(0, Q_2)$. We assume the two transmitters are power limited to P_1 and P_2 respectively. In order to determine an achievable region for the modified Gaussian genie-aided cognitive radio channel, specific forms of the random variables described in Theorem 1.5 are assumed. As in [Costa, 1983, Gallager, 1968, Han and Kobayashi, 1981], Theorem 1.5 and its Corollary can readily be extended to memoryless channels with discrete time and continuous alphabets by finely quantizing the input, output, and interference variables (Gaussian in this case). Let the time-sharing random variable W be constant. Consider the case where, for certain $\alpha, \beta \in \mathbb{R}$ and $\lambda, \overline{\lambda}, \gamma, \overline{\gamma} \in [0, 1]$, with $\lambda + \overline{\lambda} = 1$, $\gamma + \overline{\gamma} = 1$, and additional independent auxiliary random variables U_1, W_1, U_2, W_2 as in Fig. 1.13, the following hold:

$$
\begin{aligned}
U_1 &= M_1 \sim \mathcal{N}(0, \lambda P_1) \\
W_1 &= N_1 \sim \mathcal{N}(0, \overline{\lambda} P_1) \\
X_1 &= U_1 + W_1 = M_1 + N_1 \sim \mathcal{N}(0, P_1) \\
M_2 &= U_2 + \alpha X_1 \text{ where } U_2 \sim \mathcal{N}(0, \gamma P_2) \\
N_2 &= W_2 + \beta X_1 \text{ where } W_2 \sim \mathcal{N}(0, \overline{\gamma} P_2) \\
X_2 &= U_2 + W_2 \sim \mathcal{N}(0, P_2).
\end{aligned}
$$



(a) (b)

Figure 1.14. Achievable region of [Han and Kobayashi, 1981] (innermost polyhedron), Theorem 1.5 (the next to smallest), and Corollary 1.6 (the second to largest), and the intersection of the capacity region of the 2×2 MIMO broadcast channel with the outer bound on R_2 of an interference-free Gaussian channel of capacity $1/2 \log(1 + P_2/Q_2)$ (the largest region). (a) $Q_1 = Q_2 = 1$, $a_{12} = a_{21} = 0.55$, $P_1 = P_2 = 6$. (b) $Q_1 = Q_2 = 1$, $a_{12} = a_{21} = 0.55$, $P_1 = 6$, $P_2 = 1.5$.

The achievable regions thus obtained for the Gaussian genie-aided cognitive radio channel are plotted in Fig. 1.14. The innermost region (black) corresponds to the achievable region of [Han and Kobayashi, 1981], and is obtained by setting $\alpha = \beta = 0$. As expected, because of the extra information at the encoder and the partial use of a dirty-paper coding technique, the achievable region in Theorem 1.5, the second to smallest region (cyan) in Fig. 1.14, extends that of [Han and Kobayashi, 1981]. The overall achievable region, that of Corollary 1.6, further extends that of Theorem 1.5, as seen by the second largest (red) region in Fig. 1.14. An upper bound on our achievable rate region is provided by the 2×2 Gaussian MIMO broadcast channel, whose capacity was computed in [Weingarten et al., 2004]. Here, the two senders can fully cooperate (fully symmetric system). We calculate this region for input covariance constraint matrix of the form $S = \begin{pmatrix} P_1 & c \\ c & P_2 \end{pmatrix}$, for some $-\sqrt{P_1 P_2} \leq c \leq \sqrt{P_1 P_2}$ (which ensures S is positive semi-definite), and which mimics the power constraints P_1 and P_2 on each individual sender (asymmetric problem). The largest region in Fig. 1.14 is the intersection of the 2×2 Gaussian MIMO broadcast channel capacity region with the bound on $\mathcal{S}_2$'s rate $R_2 \leq \frac{1}{2} \log(1 + P_2/Q_2)$ provided by the interference-free channel in which $a_{12} = 0$.

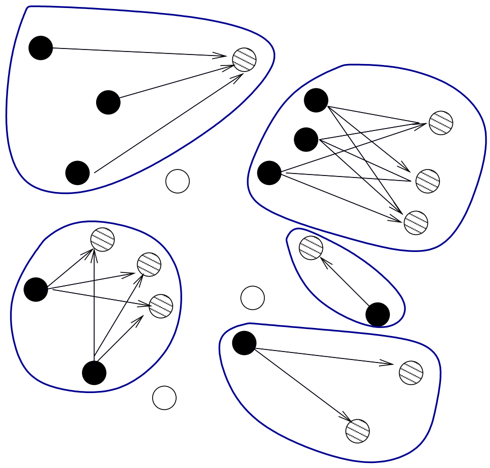


Figure 1.15. A wireless network consisting of cognitive and possibly non-cognitive devices. Black nodes are senders ($\mathbf{S_i}$), striped nodes are receivers ($\mathbf{R_i}$), and white nodes are neither (i.e., single node clusters). A directed edge is placed between each desired sender-receiver pair at each point/period in time. The graph has been partitioned into subsets of *generalized MIMO channels.*

