Optimal Bit Allocation of Limited Rate Feedback for Cooperative Jamming

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***Abstract*—In this paper, we investigate bit allocation schemes with limited rate feedback for cooperative jamming. In addition to the transmitter and receiver, we assume a passive eavesdropper and cooperative jammer are present. In order to achieve a secure communications link against the eavesdropper, the transmitter and jammer require channel state information (CSI) to be fed back to them from the receiver. Assuming feedback channels with a maximum sum feedback rate constraint, the receiver must allocate the total number of bits available to quantize the CSI between the transmitter and jammer. This requires the receiver to balance the need for a strong channel from the transmitter against the need for the jammer to accurately null the receiver and reduce the resulting interference. We propose an optimal bit allocation strategy for this problem using mean-squared error as the performance metric, and we use simulation examples to illustrate its advantage over a non-optimized feedback allocation.**

I. Introduction

Classical results in information theoretic secrecy have demonstrated that secret communication is possible if the eavesdropper's channel is degraded compared to that of the legitimate receiver [1], [2]. Since this condition cannot be guaranteed in general, attention has focused recently on the use of multiple antennas as a means of either enhancing the channel of the desired user through beamforming, or selectively degrading the eavesdropper's channel through artificial interference or jamming. The former approach requires knowledge of the eavesdropper's channel, which is a problematic assumption in many cases, while the latter can be beneficial if only limited information (or even no information) is available about the eavesdropper [3], [4].

The use of jamming from cooperating transmitters has been discussed in [5]–[12]. In most works, multiple antennas are used to mitigate the effect of the jamming on the desired receiver, but this requires accurate channel state information (CSI). In practice, perfect CSI is not possible at the cooperative jammer due to channel estimation errors in time division duplex (TDD) systems or limited feedback in the frequency division duplex (FDD) case. The design of limited feedback schemes and their performance compared to the perfect CSI case has received considerable attention for standard wireless applications (for example, see [13]), but little work has been done on the impact of limited feedback for the wiretap channel [14], [15]. In this paper, we consider the problem of limited feedback design for cooperative jamming systems, where both the data transmitter and jamming transmitter require CSI. This problem is interesting when the feedback bandwidth is limited, and the total number of feedback bits must be properly allocated between the two transmitters. In this case, one must balance the need for good beamforming gain from the data transmitter against interference leakage from the cooperative jammer. Using the mean-squared error (MSE) at the desired receiver as our performance metric, we describe an approach for obtaining an optimal feedback bit allocation. While a simple optimization is required in the general case, we show that a closed-form solution can be obtained when the transmitter and jammer precoding matrices have the same number of elements. Numerical results demonstrate the advantage of using the optimal feedback bit allocation approach.

The remainder of this paper is organized as follows. The system model and assumptions are introduced in Section II. The beamformer design problem is presented in Section III, and the optimal bit allocation algorithm is derived in Section IV. Simulation results are presented in Section V. Throughout the paper we use lowercase boldface letters to denote vectors and uppercase bold letters to denote matrices. The space of $m \times n$ complex matrices is denoted by $\mathbb{C}^{m\times n}$. The Hermitian transpose is represented by $(\cdot)^H$, $\|\cdot\|$ the Euclidean (Frobenius) norm, $\mathbb{E}[\cdot]$ the expectation operator, $tr(\cdot)$ the matrix trace.

II. System Model and Assumptions

The scenario under consideration is depicted in Fig. 1, and is comprised of a transmitter (Alice) with N_a antennas, an intended receiver (Bob) with N_b antennas, and a cooperative jammer (Helper) with N_h antennas. While the Helper is present to provide artificial interference to degrade the channel of any eavesdropper that may be present, the parameters of the eavesdropper are assumed to be completely unknown. We also make the following additional assumptions:

- Bob has perfect channel estimation, but Alice and the Helper do not know Bob's channel.
- $N_a > N_b$ and $N_h > N_b$.
- The feedback channels from Bob to Alice and the Helper are error-free and the total feedback rate is limited.
- All channels are assumed to experience independent block fading.

