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**Addressing IQ Mismatch
in Spatial Interference Suppression Systems**

A thesis submitted in partial satisfaction
of the requirements for the degree
Master of Science in Electrical Engineering

by

Gaelen Stanislaus Pereira

2012

ABSTRACT OF THE THESIS

Addressing IQ Mismatch in Spatial Interference Suppression Systems

by

Gaelen Stanislaus Pereira

Master of Science in Electrical Engineering

University of California, Los Angeles, 2012

Professor Babak Daneshrad, Chair

Second order statistic based Spatial Interference Suppression has been shown effective in suppressing strong interference from jammers or concurrent transmissions. Interference suppression abilities are especially useful to commercial systems that need to support an ever increasing number of wireless devices in the shared wireless medium, and military systems that need to communicate in the presence of strong jammers. However, implementation impairments in the RF frontend can be a major impediment to the design of such systems since the interference can be several orders of magnitude stronger than the signal.

The main contribution of this thesis is to analyze the impact of implementation impairments, in particular IQ mismatch, and to compensate for it in interference suppression systems. Investigating the factors that limit the interference suppression ability of such systems is important, as this might lead to answers about the bounds on jammer suppression or concurrent links supportable in practical communication systems. Moreover, it is important to mitigate these limits on interference suppression as we push for denser networks and more jammer resistance. This work presents architectures for IQ mismatch compensation in both narrowband and wideband interference suppression systems.

Chapter 1 presents an overview of Spatial Interference Suppression algorithms. The mechanism behind interference suppression based on second order statistics of the interference is

illustrated, and the use of spatial interference suppression as frontend filters for packet based communication systems is also discussed.

Chapter 2 discusses implementation impairments in spatial interference suppression systems. It is shown that IQ mismatch is a key limiter to the interference suppression ability of spatial interference suppression algorithms that are based on the second order statistics of the interference.

Chapter 3 presents the expanded sub-space concept, and shows how IQ mismatch in spatial interference suppression systems can be compensated for in the expanded sub-space. An interference suppression architecture for narrowband systems with flat fading channels is introduced, and simulation results are presented that show that the effect of IQ mismatch can be mitigated.

Chapter 4 extends the interference suppression architecture with IQ mismatch compensation to wideband systems with frequency selective channels. An architecture based on sub-banding is proposed that utilizes the existing blocks from wideband interference suppression systems. Simulation results in scenarios typical of indoor environments are presented to indicate the expected gains in a practical system.

Chapter 5 concludes the thesis, with a brief discussion on how the expanded sub-space concept can be used to mitigate the frequency selectivity of the channel.

Notation: Vectors are represented by lowercase bold letters, while matrices are represented by uppercase bold letters. We use $(.)^T$ to denote the transpose, $(.)^*$ to denote the conjugate and $(.)^H$ to denote the conjugate transpose of a matrix.

The thesis of Gaelen Stanislaus Pereira is approved.

Gregory J. Pottie

Danijela Cabric

Babak Daneshrad, Committee Chair

University of California, Los Angeles

2012

To my father...

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CHAPTER 1

Spatial Interference Suppression: An Overview

Communication system design today is not just about maximizing the spectral efficiency of a link. As the number of wireless devices contending for the shared wireless medium continues to rise sharply, it is imperative to have means that allow concurrent usage of the available spectrum. Central to this objective is the development of methods that allow suppression of interference caused by other transmissions in a network thus enabling concurrent links. Interference suppression capabilities are also essential to overcome the presence of strong jammers in the environment, a scenario typical of military communication systems. This chapter presents an overview of spatial interference suppression systems. Section 1.1 introduces the idea of using additional degrees of freedom by virtue of multiple antennas at the receiver for spatial interference suppression, Section 1.2 presents spatial interference suppression techniques that are based on second order statistics of the interference and Section 1.3 covers the use of these techniques as frontend filters in packet based communication systems.

1.1 Multiple Antennas and Additional Degrees of Freedom

Multiple antennas provide communication system designers with additional degrees of freedom. These additional degrees of freedom have been traditionally used to increase the capacity or the reliability of the wireless link [1]. The former is referred to as Spatial Multiplexing [2] - [4] wherein parallel independent streams of data can be sent over the multiple antennas, and the maximum number of parallel data streams is limited by the minimum of the transmit and receive antennas. The latter is referred to as Diversity, wherein an appropriate spatial combining of signals at the receiver or spreading of signals across space and time at the transmitter [5] can lead to a lower

error rate. Unlike spatial multiplexing, diversity can be provided by having multiple antennas at either end of the link. Both these methods can improve the performance of a wireless link without increasing the spectrum usage, and therefore lead to higher link spectral efficiency.

However, as wireless communication gets more pervasive, and the number of wireless devices increase sharply, it is not only essential to enable links with high spectral efficiency, but moreover to enable networks with higher sum spectral efficiency. This has fueled research towards alternate uses of multiple antennas for increasing the network's sum spectral efficiency. In fact, an approach that relies solely on increasing the link spectral efficiency can be detrimental to the network, since it might limit the number of users that can communicate concurrently. One approach to increase the number of concurrent users is to use multiple antennas at the receiver for Interference Suppression [6] - [16]. Here the objective is to support communication links in the presence of interference arising from other concurrent links in the shared wireless medium. Another approach that goes one step further is the use of multiple antennas at the transmitter for Interference Alignment [17], where users coordinate their transmissions in space and time so as to reduce the interference at the receiver.

Interference suppression was first explored in the context of antenna theory, where the main interest was in detecting radar signals in the presence of strong jammers [6] - [10]. By using phased antenna arrays, it was possible to steer the mainlobe of the antenna pattern so as to direct it towards the incoming signal while creating nulls in the direction of the jammers. While communication algorithms today are primarily developed in the digital baseband, many interference suppression techniques are still inspired by the jammer suppression research from the antenna community.

The idea of Optimum Combining, which uses the spatial covariance of the interference and noise to combine the signals from an antenna array was presented in [6]. It was shown that optimum combining maximizes the signal to interference plus noise ratio in the presence of interference, and can be better than Maximum Ratio Combining. [7] indicated the potential benefits of smart and adaptive antenna arrays to wireless systems. For a system with M antennas in the presence of N interferers, all interferers can theoretically be canceled if $N < M$. Moreover,

when $N < M$, the additional $M - N$ eigen modes can be used for diversity. [8] discussed the convergence of adaptive antenna array algorithms, and concluded that algorithms based on inversion of the spatial covariance of the interference plus noise have a rapid convergence rate. [9] extended the idea of spatial processing of signals from antenna arrays to space - time processing. Broadband interference leads to wider effective angles in the spatial covariance, due to the equivalence of frequency and direction as perceived by a narrowband antenna array. It was shown that this effect can be mitigated by processing the received signal over both space and time.

All these techniques assumed a signal with only one spatial stream. Therefore the antenna combining is merely a weighted sum in order to yield a single signal. However, with the advent of MIMO Spatial Multiplexing and the ability to transmit parallel data streams, there has been renewed interest into how to effectively use the degrees of freedom for spatial multiplexing, diversity and interference suppression. This issue is dealt with in the context of MIMO Multi-user detection, where the aim is to be able to decode independent streams of data while suppressing interference from other transmissions in the network. Multi-user MIMO detection using an iterative interference cancellation approach is presented in [11], where a signal stream is first decoded and then used to cancel out the interference from itself to other signal streams in future iterations. [12] discusses performance bounds of different MIMO receivers in the presence of interference. It is found that the MMSE receiver performs quite close to the optimal multi-user detector in the presence of interference. [13] considers the viability of antenna array processing for interference suppression in MIMO communication systems. Moreover, experimental results are presented in the presence of frequency and timing offsets. Of late, there has been work towards better modeling of RF interference [14], which consists of the co-channel interference discussed above and in addition, interference from sources like external electronics. [14] also shows that the performance of receivers can be improved if they are specifically designed for these environments, as compared to those receivers designed assuming a gaussian channel.

