Optimal Pilot Power Allocation for OFDM Systems with Transmitter and Receiver IQ Imbalances

V. K. Varma Gottumukkala and Hlaing Minn
Department of Electrical Engineering
University of Texas at Dallas
Richardson, Texas 75080
Email: {vkg071000, Hlaing.Minn}@utdallas.edu

Abstract— **In this paper, we derive optimal pilot power allocation for OFDM systems suffering from in-phase and quadrature-phase (IQ) imbalances. Existing works in literature on IQ imbalances optimize for pilot spacings and pilot designs. However, in all these works, optimal power allocation between pilot and data symbols has not been considered. Using a lower bound on the average channel capacity as a metric, we optimize for the pilot and data power allocations. Simulations show that the resulting optimal pilot power allocation increases the channel capacity along with lowering the bit error rate (BER). We further show that the power allocation is flexible in the sense that several power allocation choices exist that improve capacity compared to the equal power allocation scenario.**

I. INTRODUCTION

Orthogonal frequency division multiplexing (OFDM) systems are widely adopted in many current and future wireless systems (e.g. IEEE 802.11a/g/n, 802.16a/e, LTE) [1]–[6]. The main advantage is that it helps cope with the frequency selectivity of the channel without the need for complex equalization and it provides better resource granularity and adaptability. However, due to analog imperfections at the transmitter and receiver, there exist mismatch in amplitude and phase between the in-phase and quadrature-phase (IQ) carriers.

The effect of IQ imbalances has been dealt in detail in literature. The effect of the IQ imbalance on the OFDM system is that it creates a mirror channel that sees the conjugate of the input symbol as its input in addition to the direct channel. This poses a new challenge in designing the receiver. Several compensation algorithms exist (e.g. [7]–[10]) that compensate for the IQ imbalances if the IQ parameters are known. Pilots are typically used to learn these parameters along with the actual channel taps before compensation can be done. Several pilot designs have been proposed for estimating I/Q imbalance or combined effect of I/Q imbalance and channel [7], [11]–[18]. In [18] pilot designs were proposed to efficiently estimate these IQ imbalance parameters by turning the problem of estimating the channel taps along with the IQ imbalance parameters into a problem of estimating two parallel channels. However, in all of these pilot designs, the question of *what is the optimal pilot power allocation in the presence of IQ imbalance* hasn't been dealt with. In [19], the problem of optimal power allocation in OFDM systems where the channel is learnt through training was considered. However, it did not consider the effect of IQ imbalance in the OFDM system. In this work, we consider a more general case of [19] where the OFDM system suffers from IQ imbalances and the IQ imbalance parameters are estimated in the form of the channel taps for both direct and mirror channels. When there are no amplitude and phase mismatches between the I and Q carriers, the mirror channel becomes zero, and the problem reduces to the problem as considered in [19]. Simulations show the improved capacity and reduced bit error rate (BER) as a result of using the optimal power allocation that we derived.

We focus on the impact of I/Q imbalance while assuming ideal conditions for other impairments such as carrier frequency offset (CFO). In practical systems, the mobile station can perform frequency synchronization from the downlink received signal, after which the residual CFO for the uplink transmission would be quite small. In this scenario, neglecting the CFO as is considered in this paper is relevant. Similarly, for the downlink reception, after proper synchronization, the effects of synchronization errors would be quite negligible; and hence the following data transmission phase would fit our considered system setup well. In brief, with proper synchronization as always required in practice, the proposed approach can be applied in practical systems.

The rest of the paper is organized as follows: Section II discusses the background material and the system model to formulate our problem. Section III presents the capacity lower bound as applied to our system model. In Section IV, we optimize this capacity lower bound and derive the optimal power allocation. Simulations and results are discussed in Section V and we conclude in Section VI.

The following notations are used throughout this paper: (i) bold font lower case denotes vector, (ii) bold font capital case denotes matrix, (iii) regular font denotes scalar, (iv) $(\cdot)^T$ denotes transpose, (v) $(\cdot)^*$ denotes complex conjugate, (vi) $(\cdot)^H$ denotes conjugate transpose, (vii) $\boldsymbol{I}_N$ denotes the identity matrix of size N, (viii) $\mathrm{Tr}(\boldsymbol{X})$ denotes trace of the matrix $\boldsymbol{X}$, (ix) $||\boldsymbol{x}||$ denotes norm of vector $\boldsymbol{x}$.