The channels from Alice and the Helper to Bob are respectively given by $\mathbf{H}_{ba} \in \mathbb{C}^{N_b\times N_a}$ and $\mathbf{H}_{bh} \in \mathbb{C}^{N_b\times N_h}$, and the elements of these matrices are assumed to be independent and identically distributed (i.i.d.) and have a circularly symmetric complex Gaussian distribution with zero mean and unit variance. We further assume that $\mathbf{H}_{ba}$ and $\mathbf{H}_{bh}$ are of

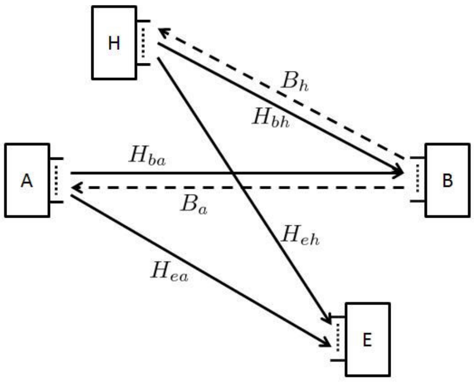

Fig. 1. System model.

full rank N_b, and that Alice transmits a single data stream s. On the other hand, the Helper transmits a d-dimensional jamming signal $\mathbf{v}$. The elements of s and $\mathbf{v}$ are zero-mean Gaussian with variance 1 and $1/d$, respectively. The jamming dimension d is constrained to be no greater than $N_h - N_b$, so that it can be designed to be orthogonal to the information signal at Bob (although the imperfect CSI will not allow this to be exactly true). We define Alice's precoder (beamformer) as $\tilde{\mathbf{w}}_a \in \mathbb{C}^{N_a \times 1}$ and the jamming precoder as $\tilde{\mathbf{W}}_h \in \mathbb{C}^{N_h \times d}$; both are normalized to have unit Frobenius norm. The transmitted power at Alice is given by P_S and that at the Helper is P_I. With these assumptions, the signals transmitted by Alice and the Helper are:

$$\mathbf{x} = \sqrt{P_S}\tilde{\mathbf{w}}_a s, \tag{1}$$

$$\mathbf{z} = \sqrt{P_I}\tilde{\mathbf{W}}_h \mathbf{v}. \tag{2}$$

The signal received by Bob is thus

$$\mathbf{y} = \sqrt{P_S}\mathbf{H}_{ba}\tilde{\mathbf{w}}_a s + \sqrt{P_I}\mathbf{H}_{bh}\tilde{\mathbf{W}}_h \mathbf{v} + \mathbf{n}\ ,$$

where the components of $\mathbf{n}$ are i.i.d. Gaussian noise with zero mean and unit variance. Assuming that Bob uses a linear receiver $\mathbf{r}^H \in \mathbb{C}^{1 \times N_b}$, the estimated data symbol is

$$\begin{aligned}\widehat{s} &= \frac{1}{\sqrt{P_S}}\mathbf{r}^H \mathbf{y} \\ &= \mathbf{r}^H \left(\mathbf{H}_{ba}\tilde{\mathbf{w}}_a s + \sqrt{\frac{P_I}{P_S}}\mathbf{H}_{bh}\tilde{\mathbf{W}}_h \mathbf{v} + \frac{1}{\sqrt{P_S}}\mathbf{n}\right). \end{aligned} \tag{3}$$

Bob designs "quantization-error-free" precoders $\mathbf{w}_a$ and $\mathbf{W}_h$ based on perfect knowledge of $\mathbf{H}_{ba}$ and $\mathbf{H}_{bh}$. Bob then quantizes $\mathbf{w}_a$ and $\mathbf{W}_h$ by selecting precoders from codebooks containing 2^{B_a} and 2^{B_h} elements. The B_a and B_h bits corresponding to the chosen precoders are separately fed back to Alice and the Helper. Since the total rate of the feedback channels is limited, Bob is only able to send a fixed total number of bits B:

$$B_a + B_h = B.$$

Let $\widehat{\mathbf{w}}_a$ and $\widehat{\mathbf{W}}_h$ represent the quantized signal precoders, and let $\Delta\mathbf{w}_a$ and $\Delta\mathbf{W}_h$ be the quantization errors:

$$\widehat{\mathbf{w}}_a \triangleq \mathbf{w}_a + \Delta\mathbf{w}_a, \tag{4}$$

$$\widehat{\mathbf{W}}_h \triangleq \mathbf{W}_h + \Delta\mathbf{W}_h. \tag{5}$$

Due to the power constraints, the actual precoders used by Alice and the Helper are normalized versions of the quantized precoders:

$$\tilde{\mathbf{w}}_a = \frac{1}{\|\widehat{\mathbf{w}}_a s\|}\widehat{\mathbf{w}}_a, \tag{6}$$

$$\tilde{\mathbf{W}}_h = \frac{1}{\left\|\widehat{\mathbf{W}}_h \mathbf{v}\right\|}\widehat{\mathbf{W}}_h. \tag{7}$$