Most linear receivers depend in some form on the second order statistics of the received signal. Interference suppression based on second order statistics is discussed in more detail below.

Figure 1.1: Interference suppression based on second order statistics

1.2 Spatial Interference Suppression using Second Order Statistics

Spatial interference suppression is based on the observation that correlated copies of the interference signal are received on each antenna. This correlation can be exploited to suppress the interference below the level of the thermal noise, which can be assumed independent on each antenna. Consider the system in Fig. 1.1. The received signal vector $\mathbf{y}$ is a linear combination of the signal vector $\mathbf{x}_s$, the interference signal vector $\mathbf{x}_i$ and the white gaussian noise vector $\mathbf{v}$. Assuming quasi-static rayleigh flat fading channels $\mathbf{H}_s$ and $\mathbf{H}_i$,

$$\mathbf{y} = \mathbf{H}_s \mathbf{x}_s + \mathbf{H}_i \mathbf{x}_i + \mathbf{v} \quad (1.1)$$

Interference suppression involves the use of a filter $\mathbf{W}$ that linearly operates on the received signal $\mathbf{y}$ to produce a clean signal $\mathbf{p}$.

$$\begin{aligned} \mathbf{p} &= \mathbf{W} \mathbf{y} \\ &= \mathbf{W} (\mathbf{H}_s \mathbf{x}_s + \mathbf{H}_i \mathbf{x}_i + \mathbf{v}) \end{aligned} \quad (1.2)$$

If the objective is to decode $\mathbf{x}_s$ directly using $\mathbf{p}$, then the optimal filter is known to be the MMSE solution,

$$\mathbf{W}_{\text{MMSE}} = \mathbf{H}_s^H (\mathbf{H}_s \mathbf{R}_s \mathbf{H}_s^H + \mathbf{H}_i \mathbf{R}_i \mathbf{H}_i^H + \mathbf{R}_v)^{-1} \quad (1.3)$$

where $\mathbf{R}_s = \mathbf{E} [\mathbf{x}_s \mathbf{x}_s^H]$, $\mathbf{R}_i = \mathbf{E} [\mathbf{x}_i \mathbf{x}_i^H]$ and $\mathbf{R}_v = \mathbf{E} [\mathbf{v} \mathbf{v}^H]$ represent the spatial covariance of the signal, interference and noise respectively. However, this linear filter relies on channel estimates, which could be hard to obtain accurately in the presence of strong interference. Therefore, it would be nice to have a filter that depends only on the second order statistics of the received

signal since this is much easier to obtain. For example, the spatial covariance $\mathbf{R}_y$ of the received signal can be obtained by merely multiplying the received signal by its conjugate and averaging over time. Assuming the time index of the signal to be n ,

$$\mathbf{R}_y = \frac{1}{n} \sum_{i=1}^n \mathbf{y}(\mathbf{i})\mathbf{y}(\mathbf{i})^H \quad (1.4)$$

Two such filters that depend only on the second order statistics of the signal have been presented in [15], and are discussed below. For any filter $\mathbf{W}$, the Signal to Interference plus Noise Ratio (SINR) ρ at the output can be computed as follows

$$\begin{aligned} \rho &= \frac{\text{Tr}(\mathbf{W}\mathbf{H}_s\mathbf{R}_s\mathbf{H}_s^H\mathbf{W}^H)}{\text{Tr}(\mathbf{W}(\mathbf{H}_i\mathbf{R}_i\mathbf{H}_i^H + \mathbf{R}_v)\mathbf{W}^H)} \\ &= \frac{\text{Tr}(\mathbf{W}\mathbf{H}_s\mathbf{R}_s\mathbf{H}_s^H\mathbf{W}^H)}{\text{Tr}(\mathbf{W}(\mathbf{R}_\gamma + \mathbf{R}_v)\mathbf{W}^H)} \end{aligned} \quad (1.5)$$

where $\text{Tr}(\cdot)$ represents the trace operation on a matrix. If the received interference vector is represented by $\gamma = \mathbf{H}_i\mathbf{x}_i$, then $\mathbf{R}_\gamma = \mathbf{H}_i\mathbf{R}_i\mathbf{H}_i^H$.

The first filter maximizes ρ , and is given by

$$\mathbf{W}_{\text{MSINR}} = \mathbf{H}_s\mathbf{R}_s\mathbf{H}_s^H(\mathbf{R}_\gamma + \mathbf{R}_v)^{-1} \quad (1.6)$$

It is worth noting here that unlike the MMSE solution which maximizes the SINR of each spatial stream, $\mathbf{W}_{\text{MSINR}}$ maximizes the ratio of the total signal power to the interference plus noise power. This is the primary reason that $\mathbf{W}_{\text{MSINR}}$ does not require estimates of the first order statistics of the signal i.e. the channel $\mathbf{H}_s$.

The second filter is based on Sample Matrix Inversion (SMI), and has been shown effective in suppressing interference [16]. This filter depends only on the second order statistics of the interference and noise, and is given by

$$\mathbf{W}_{\text{SMI}} = (\mathbf{R}_\gamma + \mathbf{R}_v)^{-1} \quad (1.7)$$

Both these filters are effective in suppressing interference as they involve matrix inversion of the spatial covariance of the interference plus noise. As depicted in Fig. 1.1, this gives rise to nulls in the spatial domain in the direction of the interference sources. Another way to look at this

is to consider the eigen value decomposition of the spatial covariance matrix. If there are m independent interference sources whose received powers are above the noise floor σ_v^2 , then the eigen values of $\mathbf{R}_{\gamma+\mathbf{v}} = \mathbf{R}_\gamma + \mathbf{R}_\mathbf{v}$ represent the received powers of these m sources. These eigen values are given by

$$\mathbf{\Lambda}_{\gamma+\mathbf{v}} = \begin{bmatrix} \sigma_{\gamma(1)}^2 & & & \\ & \dots & & \\ & & \sigma_{\gamma(m)}^2 & \\ & & & \epsilon \end{bmatrix} + \begin{bmatrix} \sigma_v^2 & & & \\ & \dots & & \\ & & \dots & \\ & & & \sigma_v^2 \end{bmatrix} \quad (1.8)$$

With r receive antennas, the maximum number of degrees of freedom at the receiver can be r . From the discussion in Section 1.1, we expect that the interference can be suppressed completely for a signal with s spatial streams as long as $m + s \leq r$. Consider the case when a single spatial stream is present along with $r - 1$ interferers ($s = 1, m = r - 1$), and $\mathbf{W}_{\text{SMI}}$ is used as the spatial interference suppression filter. The output SINR ρ can then be simplified to

$$\rho = \frac{\text{Tr}(\mathbf{W}_{\text{SMI}} \mathbf{H}_s \mathbf{R}_s \mathbf{H}_s^H \mathbf{W}_{\text{SMI}}^H)}{\text{Tr}(\mathbf{W}_{\text{SMI}})} \quad (1.9)$$