II. BACKGROUND AND SYSTEM MODEL

First we present the system model that captures the effects of IQ imbalance. Consider a single antenna system using

978-1-4244-4148-8/09/$25.00 ©2009
This full text paper was peer reviewed at the direction of IEEE Communications Society subject matter experts for publication in the IEEE "GLOBECOM" 2009 proceedings.

OFDM modulation. The frequency-independent IQ imbalance gains and phase offsets are denoted by $\{a_t^I, a_t^Q\}$ and $\{\theta_t^I, \theta_t^Q\}$. The equivalent pulse-shaping filters (i.e. overall impulse responses including D/A converters, amplifier, pulse shaping and frequency-independent imbalances) for the I and Q branches of the transmitter are denoted by $g_t^I(t)$ and $g_t^Q(t)$. The corresponding receiver side parameters are denoted by $a_r^I, a_r^Q, \theta_r^I, \theta_r^Q, g_r^I(t)$ and $g_r^Q(t)$. The low-pass equivalent channel is $h(t)$. The transmit system and receive system with IQ imbalance can each be viewed as the summation of two systems namely, the direct system and a mirror system. The impulse responses of the direct and mirror systems at the transmitter are denoted by $g_T^D(t), g_T^M(t)$ and those at receiver are denoted by $g_R^D(t), g_R^M(t)$. These are related to the IQ imbalance parameters by:

$$g_T^D(t) = \frac{1}{2}[a_t^I e^{j\theta_t^I} g_t^I(t) + a_t^Q e^{j\theta_t^Q} g_t^Q(t)] \quad (1)$$

$$g_T^M(t) = \frac{1}{2}[a_t^I e^{j\theta_t^I} g_t^I(t) - a_t^Q e^{j\theta_t^Q} g_t^Q(t)] \quad (2)$$

$$g_R^D(t) = \frac{1}{2}[a_r^I e^{-j\theta_r^I} g_r^I(t) + a_r^Q e^{-j\theta_r^Q} g_r^Q(t)] \quad (3)$$

$$g_R^M(t) = \frac{1}{2}[a_r^I e^{j\theta_r^I} g_r^I(t) - a_r^Q e^{j\theta_r^Q} g_r^Q(t)]. \quad (4)$$

The overall system can be expressed as the sum of direct and mirror channels $p(t)$ and $q(t)$ respectively, as shown in Fig. 1, and they are given by:

$$p(t) = g_T^D(t) * h(t) * g_R^D(t) + (g_T^M(t))^* * h^*(t) * g_R^M(t) \quad (5)$$

$$q(t) = g_T^M(t) * h(t) * g_R^D(t) + (g_T^D(t))^* * h^*(t) * g_R^M(t). \quad (6)$$

Now, consider a discrete-time OFDM system with N subcarriers. The discrete-time versions of the channels $p(t)$ and $q(t)$ are denoted by $\boldsymbol{p}$ and $\boldsymbol{q}$ respectively and they consist of maximum of L taps each. If the direct channel sees an input $s(t)$, the mirror channel sees an input $s^*(t)$. The time-domain training signal is denoted by $s(k), k \in [-N_{\text{CP}}, N-1]$ where N_{CP} is the number of cyclic prefix samples. The time-domain data signals are denoted by $x(k), k \in [-N_{\text{CP}}, N-1]$. The time-domain $N \times 1$ received signal vector $\boldsymbol{r}$ for one OFDM symbol after cyclic prefix removal is given by:

$$\boldsymbol{r} = (\boldsymbol{S} + \boldsymbol{X})\boldsymbol{p} + (\boldsymbol{S}^* + \boldsymbol{X}^*)\boldsymbol{q} + \boldsymbol{n} \quad (7)$$

where $\boldsymbol{S}$ denotes the training signal convolution matrix (circulant) of size $N \times L$. The elements of $\boldsymbol{S}$ are given by $S(m,k) = s(m-k), m \in [0, N-1], k \in [0, L-1]$. The data matrix $\boldsymbol{X}$ of size $N \times L$ is defined similar to $\boldsymbol{S}$ with the elements given by $X(m,k) = x(m-k), m \in [0, N-1], k \in [0, L-1]$. The complex Gaussian noise vector $\boldsymbol{n}$ is given by $\boldsymbol{n} = g_R^D \boldsymbol{w} + g_R^M \boldsymbol{w}^*$ where $\boldsymbol{w}$ consists of independent and identically distributed (iid) circularly-symmetric complex Gaussian random variables with variance σ_w^2. Thus, $\boldsymbol{n}$ consists of complex Gaussian random variables with variance denoted by σ_n^2.