III. Beamformer Design

The mean-squared error (MSE) associated with the estimate $\widehat{s}$ is

$$\text{MSE} \triangleq \mathbb{E}\left[|\widehat{s} - s|^2\right]. \tag{8}$$

According to (3), the instantaneous MSE is

$$\begin{aligned}\text{MSE}(\tilde{\mathbf{w}}_a, \tilde{\mathbf{W}}_h, \mathbf{r}) &= \left|\mathbf{r}^H \mathbf{H}_{ba}\tilde{\mathbf{w}}_a - 1\right|^2 + \mathbf{r}^H \mathbf{R}_n \mathbf{r} \\ &= \mathbf{r}^H \left(\mathbf{H}_{ba}\tilde{\mathbf{w}}_a \tilde{\mathbf{w}}_a^H \mathbf{H}_{ba}^H + \mathbf{R}_n\right)\mathbf{r} + 1 \\ &\quad - \mathbf{r}^H \mathbf{H}_{ba}\tilde{\mathbf{w}}_a - \tilde{\mathbf{w}}_a^H \mathbf{H}_{ba}^H \mathbf{r} \end{aligned} \tag{9}$$

where

$$\mathbf{R}_n = \frac{P_I}{P_S d}\mathbf{H}_{bh}\tilde{\mathbf{W}}_h \tilde{\mathbf{W}}_h^H \mathbf{H}_{bh}^H + \frac{1}{P_S}\mathbf{I}.$$

A. Optimal Decoder

Given $\tilde{\mathbf{w}}_a$ and $\tilde{\mathbf{W}}_h$, the optimal decoder $\mathbf{r}^*$ is easily found by setting the gradient of the MSE to zero, which yields the minimum MSE (MMSE) receiver

$$\mathbf{r}^* = \left[\mathbf{H}_{ba}\tilde{\mathbf{w}}_a \tilde{\mathbf{w}}_a^H \mathbf{H}_{ba}^H + \mathbf{R}_n\right]^{-1}\mathbf{H}_{ba}\tilde{\mathbf{w}}_a.$$

Using the optimal decoder, the MSE expression in (9) reduces to

$$\begin{aligned}\text{MSE}(\tilde{\mathbf{w}}_a, \tilde{\mathbf{W}}_h) &= \text{MSE}(\tilde{\mathbf{w}}_a, \tilde{\mathbf{W}}_h, \mathbf{r}^*) \\ &= \frac{1}{1 + \tilde{\mathbf{w}}_a^H \mathbf{H}_{ba}^H \mathbf{R}_n^{-1}\mathbf{H}_{ba}\tilde{\mathbf{w}}_a} \end{aligned} \tag{10}$$

where the second expression is obtained after applying the matrix inversion lemma.

B. Quantization-Error-Free Precoders

Bob designs the quantization-error-free precoders $\mathbf{w}_a$ and $\mathbf{W}_h$ that minimize the instantaneous MSE

$$\min_{\tilde{\mathbf{w}}_a, \tilde{\mathbf{W}}_h} \quad \frac{1}{1 + \tilde{\mathbf{w}}_a^H \mathbf{H}_{ba}^H \mathbf{R}_n^{-1}\mathbf{H}_{ba}\tilde{\mathbf{w}}_a},$$

subject to unit norm constraints on $\tilde{\mathbf{w}}_a$ and $\tilde{\mathbf{W}}_h$. The optimal solution to this optimization problem is given by

$$\mathbf{w}_a = [\mathbf{V}_a]_{(:,1)}, \tag{11}$$

$$\mathbf{W}_h = [\mathbf{V}_h]_{(:,N_h-d+1:N_h)}, \tag{12}$$

where $[\mathbf{V}_a]_{(:,1)}$ denotes the first right singular vector of $\mathbf{H}_{ba}$, and $[\mathbf{V}_h]_{(:,N_h-d+1:N_h)}$ denotes the last d right singular vectors of $\mathbf{H}_{bh}$. The beamformer $\mathbf{w}_a$ aligns the transmitted signal along the strongest channel mode, so that $\mathbf{H}_{ba}\mathbf{w}_a = \lambda_p \mathbf{u}_1$, where $\mathbf{u}_1$ is the first left singular vector of $\mathbf{H}_{ba}$ and λ_p is the principal singular value of $\mathbf{H}_{ba}$. The precoder $\mathbf{W}_h$ satisfies $\mathbf{H}_{bh}\mathbf{W}_h = \mathbf{0}$.