The eigen values of $\mathbf{W}_{\text{SMI}}$ are equal to

$$\begin{aligned} \mathbf{\Lambda}_{\text{SMI}} &= \begin{bmatrix} \frac{1}{\sigma_{\gamma(1)}^2 + \sigma_v^2} & & & \\ & \dots & & \\ & & \frac{1}{\sigma_{\gamma(m)}^2 + \sigma_v^2} & \\ & & & \frac{1}{\sigma_v^2} \end{bmatrix} \\ &\approx \begin{bmatrix} \epsilon & & & \\ & \dots & & \\ & & \epsilon & \\ & & & \frac{1}{\sigma_v^2} \end{bmatrix} \end{aligned} \quad (1.10)$$

where (1.10) is true when $\sigma_{\gamma(i)}^2 \gg \sigma_v^2, \sigma_{\gamma(i)}^2 \gg 1$. Thus, when the interferers are much stronger than the signal (which is assumed to be unit power) and the noise, then applying the matrix $\mathbf{W}_{\text{SMI}}$ nulls out the eigen modes corresponding to the m interferers, and leaves $r - m$ eigen modes with an SINR equal to the SNR on average. The eigen value matrix $\mathbf{\Lambda}_{\text{SMI}}$ therefore has

one channel of strength $\frac{1}{\sigma_v^2}$, and the other channels are nulled to ϵ . However, when the number of interferers is more than $r - s$, it is not possible to null the interference without nulling the signal as well, and this limits the interference suppression ability. This limit on interference suppression can be quantified in terms of the SINR gain, which is defined as the gain in SINR from the input of the filter $\mathbf{W}$ to the output

$$\rho(\text{gain})dB = \rho(\mathbf{p})dB - \rho(\mathbf{y})dB \quad (1.11)$$

$\rho(\text{gain})$ represents the amount of interference suppression. If the interference can be completely suppressed, $\rho(\text{gain})$ increases linearly with the interference strength expressed in dB . However, $\rho(\text{gain})$ saturates once the limit on the interference suppression ability of the system is reached.

1.3 Frontend Filters for Packet Based Communication Systems

Packet based communication systems, as the name suggests function on a per packet basis. This means that the receiver needs to detect the presence of a packet, perform the synchronization operations, estimate the channel and only then can the payload be decoded. In order to decode the signal without having a packet error, each of these features needs to be functional even in the presence of strong interference. Thus, it is essential to suppress the interference at the frontend i.e. soon after the Analog to Digital Conversion. This could be achieved by using a frontend filter (Fig. 1.2).

Several frontend filters based on the spatial covariance (second order statistics) of the received signal have been considered for this purpose [15]. The objective generally is to maximize the signal to interference plus noise ratio (SINR) at the output of the frontend filter. For common packet detection algorithms like those based on correlating the received signal with a known signal, this would in turn enable better packet detection. Thus, the filter $\mathbf{W}_{\text{MSINR}}$ would be the ideal candidate for such a system. However, even obtaining an accurate estimate of the spatial covariance of the signal is hard in the presence of strong interference. Therefore, it might be easier to implement a frontend filter based on $\mathbf{W}_{\text{SMI}}$, which depends only on the spatial covariance of the interference plus noise.

Figure 1.2: Frontend filter for suppressing interference in packet based communication systems

Frontend filters can also be an attractive option for packet based systems since this does not entail making changes to the existing receiver architecture. Take for example the MMSE MIMO decoder.

If the MMSE MIMO decoder is applied directly to the incoming signal $\mathbf{y}$, the estimated signal at the output of the MMSE MIMO decoder, $\hat{\mathbf{x}}_s$, can be obtained using (1.3) and is given by

$$\hat{\mathbf{x}}_s = \mathbf{H}_s^H (\mathbf{H}_s \mathbf{R}_s \mathbf{H}_s^H + \mathbf{H}_i \mathbf{R}_i \mathbf{H}_i^H + \mathbf{R}_v)^{-1} \mathbf{y} \quad (1.12)$$

However, as discussed before, this decoder needs channel estimates. Moreover, even if we can detect the payload using the MMSE decoder, we still need the frontend filter for enabling functions related to packet detection and synchronization.

Now if the incoming signal $\mathbf{y}$ is passed through the frontend filter $\mathbf{W}$, and the receiver architecture remains unchanged, then the effective channel estimate is equal to $\mathbf{W} \mathbf{H}_s$. The effective covariance of the interference plus noise equals $\mathbf{W}(\mathbf{R}_\gamma + \mathbf{R}_v) \mathbf{W}^H$. Therefore, the effective MMSE filter is equal to

$$\begin{aligned} \bar{\mathbf{W}}_{\text{MMSE}} &= (\mathbf{W} \mathbf{H}_s)^H ((\mathbf{W} \mathbf{H}_s) \mathbf{R}_s (\mathbf{W} \mathbf{H}_s)^H + \mathbf{W} \mathbf{R}_\gamma \mathbf{W}^H + \mathbf{W} \mathbf{R}_v \mathbf{W}^H)^{-1} \\ &= \mathbf{H}_s^H (\mathbf{H}_s \mathbf{R}_s \mathbf{H}_s^H + \mathbf{H}_i \mathbf{R}_i \mathbf{H}_i^H + \mathbf{R}_v)^{-1} \mathbf{W}^{-1} \end{aligned} \quad (1.13)$$

The resulting estimate obtained by passing the clean signal $\mathbf{p}$ through the effective MMSE filter

$\bar{\mathbf{W}}_{\text{MMSE}}$ is given by

$$\begin{aligned}\bar{\mathbf{x}}_s &= \mathbf{W}_{\text{MMSE}}\mathbf{p} \\ &= \mathbf{H}_s^H(\mathbf{H}_s\mathbf{R}_s\mathbf{H}_s^H + \mathbf{H}_i\mathbf{R}_i\mathbf{H}_i^H + \mathbf{R}_v)^{-1}\mathbf{W}^{-1}\mathbf{W}\mathbf{y}\end{aligned}\tag{1.14}$$

$$= \hat{\mathbf{x}}_s\tag{1.15}$$

Therefore the estimated signal remains unaffected by the application of the frontend filter. This means that this interference suppression solution can be applied without making any changes to the receiver, as depicted in Fig. 1.2.

1.4 Concluding Remarks

Spatial interference suppression exploits additional degrees of freedom by virtue of multiple antennas to null out interference. Moreover, the interference suppression ability is related to the number of interferers and signal spatial streams being lesser than the available degrees of freedom at the receiver. In the case of second order statistic based interference suppression, this translates to a limit on the number of dominant eigen modes in the interference spatial covariance matrix. Second order statistic based interference suppression is also an attractive option for frontend filters in packet based communication systems, which need robust packet detection and synchronization functionality even in the presence of strong interference.

CHAPTER 2

Implementation Impairments

RF impairments in the analog circuitry have been known to limit the performance of communication systems, and their impact is more pronounced as we move to multi-antenna and multi-carrier systems. A serious RF impairment in direct conversion receivers is IQ mismatch. While several works in the literature have looked at performance degradation and compensation schemes for IQ mismatch in communication systems [23] - [28], the effect of IQ mismatch on Spatial Interference Suppression algorithms has not been investigated before. In this chapter, we show how IQ mismatch can limit the interference suppression ability of second order statistic based interference suppression algorithms. Section 2.1 presents the IQ mismatch model. Section 2.2 explains how IQ mismatch leads to the creation of additional eigen modes in the interference spatial covariance matrix, that arise from the image band of the interference signal. Section 2.3 presents simulation results demonstrating the resulting performance degradation in flat fading channels.

2.1 IQ Mismatch

IQ mismatch occurs in quadrature mixers due to improper matching between the I and Q branches (Fig. 2.1). A simplified model of IQ mismatch is described below [23]. Any deviation from the ideal phase difference of $\frac{\pi}{2}$ between the I and Q branches is known as a phase mismatch, denoted by ϕ . A gain mismatch is caused if the amplitudes of the I and Q branches are unequal, and is denoted by α (when mentioned in dB, the gain mismatch equals $10 \log_{10}(1 + \alpha)$).