We consider a pilot design in [18] called frequency domain nulling (FDN). In particular, non-zero pilots are placed in equally spaced L tones while zero-pilots are placed in the L

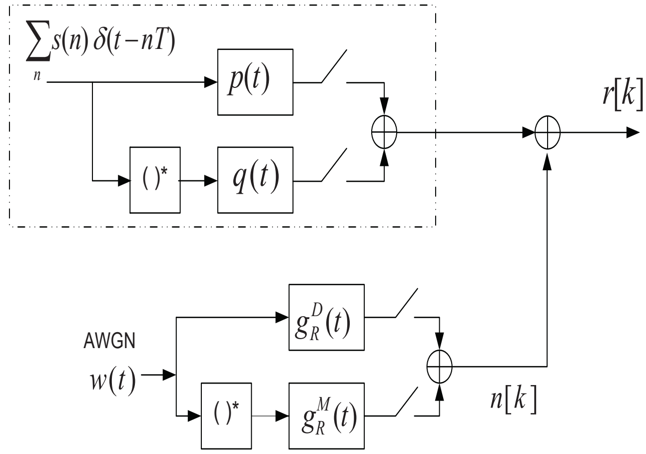

Fig. 1. System Model

mirror tones corresponding to the L tones which contain the non-zero pilots. The data signals are placed in the remaining $N-2L$ tones. This arrangement provides for interference avoidance between data and pilots in the frequency domain. The channel taps h_l are assumed to be complex Gaussian distributed with zero-mean and variance $\sigma_{h_l}^2$, and they are independent of each other.

Frequency-independent IQ imbalance is mainly caused by imperfection in the mixers and hence is prevalent in all direct-conversion radios. Frequency-dependent IQ imbalance is typically caused by imperfection in filters/amplifiers in very-wideband systems, and its frequency selectivity is typically quite mild. Hence, apart from low-cost very-wideband radios, IQ imbalance in typical systems can be well modelled as frequency-independent. We therefore consider frequency-independent IQ imbalance in this work. This causes the IQ imbalance pulse shaping filters to have single taps each. The discrete-time versions of direct and mirror channels can then be written as

$$\boldsymbol{p} = a_1\boldsymbol{h} + a_2\boldsymbol{h}^* \quad (8)$$

$$\boldsymbol{q} = b_1\boldsymbol{h} + b_2\boldsymbol{h}^* \quad (9)$$

where

$$a_1 = g_T^D g_R^D \quad (10)$$

$$a_2 = (g_T^M)^* g_R^M \quad (11)$$

$$b_1 = g_T^M g_R^D \quad (12)$$

$$b_2 = (g_T^D)^* g_R^M \quad (13)$$

are fixed constants. This makes the channels $\boldsymbol{p}$ and $\boldsymbol{q}$ iid complex Gaussian when $\boldsymbol{h}$ is iid complex Gaussian. Thus, we have $p_l \sim \mathcal{CN}(0, \sigma_{p_l}^2)$ and $q_l \sim \mathcal{CN}(0, \sigma_{q_l}^2)$, where $\mathcal{CN}$ denotes complex Gaussian and

$$\sigma_{p_l}^2 = (a_1 a_1^* + a_2 a_2^*)\sigma_{h_l}^2 \quad (14)$$

$$\sigma_{q_l}^2 = (b_1 b_1^* + b_2 b_2^*)\sigma_{h_l}^2. \quad (15)$$

We consider least-squares estimation of channels $\boldsymbol{p}$ and $\boldsymbol{q}$

as follows:

$$\hat{\boldsymbol{p}} = (\boldsymbol{S}^H\boldsymbol{S})^{-1}\boldsymbol{S}^H\boldsymbol{r} \quad (16)$$

$$\hat{\boldsymbol{q}} = (\boldsymbol{S}^T\boldsymbol{S}^*)^{-1}\boldsymbol{S}^T\boldsymbol{r}. \quad (17)$$

Then, their MSEs are given as,

$$\mathrm{MSE}_{\boldsymbol{p}} := \sigma^2_{\Delta\boldsymbol{p}} = \sigma_n^2\mathrm{Tr}[(\boldsymbol{S}^H\boldsymbol{S})^{-1}] \quad (18)$$

$$\mathrm{MSE}_{\boldsymbol{q}} := \sigma^2_{\Delta\boldsymbol{q}} = \sigma_n^2\mathrm{Tr}[(\boldsymbol{S}^H\boldsymbol{S})^{-1}]. \quad (19)$$