C. Quantized Precoders

Bob quantizes $\mathbf{w}_a$ and $\mathbf{W}_h$ to $\widehat{\mathbf{w}}_a$ and $\widehat{\mathbf{W}}_h$, and sends the indices corresponding to the closest codebook elements as feedback. The actual precoders used by Alice and the Helper are normalized to meet the power constraints:

$$\tilde{\mathbf{w}}_a = \frac{1}{\sqrt{\widehat{\mathbf{w}}_a^H \widehat{\mathbf{w}}_a}} \widehat{\mathbf{w}}_a, \tag{13}$$

$$\tilde{\mathbf{W}}_h = \frac{1}{\sqrt{tr(\widehat{\mathbf{W}}_h^H \widehat{\mathbf{W}}_h)}} \widehat{\mathbf{W}}_h. \tag{14}$$

IV. Optimal Bit Allocation Algorithm

Plugging (4) and (5) into (13) and (14) yields

$$\tilde{\mathbf{w}}_a = \frac{\mathbf{w}_a + \Delta\mathbf{w}_a}{\sqrt{1 + \mathbf{w}_a^H \Delta\mathbf{w}_a + \Delta\mathbf{w}_a^H \mathbf{w}_a + \boldsymbol{\Delta}\mathbf{w}_a^H \boldsymbol{\Delta}\mathbf{w}_a}}$$

$$\tilde{\mathbf{W}}_h = \frac{\mathbf{W}_h + \Delta\mathbf{W}_h}{\sqrt{tr(\mathbf{I} + \mathbf{W}_h^H \Delta\mathbf{W}_h + \Delta\mathbf{W}_h^H \mathbf{W}_h + \Delta\mathbf{W}_h^H \Delta\mathbf{W}_h)}},$$

and substituting the above expressions into (10) gives the following expression for the instantaneous MSE:

$$\text{MSE} = \frac{1}{1 + \frac{(\mathbf{w}_a + \Delta\mathbf{w}_a)^H \mathbf{H}_{ba}^H \mathbf{R}_n^{-1} \mathbf{H}_{ba} (\mathbf{w}_a + \Delta\mathbf{w}_a)}{1 + \mathbf{w}_a^H \Delta\mathbf{w}_a + \Delta\mathbf{w}_a^H \mathbf{w}_a + \Delta\mathbf{w}_a^H \Delta\mathbf{w}_a}},$$

where

$$\mathbf{R}_n = \frac{P_I \mathbf{H}_{bh} \Delta\mathbf{W}_h \Delta\mathbf{W}_h^H \mathbf{H}_{bh}^H}{tr(\mathbf{I} + \mathbf{W}_h^H \Delta\mathbf{W}_h + \Delta\mathbf{W}_h^H \mathbf{W}_h + \Delta\mathbf{W}_h^H \Delta\mathbf{W}_h) P_S d} + \frac{1}{P_S}\mathbf{I}.$$

Since the codebook size is fixed for all the channel realizations, the optimal bit allocation will be based on the MSE averaged over the channels. When B is large, the entries of $\Delta\mathbf{w}_a$ and $\Delta\mathbf{W}_h$ can be modeled as zero-mean circularly symmetric complex Gaussian random variables, independent of $\mathbf{w}_a$ and $\mathbf{W}_h$, with i.i.d. components each having variance $\sigma_{\Delta a}^2$ and $\sigma_{\Delta h}^2$, respectively. We assume a codebook design that minimizes mean squared quantization error

$$\sigma_{\Delta a}^2 = \frac{1}{N_a} \sum_{i=1}^{N_a} \mathbb{E}\left[|(\Delta\mathbf{w}_a)_i|^2 \right]$$

$$\sigma_{\Delta h}^2 = \frac{1}{N_h d} \sum_{j=1}^{d} \sum_{i=1}^{N_h} \mathbb{E}\left[|(\Delta\mathbf{W}_h)_{ij}|^2 \right].$$