In the presence of IQ mismatch, the local oscillator signal $y_{LO}(t)$, operating at a frequency

Figure 2.1: IQ mismatch in quadrature mixers

ω_0 is given by

$$y_{LO}(t) = (1 + \alpha) \cos(\omega_0 t + \frac{\phi}{2}) - j(1 - \alpha) \sin(\omega_0 t - \frac{\phi}{2}) \quad (2.1)$$

After some algebraic manipulation, this yields

$$y_{LO}(t) = \underbrace{(\cos \frac{\phi}{2} + j\alpha \sin \frac{\phi}{2})}_{u_1} e^{-j\omega_0 t} + \underbrace{(\alpha \cos \frac{\phi}{2} - j \sin \frac{\phi}{2})}_{v_1} e^{j\omega_0 t} \quad (2.2)$$

$$= u_1 e^{-j\omega_0 t} + v_1 e^{j\omega_0 t} \quad (2.3)$$

Now if $y(t)$ is the complex baseband equivalent of the incoming signal $y_{RF}(t)$,

$$y_{RF}(t) = y(t)e^{j\omega_0 t} + y^*(t)e^{-j\omega_0 t} \quad (2.4)$$

The output of the quadrature mixer will be distorted due to IQ mismatch. Using (2.3), the complex baseband equivalent of the distorted signal $z(t)$ can be represented by

$$\begin{aligned} z(t) &= y_{RF}(t)y_{LO}(t) \\ &= u_1 y(t) + v_1 y^*(t) \end{aligned} \quad (2.5)$$

where we assume that the higher frequency components of $z(t)$ have been filtered out. The equivalent signal in the frequency domain $Z(f)$ can be obtained from the property of the fourier transform, $y^*(t) \longleftrightarrow Y^*(-f)$

$$Z(f) = u_1 Y(f) + v_1 Y^*(-f) \quad (2.6)$$

We therefore see that the baseband signal $Z(f)$ is affected by interference from the image band $Y^*(-f)$ of the received signal, as illustrated in Fig. 2.2. u_1 and v_1 denote the contributions

Figure 2.2: Additional interference from the image band due to IQ mismatch

of the desired signal band and the image band to the received signal. In the ideal case when there is no IQ mismatch, $\alpha = 0$, $\phi = 0$ and therefore $v_1 = 0$. Thus in the absence of IQ mismatch, direct conversion receivers have infinite image suppression. However, analog receivers today can achieve atmost 35-40 dB of image suppression [24]. Interference suppression systems typically have to deal with interference signals which can be more than 30 dB stronger than the desired signal, which means that the image band of the interferer might be of the order of or even higher than the signal strength (Fig. 2.2).

The effect of IQ mismatch can be further observed in the time domain. The received signal in the complex baseband, y can be written in terms of its I and Q components, y_I and y_Q respectively.

$$y = y_I + jy_Q \quad (2.7)$$

Using this equation, and substituting the values of u_1 and v_1 in (2.5) leads to

$$z = \left(\cos \frac{\phi}{2} + j\alpha \sin \frac{\phi}{2}\right)(y_I + jy_Q) + \left(\alpha \cos \frac{\phi}{2} - j \sin \frac{\phi}{2}\right)(y_I - jy_Q) \quad (2.8)$$

The I and Q components of the distorted signal, z_I and z_Q can be simplified to

$$z_I = [(1 + \alpha) \cos \frac{\phi}{2}]y_I - [(1 + \alpha) \sin \frac{\phi}{2}]y_Q \quad (2.9)$$

$$z_Q = [(1 - \alpha) \cos \frac{\phi}{2}]y_Q - [(1 - \alpha) \sin \frac{\phi}{2}]y_I \quad (2.10)$$

Hence we see that IQ mismatch causes a mixing of the I and Q components y_I and y_Q , which would otherwise have been orthogonal. In the absence of IQ mismatch $\alpha = 0$ and $\phi = 0$. Therefore $z_I = y_I$ and $z_Q = y_Q$. This is also evident from (2.5), since $v_1 = 0$ in the absence of IQ mismatch.

In general, IQ mismatch might occur due to any kind of mismatch between the I and Q branches, not necessarily arising from the quadrature mixer. For example, if the gain on the I branch is different from the gain on the Q branch, this again would result in IQ mismatch. (2.5) - (2.6) still hold even in such cases, but the parameters u_1 and v_1 are different.

Assuming the IQ mismatch on each RF chain to be independent, the distortion caused by IQ mismatch (2.5) can be extended to the multi-antenna case

$$\begin{aligned} \mathbf{z} &= \mathbf{U}\mathbf{y} + \mathbf{V}\mathbf{y}^* \\ &= \mathbf{U}\mathbf{H}_s\mathbf{x}_s + \mathbf{V}\mathbf{H}_s^*\mathbf{x}_s^* + \mathbf{U}\boldsymbol{\gamma} + \mathbf{V}\boldsymbol{\gamma}^* + \mathbf{w} \end{aligned} \quad (2.11)$$

$$= \mathbf{U}\mathbf{H}_s\mathbf{x}_s + \mathbf{V}\mathbf{H}_s^*\mathbf{x}_s^* + \boldsymbol{\psi} + \mathbf{w} \quad (2.12)$$

where $\mathbf{U} = \text{diag}(u_1, \dots, u_r)$, $\mathbf{V} = \text{diag}(v_1, \dots, v_r)$, $\mathbf{w}$ is the effective noise after IQ mismatch and we define $\boldsymbol{\psi} = \mathbf{U}\boldsymbol{\gamma} + \mathbf{V}\boldsymbol{\gamma}^*$, the resultant received interference signal vector.

2.2 Performance of Spatial Interference Suppression

Let us consider the spatial interference suppression algorithms from Section 1.2 applied to the distorted signal $\mathbf{z}$. The spatial covariance matrix of the interference plus noise $\mathbf{R}_{\boldsymbol{\psi}+\mathbf{w}}$ is given by

$$\begin{aligned} \mathbf{R}_{\boldsymbol{\psi}+\mathbf{w}} &= \mathbf{E}[\boldsymbol{\psi}\boldsymbol{\psi}^H] + \mathbf{E}[\mathbf{w}\mathbf{w}^H] \\ &= \mathbf{U}\mathbf{R}_\gamma\mathbf{U}^H + \mathbf{V}\mathbf{R}_\gamma^*\mathbf{V}^H + \mathbf{R}_w \end{aligned} \quad (2.13)$$

Figure 2.3: Creation of additional eigen modes in the interference spatial covariance matrix

Assuming m interference sources and r receive antennas, the eigen values of this matrix are given by

$$\mathbf{\Lambda}_{\psi+\mathbf{w}} = \begin{bmatrix} \sigma_{\psi(1)}^2 & & & \\ & \dots & & \\ & & \dots & \\ & & & \sigma_{\psi(\min(2m,r))}^2 \end{bmatrix} + \begin{bmatrix} \sigma_w^2 & & & \\ & \dots & & \\ & & \dots & \\ & & & \sigma_w^2 \end{bmatrix} \quad (2.14)$$

We observe from (2.13) that once the image band of the interference signal ($\mathbf{V}\mathbf{R}_\gamma^*\mathbf{V}^H$) rises above the noise floor ($\mathbf{R}_w$), it creates additional eigen modes in the spatial covariance matrix. This is because every tone in the image band of the interference signal is independent of the corresponding tone in the primary band of the interference signal. In the worst case the number of interference eigen modes double (2.14), as every strong interferer now generates interference from its image band as well. This would happen when the image band of all interferers rise above the noise floor in Fig. 2.3.

In Chapter 1, it was shown that the interference can be completely suppressed as long as the total number of interference sources and signal spatial streams is not more than the degrees of freedom at the receiver (1.10). If we have s signal spatial streams, m interferers and r receive antennas, then (2.14) implies that the interference can be completely suppressed only if

Table 2.1: Simulation Parameters

Parameter	Value
IQ Gain mismatch (α)	0.1 dB
IQ Phase mismatch (ϕ)	0.5 deg

$m \leq \lfloor \frac{r-s}{2} \rfloor$. If there are more interference sources, part of the signal ends up being nulled as well. Thus we can effectively suppress only half as many interferers as before. Moreover, the SINR gain which indicates the amount of interference suppression saturates once we have additional interference sources.