For the pilot design considered in this paper, the above can be simplified further. We have $\boldsymbol{S}^H\boldsymbol{S} = P_s\boldsymbol{I}_L$, where $P_s = ||\boldsymbol{s}||^2$ is the total power in the training signal. Therefore, we get,

$$\sigma^2_{\Delta\boldsymbol{p}} = \sigma_n^2 L/P_s \quad (20)$$

$$\sigma^2_{\Delta\boldsymbol{q}} = \sigma_n^2 L/P_s. \quad (21)$$

The frequency domain taps can then be obtained as follows:

$$[\hat{P}_0, \hat{P}_1, ..., \hat{P}_{N-1}]^T = \sqrt{N}F_L\hat{\boldsymbol{p}} \quad (22)$$

and

$$[\hat{Q}_0, \hat{Q}_1, ..., \hat{Q}_{N-1}]^T = \sqrt{N}F_L\hat{\boldsymbol{q}} \quad (23)$$

where $\mathbf{F}_L$ consists of the first L columns of the unitary $N\times N$ DFT matrix $\mathbf{F}$ whose $(m,n)\mathrm{th}$ element is given by $F(m,n) = \frac{1}{\sqrt{N}}e^{-\frac{j2\pi mn}{N}}$, $m \in [0, N-1]$ and $n \in [0, N-1]$. We also have

$$[P_0, P_1, ..., P_{N-1}]^T = \sqrt{N}F_L\boldsymbol{p} \quad (24)$$

and

$$[Q_0, Q_1, ..., Q_{N-1}]^T = \sqrt{N}F_L\boldsymbol{q}. \quad (25)$$

Note that P_i for any $i \in [0, N-1]$ is $\mathcal{CN}(0, \sigma_P^2)$ where $\sigma_P^2 := \sum_{l=0}^{L-1}\sigma^2_{p_l}$. Similarly, Q_i for any $i \in [0, N-1]$ is $\mathcal{CN}(0, \sigma_Q^2)$ where $\sigma_Q^2 := \sum_{l=0}^{L-1}\sigma^2_{q_l}$. Let $\mathcal{I}$ denote the index set consisting of the M data tones and $\mathcal{J}$ denote the index set consisting of the K pilot tones. Let $\bar{N} = N + N_{\mathrm{CP}}$ be the total number of samples in one OFDM symbol including the cyclic prefix. In the following, we consider cyclic prefix of length L, i.e. $N_{\mathrm{CP}} = L$.

III. Capacity Lower Bound

To enhance the capacity of our system, we find the optimal pilot power allocation using a lower bound on the channel capacity. Let the total transmit power be $P = P_x + P_s$, where $P_x = E\{||\mathbf{x}||^2\}$ and $P_s = ||\mathbf{s}||^2$. Under the total power constraint, and following standard steps (as in [20]–[23]), we have the channel capacity (normalized per transmit symbol) averaged over the channel matrix as,

$$C_{\mathrm{ideal}} = \frac{M}{\bar{N}}E[\log(1+\rho_{\mathrm{ideal}}|g|^2)] \quad (26)$$

where

$$\rho_{\mathrm{ideal}} := \frac{(\sigma_P^2+\sigma_Q^2)P_x}{M\sigma_n^2}. \quad (27)$$

Incorporating the channel estimation error, we apply a lower bound (as found in [20], [24]), and for joint detection of X_i and X^*_{-i}, after summing across subcarriers, we obtain for the case of frequency-independent IQ imbalances,

$$C \geq \underline{\mathrm{C}} := \quad (28)$$

$$\frac{1}{\bar{N}}\sum_{i\in\mathcal{I}}E\left\{\log\left(1+\frac{|\hat{P}_iX_i+\hat{Q}_iX^*_{-i}|^2}{E\{|\Delta P_iX_i+\Delta Q_iX^*_{-i}|^2\}+M\sigma_n^2}\right)\right\} \quad (29)$$

We then define:

$$g = \frac{\hat{P}_iX_i+\hat{Q}_iX^*_{-i}}{\sqrt{E\{|\hat{P}_iX_i+\hat{Q}_iX^*_{-i}|^2\}}}. \quad (30)$$