Cognitive Radio Channel: the Causal Case

In practice, the message x_1^n that $\mathcal{S}_1$ wants to transmit cannot be non-causally given to $\mathcal{S}_2$. The transmitter $\mathcal{S}_2$ must obtain the message in real time, and one possible way to do so is by exploiting proximity to $\mathcal{S}_1$. As in Section 2, this proximity can be modeled by a reduction G in path loss, or equivalently, an increase in capacity between $\mathcal{S}_1$ and $\mathcal{S}_2$, relative to the channels between the senders and the receivers. If, for example, the channel between $\mathcal{S}_1$ and $\mathcal{S}_2$ is an AWGN channel, then the capacity would increase, for a factor $G \geq 1$, to $C = \frac{1}{2}\log\left(1 + G \cdot \frac{P_1}{Q}\right)$, where Q is the additive Gaussian noise power. Alternatively, if $\mathcal{S}_1$ and $\mathcal{S}_2$ are base-stations, then it may be possible for S_2 to obtain S_1's message through a high bandwidth wired connection (if one exists) in real time. In the Gaussian cognitive radio channel model, all receivers know the channel between themselves and the relevant sender(s). In addition, both senders and receivers know the interference channel parameters a_{12} and a_{21}. In [Devroye et al., 2004], several protocols that allow $\mathcal{S}_2$ to causally obtain $\mathcal{S}_1$'s message were proposed, and corresponding achievable regions were derived. Three of them use a two-phase approach as in Section 2. Details of the protocols and achievable regions can be found in [Devroye et al., 2004]. We should note that the genie-aided cognitive radio channel achievable region provides an outer bound on a causal achievable region which uses the same coding strategy.

Cognitive Multiple Access Networks

In the previous section, we provided an achievable region for a two sender, two receiver *cognitive radio channel*. The achievable region was extended to *cognitive multiple access networks* in [Devroye et al., 2005]. We consider an arbitrary wireless network consisting of cognitive and possibly non-cognitive radio devices, as illustrated in Fig. 1.15. At each point/period in time, certain devices in sending mode wish to transmit to other devices in receiving mode. At each point/period in time, the wireless network can be represented as a directed graph by drawing a directed edge between every sender-receiver pair, as in Fig. 1.15. We define a *generalized MIMO channel* $(\mathbf{S}, \mathbf{R})$ as a connected bipartite directed graph where each sender node in $\mathbf{S}$ transmits to a subset of the receiver nodes in $\mathbf{R}$, and the channel is fully described by the conditional probability $P(\mathbf{r}|\mathbf{s})$. The *generalized MIMO channel* reduces to well-studied channels in certain cases. When a cluster consists of a single sender, it becomes a broadcast channel. When a cluster consists of a single receiver, it becomes a multiple access channel (MAC). When all receivers in a cluster are connected to all senders in a cluster, we have a vector MAC. The following lemma can readily be proved [Devroye et al., 2005].

LEMMA 1.7 *Any cognitive network can be partitioned into a set of* generalized MIMO channels $(\mathbf{S_i}, \mathbf{R_i})$ *where each sender node in* $\mathbf{S_i}$ *only transmits to a subset of the receiver nodes* $\mathbf{R_i}$. □

As described in the Introduction, after partitioning a wireless network into clusters, one can consider both *inter* and *intra* cluster competitive, cognitive, and cooperative behavior. In a cognitive multiple access network, clusters consist of classical information theoretic multiple access channels. The capacity region of the *intra-cluster cognitive MAC* has previously been considered in the classic information theoretic context of [Van der Meulen, 1971, Van der Meulen, 1977]. In [Devroye et al., 2005], an achievable region for two MAC channel clusters that simultaneously transmit and interfere has been computed in the case that one MAC cluster knows the messages to be sent by the other MAC cluster. In the Gaussian case, [Devroye et al., 2005] also numerically evaluated an achievable region for cognitive behavior and compared it to the achievable regions under competitive behavior as well as cooperative behavior.

5. Summary and Remarks

In this chapter, we have developed a general framework of wireless networks in the context of competition, cooperation and cognition. For the cooperation paradigm, we have provided a bandwidth efficient approach to compound Gaussian relay channels, and shown the existence of a cooperative code which is good over wide range of channels. We have also presented the design of a cooperative diversity system on an OFDM platform to demonstrate cooperative diversity gains in practical wireless systems. As an alternative to cooperative diversity, we have also introduced cooperative beamforming concepts along with their respective performance analyses. For cognitive radio channels, we have defined a more flexible and potentially more efficient transmission model, and constructed an achievable region. Finally, we have discussed extensions of the idea to cognitive multiple access networks.

We conclude by discussing some research opportunities in this emerging field. One possible extension of the cooperative diversity scheme is to investigate the effect of full asynchronism between nodes. Another extension is to investigate more refined cooperation on part of the relays. A number of open issues remain for cooperative beamforming as well, such as applicability of beamforming when the destination or nodes in the cluster are in rapid motion or the channel suffers severe multipath fading. There is still a lot of work to do in the context of practical implementation, such as development of more efficient protocols, and synchronization and channel estimation algorithms. From the viewpoint of higher layer protocols, development of clustering protocols and link layer error control protocols is an important topic for both approaches.

As the cognitive radio channel, which captures the essence of asymmetric cooperation, has only recently been introduced, numerous promising research directions exist. From a theoretical perspective, the capacity of this channel, as well as its causal version are still open problems. Some achievable regions have already been calculated, and the development of tight upper bounds on the cognitive and causal cognitive radio channels will advance the field towards this final goal. Extensions of the cognitive radio channel to fading and compound channels is another research area to be explored. Practically, coding protocols and schemes that enable cognitive transmission must be devised, and issues similar to those encountered in full cooperation, such as synchronization and channel estimation, will naturally arise here as well.

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