2.3 Simulation Results

Sample Matrix Inversion based spatial interference suppression is used as a frontend filter for a MIMO system with 4 antennas. The signal uses QPSK modulation, and the interference is generated using independent gaussian sources. The system is simulated in a rayleigh flat fading channel environment. The IQ mismatch parameters are given in Table 2.1. These parameters are typical of hardware implementations today [31].

Fig. 2.4a demonstrates the effect of IQ mismatch on the SINR Gain. Since the total number of signal spatial streams and interference sources is no more than the degrees of freedom at the receiver (which is 4, due to four antennas), in the absence of IQ mismatch the interference can be suppressed completely. This is obvious from Fig. 2.4b, where the SINR gain is found to be equal to the Interference to Noise Ratio (INR), indicating that the interference is indeed being nulled to below the noise floor.

In the presence of IQ mismatch, as discussed in Section 2.2 only 1 interference source can be suppressed if there are 2 spatial streams. This can be seen from Fig. 2.4a, where the SINR gain remains unaffected if there is only 1 interference source. However, the SINR gain saturates as the interference increases if there are 2 interference sources.

Another interesting point is that the SINR gain saturates around 40 dB, and this is the

(a) Interference suppression ability as a function of the number of interferers (SINR Gain vs SIR for a signal with two spatial streams at 25dB SNR)

(b) Suppression of interference to the noise floor (SINR Gain vs INR for a signal with two spatial streams at 25dB SNR)

Figure 2.4: Limited interference suppression due to IQ mismatch

amount by which the interference from the image band due to IQ mismatch can be suppressed using analog processing. Beyond this point, the image band of the interference rises above the noise floor and creates additional eigen modes in the interference spatial covariance matrix, thereby limiting the interference suppression ability. This can be seen from Fig. 2.5. When the interference is weak, there are only two dominant eigen modes in the spatial covariance matrix

Figure 2.5: Impact of IQ mismatch on eigen modes of interference spatial covariance matrix (Eigen values of the interference plus noise when two interferers are present, 25dB SNR)

which correspond to the two interferers. The remaining two eigen modes are occupied by the noise. However, as the interference rises, the other two eigen modes are also occupied by the interference. These additional eigen modes are observed to be roughly 40dB weaker than the interference eigen modes.

Finally, Fig. 2.6 illustrates the impact of increasing IQ mismatch on the SINR gain. As expected, the interference suppression ability reduces as the IQ mismatch increases.

2.4 Concluding Remarks

In this chapter it was shown how IQ mismatch limits the interference suppression ability of second order statistic based interference suppression algorithms. The image band of the interference gives rise to additional eigen modes in the interference spatial covariance matrix, which limits the amount of interference that can be suppressed. In flat fading channel environments, this causes the SINR gain of the frontend filter to saturate, and the performance of the packet based communication system can be impacted adversely.

Figure 2.6: Impact of IQ mismatch parameters (SINR Gain vs SIR for a signal with two spatial streams, and two interferers in the environment, 25dB SNR)

CHAPTER 3

IQ Mismatch Compensation for Narrowband Spatial Interference Suppression Systems

Several techniques have been proposed for IQ mismatch compensation in communication systems. Many of these compensation schemes are based on estimating the IQ mismatch parameters [25], or involve the use of some sort of training signals for estimating the effective channel after IQ mismatch [26] [27]. However, estimating the IQ mismatch parameters can be extremely challenging in the presence of strong interference. Moreover, IQ mismatch limits the interference suppression ability at the frontend, and therefore schemes that compensate for IQ mismatch during channel estimation and decoding have to deal with noisy training signals. Therefore these compensation techniques cannot be directly applied to interference suppression systems. This chapter presents the concept of expanding the received signal sub-space to compensate for IQ mismatch at the frontend in spatial interference suppression systems. The idea has also been used to correct for IQ mismatch in MIMO-OFDM systems [28]. Section 3.1 presents the expanded subspace concept for mitigating IQ mismatch, Section 3.2 proposes an interference suppression architecture that is inspired by this idea and Section 3.3 presents simulation results demonstrating the additional interference suppression achievable through this architecture.

3.1 The Expanded Subspace Concept

Chapter 2 showed how IQ mismatch causes interference from the conjugate of the image band of the signal when observed in the frequency domain, or equivalently from the conjugate of the signal when observed in the time domain. This suggests, on an intuitive level, that it might be possible to mitigate this effect if the signal and its conjugate are considered together, since

this results in more information for the decoding process. Consider the following formulation for MIMO systems [28]. From Chapter 2, we have the following equations

$$\mathbf{z} = \mathbf{U}\mathbf{H}_s\mathbf{x}_s + \mathbf{V}\mathbf{H}_s^*\mathbf{x}_s^* + \psi + \mathbf{w} \quad (3.1)$$

$$\mathbf{z}^* = \mathbf{V}^*\mathbf{H}_s\mathbf{x}_s + \mathbf{U}^*\mathbf{H}_s^*\mathbf{x}_s^* + \psi^* + \mathbf{w}^* \quad (3.2)$$

where (3.2) is the conjugate of (3.1). At sample time n , this can be written as

$$\underbrace{\begin{pmatrix} \mathbf{z}(\mathbf{n}) \\ \mathbf{z}^*(\mathbf{n}) \end{pmatrix}}_{\tilde{\mathbf{z}}} = \underbrace{\begin{pmatrix} \mathbf{U}\mathbf{H}_s & \mathbf{V}\mathbf{H}_s^* \\ \mathbf{V}^*\mathbf{H}_s & \mathbf{U}^*\mathbf{H}_s^* \end{pmatrix}}_{\tilde{\mathbf{H}}} \underbrace{\begin{pmatrix} \mathbf{x}(\mathbf{n}) \\ \mathbf{x}^*(\mathbf{n}) \end{pmatrix}}_{\tilde{\mathbf{x}}} + \underbrace{\begin{pmatrix} \psi(\mathbf{n}) \\ \psi^*(\mathbf{n}) \end{pmatrix}}_{\tilde{\psi}} + \underbrace{\begin{pmatrix} \mathbf{w}(\mathbf{n}) \\ \mathbf{w}^*(\mathbf{n}) \end{pmatrix}}_{\tilde{\mathbf{w}}} \quad (3.3)$$

$$\tilde{\mathbf{z}} = \tilde{\mathbf{H}}\tilde{\mathbf{x}} + \tilde{\psi} + \tilde{\mathbf{w}} \quad (3.4)$$

Here $\mathbf{U}\mathbf{H}_s$ and $\mathbf{U}^*\mathbf{H}_s^*$ in the effective channel matrix of (3.3) represent contribution of the desired signal. By using the received signal and its conjugate to decode the signal, the information from $\mathbf{V}\mathbf{H}_s^*$ and $\mathbf{V}^*\mathbf{H}_s$ can now be used to estimate the signal instead of considering it as interference. As long as it is possible to estimate the expanded channel matrix $\tilde{\mathbf{H}}$ of size $2r \times 2s$ (where r is the number of receive antennas and s is the number of signal spatial streams) the effect of IQ mismatch can be mitigated.

It is important to note here that the parameters of the IQ mismatch are not explicitly being estimated, rather it is the effective channel after IQ mismatch that is estimated. Estimating the parameters of the IQ mismatch can be very challenging in the presence of strong interference. Moreover, since the degradation is caused by the interferer's image band, the use of pilot signals to estimate these parameters will not be effective. Thus compensation techniques that rely on the estimation of the IQ mismatch parameters cannot be directly applied to the interference suppression problem.