Substituting the above g in (29), we obtain

$$\underline{\mathrm{C}} = \frac{1}{\bar{N}}\sum_{i\in\mathcal{I}}E\left\{\log\left(1+|g|^2\frac{\sigma_P^2+\sigma^2_{\Delta\boldsymbol{p}}+\sigma_Q^2+\sigma^2_{\Delta\boldsymbol{q}}}{\sigma^2_{\Delta\boldsymbol{p}}+\sigma^2_{\Delta\boldsymbol{q}}+M\sigma_n^2/P_x}\right)\right\} \quad (31)$$

where we have used

$$E\{|\hat{P}_i|^2\} = \sigma_P^2 + E\{|\Delta P_i|^2\} \quad (32)$$

and

$$E\{|\hat{Q}_i|^2\} = \sigma_Q^2 + E\{|\Delta Q_i|^2\}. \quad (33)$$

We can verify that [19]

$$E\{|\Delta P_i|^2\} = \sigma^2_{\Delta\boldsymbol{p}} \quad \forall i \in \mathcal{I} \quad (34)$$

and

$$E\{|\Delta Q_i|^2\} = \sigma^2_{\Delta\boldsymbol{q}} \quad \forall i \in \mathcal{I}. \quad (35)$$

Using the above, we can write (31) as

$$\underline{\mathrm{C}} = \frac{M}{\bar{N}}E\{\log(1+\rho|g|^2)\} \quad (36)$$

where

$$\rho = \frac{\sigma_P^2+\sigma^2_{\Delta\boldsymbol{p}}+\sigma_Q^2+\sigma^2_{\Delta\boldsymbol{q}}}{\sigma^2_{\Delta\boldsymbol{p}}+\sigma^2_{\Delta\boldsymbol{q}}+\frac{M\sigma_n^2}{P_x}}. \quad (37)$$

IV. Optimal Pilot Power Allocation

We define the ratio of the data power to total power to be α with $0 < \alpha < 1$. Then, $P_x = \alpha P$ and $P_s = (1-\alpha)P$ where P is the total power in one OFDM symbol. From (36), we can see that once M and $\bar{N} = N + L$ are fixed, the capacity lower bound becomes a function of ρ only. Since $\log(\cdot)$ is an increasing function, for fixed M and K, $\underline{\mathrm{C}}$ is maximized when ρ is maximized and hence the optimal α can be found out by maximizing the expression for ρ. The optimal α can be found numerically by finding the value of α that maximizes (37) over a discrete set of values. For fixed M and $K = N - M$, a closed-form solution for α can be obtained for high SNR regime. At high SNR, channel estimation error is very small, hence the error variance terms $\sigma^2_{\Delta\boldsymbol{p}}$ and $\sigma^2_{\Delta\boldsymbol{q}}$ approach zero and we have $\sigma_P^2+\sigma^2_{\Delta\boldsymbol{p}} \cong \sigma_P^2$ and $\sigma_Q^2+\sigma^2_{\Delta\boldsymbol{q}} \cong \sigma_Q^2$. Also, $\sigma^2_{\Delta\boldsymbol{p}} = \sigma_n^2\mathrm{Tr}[(S^HS)^{-1}] = \sigma_n^2L/P_s$ and $\sigma^2_{\Delta\boldsymbol{q}} = \sigma_n^2\mathrm{Tr}[(S^HS)^{-1}] = \sigma_n^2L/P_s$. Substituting these approximations and $P_x = \alpha P$ and $P_s = (1-\alpha)P$ in (37), we obtain

$$\rho = \frac{\sigma_P^2+\sigma_Q^2}{\sigma_n^2L/P_s+\sigma_n^2L/P_s+M\sigma_n^2/P_x} \quad (38)$$

i.e.,

$$\begin{aligned}\rho &= \frac{\sigma_P^2+\sigma_Q^2}{2\sigma_n^2 L/P_s + M\sigma_n^2/P_x} \\ &= \frac{\sigma_P^2+\sigma_Q^2}{\sigma_n^2}\frac{1}{2L/P_s+M/P_x} \\ &= \frac{\sigma_P^2+\sigma_Q^2}{\sigma_n^2}\frac{1}{\frac{2L}{(1-\alpha)P}+\frac{M}{\alpha P}} \\ &= \frac{P(\sigma_P^2+\sigma_Q^2)}{\sigma_n^2}\frac{1}{\frac{2L}{(1-\alpha)}+\frac{M}{\alpha}} \\ &= M\frac{P(\sigma_P^2+\sigma_Q^2)}{M\sigma_n^2}\frac{\alpha(1-\alpha)}{2L\alpha+M(1-\alpha)} \\ &= M\rho_{\text{snr}}\frac{\alpha(1-\alpha)}{2L\alpha+M(1-\alpha)} \qquad (39)\end{aligned}$$