The mean squared quantization errors are lower-bounded by the corresponding rate-distortion functions [16], i.e.,

$$\sigma_{\Delta a}^2 \geq D(R_a) = \sigma_a^2 2^{-R_a}$$
$$\sigma_{\Delta h}^2 \geq D(R_h) = \sigma_h^2 2^{-R_h}$$

where $R_a = \frac{B_a}{N_a}$ and $R_h = \frac{B_h}{N_h d}$ are the number of descriptive bits per element of $\mathbf{w}_a$ and $\mathbf{W}_h$. Because $\mathbf{w}_a$ and $\mathbf{W}_h$ are unit norm from (11) and (12), we have $\mathbb{E}[\mathbf{w}_a^H \mathbf{w}_a] = \sigma_a^2 N_a = 1$ and $\mathbb{E}[\mathbf{W}_h^H \mathbf{W}_h] = \sigma_h^2 N_h \mathbf{I} = \mathbf{I}$. Thus, $\sigma_a^2 = \frac{1}{N_a}$ and $\sigma_h^2 = \frac{1}{N_h}$. The MSE of the quantization errors is thus

$$\sigma_{\Delta a}^2 = \frac{1}{N_a} 2^{-\frac{B_a}{N_a}}$$
$$\sigma_{\Delta h}^2 = \frac{1}{N_h} 2^{-\frac{B_h}{N_h d}}.$$

We assume that the quantization errors are independent of the actual channels and quantization-error-free precoders. The MSE averaged over the quantization errors for a given channel realization is approximately[1]

$$\mathbb{E}_{\Delta\mathbf{w}_h}[\mathbf{R}_n] \approx \frac{P_I \sigma_{\Delta h}^2 \mathbf{H}_{bh} \mathbf{H}_{bh}^H}{P_S d (1 + N_h \sigma_{\Delta h}^2)} + \frac{1}{P_S}\mathbf{I}$$

$$\mathbb{E}_{\Delta\mathbf{w}_a, \Delta\mathbf{w}_h}[\text{MSE}] \approx \frac{1}{1 + \frac{\lambda_p^2 \mathbf{u}_1^H \left(\mathbb{E}_{\Delta\mathbf{w}_h}[\mathbf{R}_n]\right)^{-1} \mathbf{u}_1 + tr\left(\mathbf{H}_{ba}^H \left(\mathbb{E}_{\Delta\mathbf{w}_h}[\mathbf{R}_n]\right)^{-1} \mathbf{H}_{ba}\right) \sigma_{\Delta a}^2}{1 + N_a \sigma_{\Delta a}^2}}.$$

Given that the elements of $\mathbf{H}_{ba}$ and $\mathbf{H}_{bh}$ are independent of each other, the MSE averaged over all channel realizations can be approximated as

$$\mathbb{E}_{\mathbf{H}_{bh}} \mathbb{E}_{\Delta\mathbf{w}_h}[\mathbf{R}_n] \approx \frac{d + (P_I + d) N_h \sigma_{\Delta h}^2}{P_S d (1 + N_h \sigma_{\Delta h}^2)} \mathbf{I}$$

$$\overline{\text{MSE}} = \mathbb{E}_{\mathbf{H}_{ba}, \mathbf{H}_{bh}} \mathbb{E}_{\Delta\mathbf{w}_a, \Delta\mathbf{w}_h}[\text{MSE}] \approx \frac{1}{1 + \frac{P_S d (1 + N_h \sigma_{\Delta h}^2)\left(\overline{\lambda}_p^2 + N_b N_a \sigma_{\Delta a}^2\right)}{\left(d + (P_I + d) N_h \sigma_{\Delta h}^2\right)\left(1 + N_a \sigma_{\Delta a}^2\right)}},$$

where $\overline{\lambda}_p$ is the root mean square principal singular value averaged over n prior channel realizations of $\mathbf{H}_{ba}$: $\overline{\lambda}_p = \sqrt{\frac{1}{n}\sum_{i=1}^{n} \lambda_p^2(i)}$.

The optimal bit allocation problem can be expressed as

$$\min_{B_a, B_h} \quad \overline{\text{MSE}} = \frac{1}{1 + \frac{P_S d \left(1 + 2^{-\frac{B_h}{N_h d}}\right)\left(\overline{\lambda}_p^2 + N_b 2^{-\frac{B_a}{N_a}}\right)}{\left(d + (P_I + d) 2^{-\frac{B_h}{N_h d}}\right)\left(1 + 2^{-\frac{B_a}{N_a}}\right)}}$$

$$\text{s.t.} \quad B_a + B_h = B$$
$$B_a, B_h \in \mathbb{Z}^+,$$

where $\mathbb{Z}^+$ is the set of non-negative integers. To solve this problem, we first relax the integer constraint. This leads to a standard convex optimization problem and can be solved by standard methods. A numerical solution can be found by solving the following equation based on the resulting KKT conditions:

$$N_h d^2 (\overline{\lambda}_p^2 - N_b) 2^{\frac{B_h}{N_h d}} - N_a P_I N_b 2^{-\frac{B}{N_a}} 2^{\frac{B_h}{N_a}}$$
$$+ N_h d (P_I + 2d)(\overline{\lambda}_p^2 - N_b) - N_a P_I (N_b + \overline{\lambda}_p^2)$$
$$+ N_h d (P_I + d)(\overline{\lambda}_p^2 - N_b) 2^{-\frac{B_h}{N_h d}} - N_a P_I \overline{\lambda}_p^2 2^{\frac{B}{N_a}} 2^{-\frac{B_h}{N_a}} = 0,$$

where we take a positive root of the equation. A closed-form solution can be obtained when the transmitter and jammer

[1] Here we use the approximation: $\mathbb{E}[f(X)/g(X)] \approx \mathbb{E}[f(X)]/\mathbb{E}[g(X)]$.

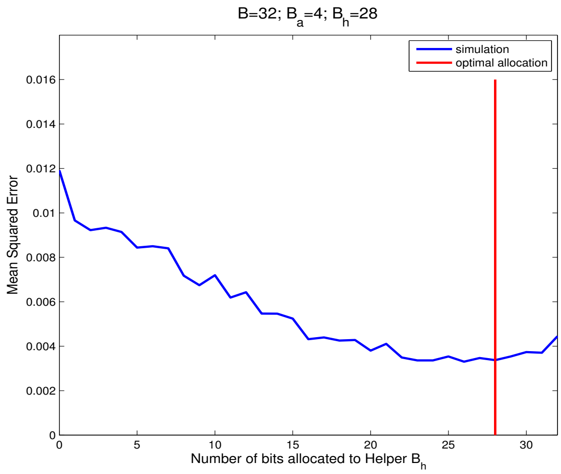

Fig. 2. Average MSE versus B_h for $B = 32$. The proposed allocation is in red.

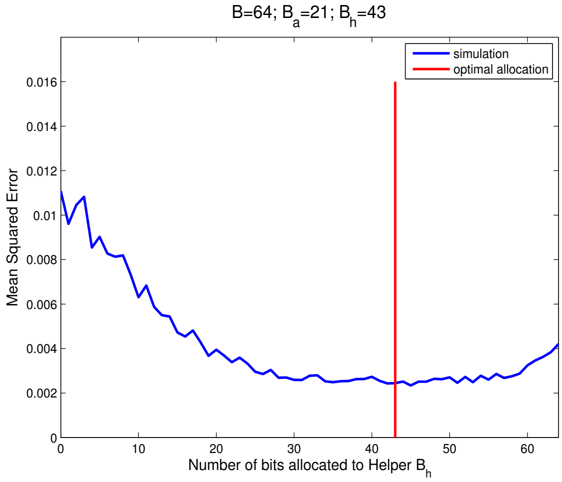

Fig. 3. Average MSE versus B_h for $B = 64$. The proposed allocation is in red.

precoders have an equal number of elements: $N_a = N_h d = N$. In this case, the number of bits allocated to the Helper is

$$B_h = -N \log_2 \frac{(P_I+d)N_b - d\bar{\lambda}_p^2 - \sqrt{P_I(N_b-\bar{\lambda}_p^2)(1-2^{-\frac{B}{N}})((P_I+d)N_b - d\bar{\lambda}_p^2 2^{\frac{B}{N}})}}{P_I\bar{\lambda}_p^2(1-2^{\frac{B}{N}}) - (P_I+d)N_b + d\bar{\lambda}_p^2}.$$

The integer bit allocation can then be found by rounding the solution for B_h in the problems above to the nearest integer less than or equal to B.

V. Simulation Results

For the simulation results in this section, we assume $N_a = N_h = 3$, $N_b = 2$ and $d = 1$, so that the closed-form solution can be applied. The power at Alice and Helper are fixed at 100. In Fig. 2 and Fig. 3, simulation results based on 100 channel realizations are shown for $B = 32$ and $B = 64$, respectively. The average MSE is plotted versus B_h, and our closed-form solution is indicated by the red line. In both cases, our solution provides an allocation that essentially yields the lowest possible average MSE. We see that the Helper commands a larger fraction of the feedback bits than Alice, indicating that interference leakage from the Helper is more detrimental to the average MSE at Bob than reduced beamforming gain from Alice. As the number of feedback bits is decreased, a higher fraction of the total number of bits is allocated to the Helper.

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