In fact, in a packet based communication system that uses a frontend filter for interference suppression, even using (3.3) to decode the signal can be problematic. This is because in the

presence of IQ mismatch the interference is not suppressed completely at the output of the frontend filter $\mathbf{W}$, which might lead to poor estimation of the effective channel $\tilde{\mathbf{H}}$ or even worse the packet might not be detected at all. It is therefore essential to suppress the interference at the frontend completely even if we can compensate for IQ mismatch at the MIMO decoder by using the architecture proposed in [28].

The idea of considering the signal and its conjugate jointly can be extended to the frontend filter $\mathbf{W}$. Using the system of equations (3.3), the spatial covariance $\mathbf{R}_{\tilde{\psi}+\tilde{\mathbf{w}}}$ of the interference plus noise can be computed to be

$$\mathbf{R}_{\tilde{\psi}+\tilde{\mathbf{w}}} = \mathbf{E} \left[\begin{pmatrix} \psi + \mathbf{w} \\ \psi^* + \mathbf{w}^* \end{pmatrix} \begin{pmatrix} \psi + \mathbf{w} & \psi^* + \mathbf{w}^* \end{pmatrix}^H \right] \quad (3.5)$$

The eigen values of this covariance matrix can be represented as

$$\mathbf{\Lambda}_{\tilde{\psi}+\tilde{\mathbf{w}}} = \begin{bmatrix} \sigma_{\psi(1)}^2 & & & \\ & \dots & & \\ & & \sigma_{\psi(2m)}^2 & \\ & & & \epsilon \end{bmatrix} + \begin{bmatrix} \sigma_w^2 & & & \\ & \dots & & \\ & & \dots & \\ & & & \sigma_w^2 \end{bmatrix} \quad (3.6)$$

This is in essence an expansion of the received signal sub-space to achieve $2r$ degrees of freedom, where r is the number of receive antennas. However, the distorted interference signal consists of interference from m interferers and its m mirror tones, and therefore still has only $2m$ eigen modes in the expanded sub-space. For a signal with s spatial streams, the interference can now be completely suppressed as long as $2m \leq 2r - 2s$ or $m + s \leq r$. This is the same condition as that in Section 1.2. Thus, if the interference could be completely suppressed in the absence of IQ mismatch using spatial interference suppression, then using the expanded sub-space concept the interference can be completely suppressed even in the presence of IQ mismatch.

While it might appear that we are doubling the number of complex degrees of freedom from r to $2r$, in essence the expanded sub-space is merely a transformation of the $2r$ real degrees of freedom. This can be seen by considering (2.9) - (2.10), written in matrix form.

$$\begin{pmatrix} \mathbf{z_I} \\ \mathbf{z_Q} \end{pmatrix} = \begin{pmatrix} (1 + \alpha) \cos \frac{\phi}{2} & -(1 + \alpha) \sin \frac{\phi}{2} \\ (1 - \alpha) \sin \frac{\phi}{2} & (1 - \alpha) \cos \frac{\phi}{2} \end{pmatrix} \begin{pmatrix} \mathbf{y_I} \\ \mathbf{y_Q} \end{pmatrix} \quad (3.7)$$

Figure 3.1: Interference suppression architecture with IQ mismatch compensation for narrow-band receivers

The expanded sub-space solution can be reformulated using (3.7), where the expanded sub-space would be formed by the I and Q components $\mathbf{z}_I$ and $\mathbf{z}_Q$, instead of the signal and its conjugate $\mathbf{z}$ and $\mathbf{z}^*$. If the spatial covariance of the interference is now computed using (3.7), then an equivalent sub-space with $2r$ real degrees of freedom is formed, and the same argument for interference suppression will apply.

This concept of expanding the sub-space can be applied to second order statistic based interference suppression in general, since these filters involve inversion of covariance matrices.

3.2 Architecture for IQ Mismatch Compensation

Fig. 3.1 depicts an interference suppression architecture with IQ mismatch compensation, which is based on the expanded subspace concept. The basic idea is to compute the interference suppression filter by considering the signal and its conjugate together. The interference suppression

filter $\tilde{\mathbf{W}}_{\text{SMI}}$ can be calculated in terms of the expanded spatial covariance matrix $\mathbf{R}_{\psi+\mathbf{w}}$ as

$$\tilde{\mathbf{W}}_{\text{SMI}} = \mathbf{R}_{\psi+\mathbf{w}}^{-1} \quad (3.8)$$

This filter is of size $2r \times 2r$, and is formed by inverting a matrix of size $2r \times 2r$. As noted in [28], the additional complexity associated with the expanded subspace concept is inverting a matrix with double the dimensions. The distorted signal $\tilde{\mathbf{z}}$ is filtered to produce $\tilde{\mathbf{p}}$, which is given by

$$\underbrace{\begin{pmatrix} \mathbf{p}(\mathbf{n}) \\ \mathbf{p}^*(\mathbf{n}) \end{pmatrix}}_{\tilde{\mathbf{p}}} = \tilde{\mathbf{W}}_{\text{SMI}} \underbrace{\begin{pmatrix} \mathbf{z}(\mathbf{n}) \\ \mathbf{z}^*(\mathbf{n}) \end{pmatrix}}_{\tilde{\mathbf{z}}} \quad (3.9)$$

We note here that $\mathbf{p}^*(\mathbf{n})$ is merely the conjugate of the desired signal $\mathbf{p}(\mathbf{n})$. Therefore $\mathbf{p}^*(\mathbf{n})$ need not be computed explicitly, and the lower half of the matrix $\tilde{\mathbf{W}}_{\text{SMI}}$ does not have to be computed. If $\tilde{\mathbf{W}}_{\text{SMI}}$ is written in block matrix form consisting of four $r \times r$ matrices,

$$\tilde{\mathbf{W}}_{\text{SMI}} = \begin{pmatrix} \mathbf{W}_{11} & \mathbf{W}_{12} \\ \mathbf{W}_{21} & \mathbf{W}_{22} \end{pmatrix} \quad (3.10)$$

Then effectively only (3.11) needs to be implemented

$$\mathbf{p}(\mathbf{n}) = \begin{pmatrix} \mathbf{W}_{11} & \mathbf{W}_{12} \end{pmatrix} \begin{pmatrix} \mathbf{z}(\mathbf{n}) \\ \mathbf{z}^*(\mathbf{n}) \end{pmatrix} \quad (3.11)$$

However, this still needs the inversion of the $2r \times 2r$ matrix $\mathbf{R}_{\psi+\mathbf{w}}$. It is worth noting here that it is now essential to decode the signal and its conjugate jointly (using (3.3)) in the generic receiver following the interference suppression filter. This is because (3.11) combines the signal with its conjugate, and the resulting mixing needs to be corrected at the decoder.

3.3 Simulation Results

Sample Matrix Inversion based spatial interference suppression [16] was used as the frontend filter for an 802.11n compliant MIMO-OFDM transceiver with four antennas and complete synchronization and packet detection functionality [29]. The simulations were carried out using the flat fading channel model from the 802.11n standard [30] - Channel A. As a representative case,

Table 3.1: Simulation Parameters

Parameter	Value
IQ Gain mismatch (α)	0.1 dB
IQ Phase mismatch (ϕ)	0.5 deg

the system is simulated using MCS 10 [29] (QPSK modulation with rate $\frac{3}{4}$ FEC and 2 spatial streams) in the presence of two gaussian interference sources. The IQ mismatch parameters are given in Table 3.1.

Fig. 3.2a illustrates the SINR gain at packet synch as a function of the Signal to Interference (SIR) ratio. As discussed in Chapter 2, IQ mismatch limits the interference suppression ability to around 40dB, which is the amount of image suppression expected from analog processing. By using the interference suppression architecture proposed in this chapter, it is possible to eliminate this saturation, since even the image band of the interference signal can be suppressed in the expanded sub-space.