where

$$\rho_{\text{snr}} = \frac{P(\sigma_P^2+\sigma_Q^2)}{M\sigma_n^2} \qquad (40)$$

is the output SNR. The above value of ρ is maximized by differentiating it with respect to α, setting it equal to zero and solving the resulting equation. The optimal α for high SNR regime is then given by

$$\alpha_{\text{opt}} = \frac{1}{1+\sqrt{\frac{2L}{M}}}. \qquad (41)$$

Note that this solution resembles the solution found in [19] but with a channel length of $2L$ instead of the length L due to the mirror channel introduced by the IQ imbalance.

We can see that neither the transmitter nor the receiver need to know apriori the IQ imbalance parameters. The receiver learns the IQ parameters by using pilot symbols. The optimal power allocation at high SNR depends only on the number of pilot tones and the number of data tones. Thus the result of our power allocation does not require knowledge of IQ imbalance parameters. The benefits of the derived results are reflected in the improvement in capacity in Figs. 2 and 4 and BER in Fig. 3.

V. Simulations and Results

We use simulations to compare the performance of our system under the optimal power allocation derived above and equal power allocation of $\alpha_{\text{equal}} = \frac{1}{1+(\frac{2L}{M})}$. For simulation, we use $N = 64$ subcarriers and we consider $L = 8$ time domain channel taps for both $\boldsymbol{p}$ and $\boldsymbol{q}$ channels. We use $K = 2L = 16$ pilot tones and $M = N - 2L = 48$ data tones within one OFDM symbol. Fig. 2 shows comparison of capacity for four different cases: 1) when channel state is perfectly known and no training is required, 2) optimal power allocation with imperfect channel estimation, 3) optimal power allocation with perfect channel estimation, 4) equal power allocation where $\alpha = \alpha_{\text{equal}}$. From the results, we can see that optimal power allocation improves the channel capacity compared to equal power allocation case, while there is a significant loss due to imperfect channel estimation and also compared to the case when channel state information (CSI) is known.

Fig. 3 shows the BER performance when the optimal power allocation is done. We can see that there is slight performance improvement compared to equal power allocation case. And significant difference exists compared to perfect channel estimation case.

Fig. 4 shows the variation of capacity as a function of α. As expected, the maximum is seen at $\alpha = 0.63$ compared to the calculated value of $\alpha_{\text{opt}} = \frac{1}{1+\sqrt{\frac{2L}{M}}} = 0.634$. Also the relatively flat peak of the capacity curve shows that several choices exist that give similar capacity performance. This also validates the results seen in Fig. 2 and Fig. 3 that there is only a marginal improvement in capacity and BER from the equal power case ($\alpha_{\text{equal}} = \frac{1}{1+\frac{2L}{M}} = 0.75$) which lies close to the flat peak of the capacity curve. However, when the pilot power is significantly boosted, for e.g. $\alpha = 0.2$, a degradation of about 21% decrease in capacity is observed at $\text{SNR} = 20\text{dB}$.

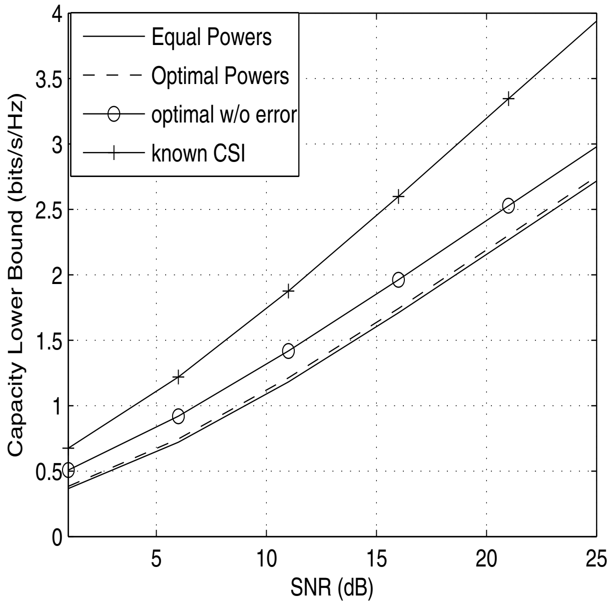

Fig. 2. Capacity Lower Bound versus SNR

VI. Conclusions

In this paper, we considered an OFDM system suffering from frequency-independent in-phase and quadrature-phase imbalances and derived an optimal power allocation between pilots and data symbols for a fixed total transmit power constraint. Using a lower bound on the average channel capacity as a metric, we optimized for the power allocation between pilot and data symbols. We obtained a closed form solution for high SNR scenario. Simulation results show that the resulting power allocation leads to increased channel capacity and also reduced BER compared to equal power allocation. Also, the slight increase in performance of the optimal power allocation

978-1-4244-4148-8/09/$25.00 ©2009
This full text paper was peer reviewed at the direction of IEEE Communications Society subject matter experts for publication in the IEEE "GLOBECOM" 2009 proceedings.

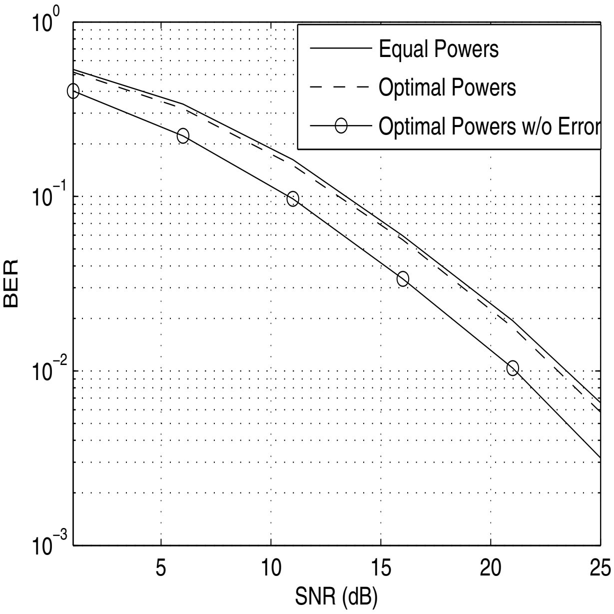

Fig. 3. BER versus SNR

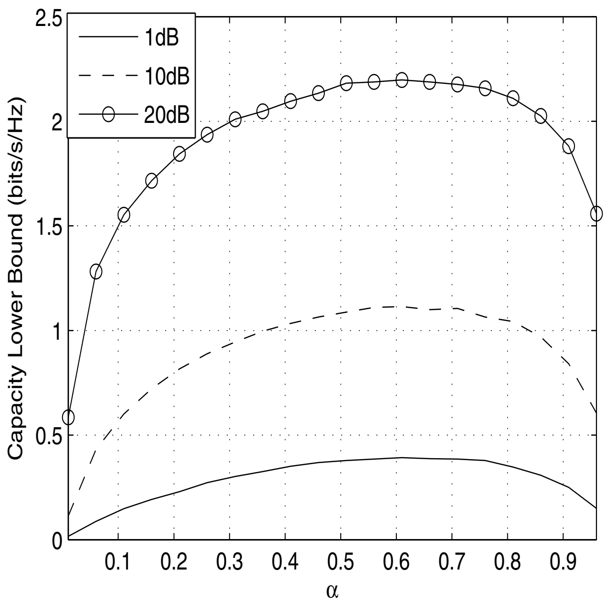

Fig. 4. Capacity Lower Bound versus $\alpha = \frac{P_x}{P}$

compared to the equal power allocation shows that equal power allocation is close to optimal power allocation.

References

[1] *Wireless LAN Medium Access Control (MAC) and Physical Layer (PHY) Specifications: High-Speed Physical Layer in the 5 GHz Band*, IEEE Standard 802.11a, 1999.

[2] *Broadband Wireless Access: IEEE MAN standard*, IEEE LAN/MAN Standards Committee IEEE 802.16a, 2003.

[3] *Broadband Wireless Access: IEEE MAN Standard*, IEEE LAN/MAN standards committee IEEE 802.16e, 2005.

[4] *Mobile Broadband Wireless Access: IEEE MAN standard*, IEEE LAN/MAN Standards Committee IEEE 802.20 Std.

[5] *3GPP, Long-Term Evolution (LTE), "Physical Channels and Modulation" (Release 8)*, 3GPP TS 36.211 v1.2.0, June 2007.

[6] *3GPP2, "Physical Layer form Ultra Mobile Broadband (UMB) Air Interface Specification,"*, 3GPP2 C.S0084-001-0 v3.0, August 2008.