The Packet Error Rate (PER) of the system is shown in Fig. 3.2b. There are two spatial streams and two interferers, which is equal to the number of degrees of freedom at the receiver (that has four antennas). In the absence of IQ mismatch the PER curve is flat for decreasing values of the Signal to Interference ratio (SIR). This indicates that in a flat fading channel the interference can be suppressed completely irrespective of the magnitude of the interference. As discussed in Chapter 1, it is possible to use the degrees of freedom at the receiver either to suppress interference or receive multiple signal spatial streams. This is further supported by the fact that M interferers can be completely suppressed in a narrowband channel as long as the available degrees of freedom at the receiver $N > M$ [7].

In the presence of IQ mismatch, the degrees of freedom at the receiver are no longer sufficient to suppress the interference completely. The PER is a function of packet detection, synchronization, channel estimation and decoding. As discussed in Section 3.1, [28] presents an approach for estimation of MIMO signals in the presence of IQ mismatch, which compensates for the IQ mismatch at the decoding stage. The IQ mismatch case in Fig. 3.2b indicates the PER when

(a) SINR Gain vs SIR for an 802.11n compliant system using Sample Matrix Inversion based interference suppression (MCS 10, 2 interferers, 25dB SNR)

(b) Packet Error Rate vs SIR for an 802.11n compliant system using Sample Matrix Inversion based interference suppression (MCS 10, 2 interferers, 25dB SNR)

Figure 3.2: IQ mismatch compensation in flat fading channels (Channel A)

$\mathbf{W}_{\text{SMI}}$ is used as the frontend filter, and the MMSE MIMO decoder for the receiver is based on the IQ mismatch compensation scheme presented in [28]. Despite the compensation, the PER still degrades since the interference cannot be suppressed at the frontend, thus severely affecting the other parts of the receiver, such as packet detection and channel estimation.

If the interference suppression architecture presented in this chapter is used for the frontend

Figure 3.3: Packet Detection Error Rate vs SIR for an 802.11n compliant system using Sample Matrix Inversion based interference suppression (MCS 10, 2 interferers, 25dB SNR, Channel A)

filter along with the MMSE MIMO decoder from [28], it is possible to mitigate the effect of IQ mismatch. This is evident from the compensation case in Fig. 3.2b. However, this MMSE MIMO decoder involves the estimation of a channel matrix that has 4x more coefficients, and the effective noise covariance has 4x more coefficients as well. Therefore, there is a small performance gap with respect to the no IQ mismatch case.

Fig. 3.3 illustrates the contribution of the Packet Detection Error Rate (PDER) to the Packet Error Rate. We see here that IQ mismatch causes a degradation to the PDER, which supports the claim that IQ mismatch needs to be compensated for at the frontend, and not just at the decoder. By using the interference suppression architecture presented in this chapter, it is possible to suppress the interference completely so that Packet Detection no longer contributes significantly to the Packet Error Rate. Also, the increase in Packet Detection Error begins to show after the Packet Error Rate has started degrading. This is because channel estimation and synchronization errors start effecting the performance of the system before we reach the extreme case of the packet not being detected at all.

Fig. 3.4 provides more insight into interference suppression in the expanded sub-space. Fig. 3.4a shows that additional eigen modes are generated in the interference spatial covariance matrix due to IQ mismatch, and this limits the interference suppression as the eigen modes now double

(a) Eigen values of the interference plus noise in the spatial covariance matrix (2 interferers, 25dB SNR)

(b) Eigen values of the interference plus noise in the expanded spatial covariance matrix (2 interferers, 25dB SNR)

Figure 3.4: Compensating for IQ mismatch in the expanded sub-space (Flat fading channel)

thus occupying the complete sub-space. However, in the expanded sub-space the degrees of freedom at the receiver are doubled while the eigen modes of the interference remain the same. Therefore, there are now additional eigen modes that only have contribution of the thermal noise.

3.4 Concluding Remarks

This chapter introduced an interference suppression architecture to compensate for IQ mismatch, that is inspired by the idea of expanding the received signal sub-space. It was shown for flat fading channels that compensating for IQ mismatch at the MIMO decoder is insufficient in a packet based communication system, since packet detection functionality can be hampered. However, the architecture presented here mitigates IQ mismatch at the frontend filter itself, and can eliminate the limit on interference suppression due to IQ mismatch almost completely.

CHAPTER 4

IQ Mismatch Compensation for Wideband Spatial Interference Suppression Systems

Wideband systems typically have to deal with frequency selective channels. The channel's frequency selectivity can hamper the performance of interference suppression, since the delayed copies of the interference signal give rise to additional interference eigen modes in the spatial covariance matrix. Traditionally, multipath in wideband communication systems has been taken care of by building equalizers that depend on training data or using OFDM systems that require a Cyclic Prefix. However, the interference signal can be assumed to contain neither of these. In such a case, a convenient approach is to sub-band the signal [10] so that the effective channel in each of the sub-bands is more or less flat, and to apply the frontend filter to every sub-band. Section 4.1 discusses sub-banding architectures in more detail, Section 4.2 explains the expanded subspace concept in the frequency domain, Section 4.3 presents an interference suppression architecture that compensates for IQ mismatch in wideband systems and Section 4.4 presents simulation results in scenarios typical of indoor environments.

4.1 Sub-banding Architecture for Frequency Selective Channels

It was noted in [9] [10] that increasing the signal bandwidth in an antenna array increases the dispersion across the array. An effective approach to undo this effect is to divide the complete bandwidth into smaller segments. This is the idea behind sub-banding (Fig. 4.1). Sub-banding can be done using polyphase filter banks [19], but they tend to be computationally expensive. A more efficient way to implement the sub-banding operation is by using the Fast Fourier Transform (FFT), which is a linear transformation.

Figure 4.1: Sub-banding to convert wideband interference into smaller narrowband interference sources

A sub-banding architecture based on the FFT is depicted in Fig. 4.2. Since the bandwidth of each sub-band might be much smaller than the original bandwidth (depending on the number of FFT bins), one would expect the effective channel in each of these sub-bands to be more or less flat. However, the FFT operation creates filters with high side lobes [21]. These side-lobes cause the energy from neighboring sub-bands to leak into the primary sub-band, and this again causes a spreading of eigen modes. Typically, some sort of windowing function (like Hamming window) is used to suppress the energy from the side lobes [20]. However, this can cause the primary sub-band to become wider. As long as the effective channel in each of the sub-bands can be assumed to be flat, the interference suppression operation which was valid for narrowband systems can now be applied to each sub-band individually, and the resulting signals from each sub-band can be combined using the IFFT. The frontend filtering taking place in the frequency domain can be represented by the equation below.

$$\mathbf{P}(\mathbf{k}) = \mathbf{W}_{\text{SMI}} \mathbf{Y}(\mathbf{k}) \quad (4.1)$$

where $\mathbf{Y}(\mathbf{k})$, $\mathbf{P}(\mathbf{k})$ are the received signal before and after the interference suppression, in the frequency band with index k . $\mathbf{W}_{\text{SMI}}$ is the interference suppression filter that is calculated as the inverse of the spatial covariance of the interference plus noise for each sub-band.

The number of sub-bands is typically decided based on the frequency selectivity of the channel

Figure 4.2: Interference suppression architecture with sub-banding for wideband receivers

i.e. a channel that is more frequency selective will require a larger number of sub-bands to prevent additional interference eigen modes in the spatial covariance matrix. For a N -point FFT, the complexity associated with the interference suppression operation is N matrix inversions each of size $r \times r$, apart from the cost of the FFT and IFFT operations themselves.

The FFT and IFFT operations can be implemented using the Overlap and Add (OLA) method [22]. Here the N -point FFT is calculated from overlapping samples, and a new FFT is computed every D samples, where D is referred to as the overlap step. The windowing functions can also be implemented by matching the window length to the size of the FFT.