[7] A. Tarighat, R. Bagheri, and A. H. Sayed, "Compensation schemes and performance analysis of IQ imbalances in OFDM receivers," *IEEE Trans. Signal Processing*, vol. 53, no. 8, pp. 3257–3268, Aug. 2005.

[8] M. Valkama, M. Renfors, and V. Koivunen, "Compensation of frequency-selective I/Q imbalances in wideband receivers: models and algorithms," in *Proc. IEEE Third Workshop on Signal Processing Advances in Wireless Communications (SPAWC '01)*, 20–23 March 2001, pp. 42–45.

[9] J. Tubbax, A. Fort, L. Van der Perre, S. Donnay, M. Engels, M. Moonen, and H. De Man, "Joint compensation of IQ imbalance and frequency offset in OFDM systems," in *IEEE Global Telecommunications Conference, 2003. GLOBECOM'03*, vol. 4, 2003.

[10] D. Tandur and M. Moonen, "Joint Adaptive Compensation of Transmitter and Receiver IQ Imbalance Under Carrier Frequency Offset in OFDM-Based Systems," *IEEE Trans. Signal Processing*, vol. 55, no. 11, pp. 5246–5252, Nov. 2007.

[11] W. Kirkland and K. Teo, "I/Q distortion correction for OFDM direct conversion receiver," *Electronics Letters*, vol. 39, no. 1, pp. 131–133, 2003.

[12] L. Brotje, S. Vogeler, K. Kammeyer, R. Rueckriem, and S. Fechtel, "Estimation and Correction of transmitter-caused I/Q Imbalance in OFDM Systems," in *Proc. 7th International OFDM Workshop*, 2002, pp. 178–182.

[13] Y. Egashira, Y. Tanabe, and K. Sato, "A Novel IQ Imbalance Compensation Method with Pilot-Signals for OFDM System," 2008.

[14] R. Chrabieh and S. Soliman, "IQ Imbalance Mitigation via Unbiased Training Sequences," in *IEEE Global Telecommunications Conference, 2007. GLOBECOM'07*, 2007, pp. 4280–4285.

[15] E. Lopez-Estraviz, S. De Rore, F. Horlin, and L. Van der Perre, "Optimal training sequences for joint channel and frequency-dependent IQ imbalance estimation in OFDM-based receivers," in *IEEE International Conference on Communications, 2006. ICC'06*, vol. 10, 2006.

[16] E. Lopez-Estraviz, S. De Rore, F. Horlin, and A. Bourdoux, "Pilot design for Joint Channel and Frequency-Dependent Transmit/Receive IQ Imbalance Estimation and Compensation in OFDM-Based Transceivers," in *IEEE International Conference on Communications, 2007. ICC'07*, 2007, pp. 4861–4866.

[17] T. Schenk, P. Smulders, and E. Fledderus, "Estimation and compensation of TX and RX IQ imbalance in OFDM based MIMO systems," in *Proc. IEEE Radio and Wireless Symposium (RWS 2006*, 2006, pp. 215–218.

[18] H. Minn and D. Munoz, "Pilot designs for channel estimation of OFDM systems with frequency dependent I/Q imbalances," *accepted in IEEE WCNC*, Apr. 2009.

[19] S. Ohno and G. B. Giannakis, "Capacity maximizing MMSE-optimal pilots for wireless OFDM over frequency-selective block Rayleigh-fading channels," *IEEE Trans. Inform. Theory*, vol. 50, no. 9, pp. 2138–2145, Sept. 2004.

[20] E. Biglieri, J. Proakis, and S. Shamai, "Fading channels: information-theoretic and communications aspects," *IEEE Trans. Inform. Theory*, vol. 44, no. 6, pp. 2619–2692, Oct. 1998.

[21] T. Cover, J. Thomas, J. Wiley, and W. InterScience, *Elements of information theory*. Wiley New York, 1991.

[22] T. L. Marzetta and B. M. Hochwald, "Capacity of a mobile multiple-antenna communication link in Rayleigh flat fading," *IEEE Trans. Inform. Theory*, vol. 45, no. 1, pp. 139–157, Jan. 1999.

[23] I. Telatar, "Capacity of multi-antenna Gaussian channels," *European transactions on telecommunications*, 1999.

[24] M. Medard, "The effect upon channel capacity in wireless communications of perfect and imperfect knowledge of the channel," *IEEE Trans. Inform. Theory*, vol. 46, no. 3, pp. 933–946, May 2000.