4.2 The Expanded Subspace Concept in Frequency Domain

Chapter 3 presented the expanded sub-space concept, and an interference suppression architecture based on that in the time domain. Similar to the manner in which interference suppression for wideband systems is done in the frequency domain, the interference suppression architecture with IQ mismatch compensation can also be extended to the frequency domain. Consider the

equation (3.3) from Chapter 3.

$$\underbrace{\begin{pmatrix} \mathbf{z}(\mathbf{n}) \\ \mathbf{z}^*(\mathbf{n}) \end{pmatrix}}_{\tilde{\mathbf{z}}} = \underbrace{\begin{pmatrix} \mathbf{U}\mathbf{H}_s & \mathbf{V}\mathbf{H}_s^* \\ \mathbf{V}^*\mathbf{H}_s & \mathbf{U}^*\mathbf{H}_s^* \end{pmatrix}}_{\tilde{\mathbf{H}}} \underbrace{\begin{pmatrix} \mathbf{x}(\mathbf{n}) \\ \mathbf{x}^*(\mathbf{n}) \end{pmatrix}}_{\tilde{\mathbf{x}}} + \underbrace{\begin{pmatrix} \psi(\mathbf{n}) \\ \psi^*(\mathbf{n}) \end{pmatrix}}_{\tilde{\psi}} + \underbrace{\begin{pmatrix} \mathbf{w}(\mathbf{n}) \\ \mathbf{w}^*(\mathbf{n}) \end{pmatrix}}_{\tilde{\mathbf{w}}} \quad (4.2)$$

$$\tilde{\mathbf{z}} = \tilde{\mathbf{H}}\tilde{\mathbf{x}} + \tilde{\psi} + \tilde{\mathbf{w}} \quad (4.3)$$

The equivalent signal in the frequency domain $\mathbf{Z}(k)$ can be obtained using the property of the N -point Discrete Fourier Transform $z^*(n) \longleftrightarrow Z^*(N - k + 2)$ (assuming $1 \leq n \leq N$ and $1 \leq k \leq N$)

$$\underbrace{\begin{pmatrix} \mathbf{Z}(k) \\ \mathbf{Z}^*(N - k + 2) \end{pmatrix}}_{\tilde{\mathbf{Z}}} = \tilde{\mathbf{\Lambda}} \underbrace{\begin{pmatrix} \mathbf{X}(k) \\ \mathbf{X}^*(N - k + 2) \end{pmatrix}}_{\tilde{\mathbf{X}}} + \underbrace{\begin{pmatrix} \mathbf{\Psi}(k) \\ \mathbf{\Psi}^*(N - k + 2) \end{pmatrix}}_{\tilde{\mathbf{\Psi}}} + \underbrace{\begin{pmatrix} \mathbf{W}(k) \\ \mathbf{W}^*(N - k + 2) \end{pmatrix}}_{\tilde{\mathbf{W}}} \quad (4.4)$$

$$\tilde{\mathbf{Z}} = \tilde{\mathbf{\Lambda}}\tilde{\mathbf{X}} + \tilde{\mathbf{\Psi}} + \tilde{\mathbf{W}} \quad (4.5)$$

where k is the sub-band index, and $\tilde{\mathbf{\Lambda}}$ is the equivalent channel in the frequency domain. Here again, as long as it is possible to estimate the expanded channel matrix $\tilde{\mathbf{\Lambda}}$ of size $2r \times 2s$ (where r is the number of receive antennas and s is the number of signal spatial streams) the effect of IQ mismatch can be mitigated.

The spatial covariance $\mathbf{R}_{\tilde{\mathbf{\Psi}} + \tilde{\mathbf{W}}}$ of the interference plus noise can be computed by considering the signal and the conjugate of its image tone jointly in the frequency domain and averaging over time.

$$\mathbf{R}_{\mathbf{\Psi} + \mathbf{W}} = \mathbf{E} \left[\begin{pmatrix} \mathbf{\Psi} + \mathbf{W} \\ \mathbf{\Psi}^* + \mathbf{W}^* \end{pmatrix} \begin{pmatrix} \mathbf{\Psi} + \mathbf{W} & \mathbf{\Psi}^* + \mathbf{W}^* \end{pmatrix}^H \right] \quad (4.6)$$

The eigen values of this covariance matrix can be represented as

$$\mathbf{\Lambda}_{\tilde{\Psi}+\tilde{\mathbf{W}}} = \begin{bmatrix} \sigma_{\tilde{\Psi}(1)}^2 & & & \\ & \dots & & \\ & & \sigma_{\tilde{\Psi}(2m)}^2 & \\ & & & \epsilon \end{bmatrix} + \begin{bmatrix} \sigma_{\tilde{\mathbf{W}}}^2 & & & \\ & \dots & & \\ & & \dots & \\ & & & \sigma_{\tilde{\mathbf{W}}}^2 \end{bmatrix} \quad (4.7)$$

Thus if the effective channel in each of the sub-bands was truly flat, then for a signal with s spatial streams, m interferers could have been completely suppressed with r receive antennas as long as $2m \leq 2r - 2s$ or $m + s \leq r$. However, despite the sub-banding architecture there will be some spreading of eigen modes as described in Section 4.1, which is why the interference can never be completely suppressed. Nevertheless, the expanded sub-space concept can still be used to compensate for IQ mismatch till the spreading of eigen modes becomes large.

4.3 Architecture for IQ Mismatch Compensation in Wideband Systems

Fig. 4.3 depicts an interference suppression architecture with IQ mismatch compensation, which is based on the expanded subspace concept. The idea is to convert the time domain signal into the frequency domain, suppress the interference in the expanded sub-space and then convert it back to the time domain. The received signal $\mathbf{z}$ is first sub-banded by passing it through an FFT, and the spatial covariance of the interference plus noise $\mathbf{R}_{\tilde{\Psi}+\tilde{\mathbf{W}}}$ is computed by considering a sub-band and the conjugate of its image band jointly. The modified frontend filter $\tilde{\mathbf{W}}_{\text{SMI}}$ can then be computed as

$$\tilde{\mathbf{W}}_{\text{SMI}} = \mathbf{R}_{\tilde{\Psi}+\tilde{\mathbf{W}}}^{-1} \quad (4.8)$$

This filter is of size $2r \times 2r$. Again, as noted in [28] and in Chapter 3, the additional complexity associated with the expanded subspace concept is inverting a matrix with double the dimensions. However, since the sub-band and the image band are considered together, the interference suppression needs to be performed for only half the total number of sub-bands. The filtering is applied to the signal $\tilde{\mathbf{Z}}$ formed by considering the signal and its image tone together in the

Figure 4.3: Interference suppression architecture with IQ mismatch compensation for wideband receivers

frequency domain, and the output of the filter $\tilde{\mathbf{P}}$ is given by

$$\underbrace{\begin{pmatrix} \mathbf{P}(\mathbf{k}) \\ \mathbf{P}^*(\mathbf{N} - \mathbf{k} + 2) \end{pmatrix}}_{\tilde{\mathbf{P}}} = \tilde{\mathbf{W}}_{\text{SMI}} \underbrace{\begin{pmatrix} \mathbf{Z}(\mathbf{k}) \\ \mathbf{Z}^*(\mathbf{N} - \mathbf{k} + 2) \end{pmatrix}}_{\tilde{\mathbf{Z}}} \quad (4.9)$$

Finally the clean signal $\mathbf{p}$ is produced by passing each sub-band and the conjugate of its image band through an IFFT. Therefore, by using (4.9) instead of (4.1), it is possible to compensate for IQ mismatch.

Since the FFT and IFFT blocks are already present in wideband interference suppression systems, this scheme would need $\frac{N}{2}$ matrix inversions each of size $2r \times 2r$. Thus the resulting complexity cost is an increase in the size of the spatial covariance matrix to be inverted, and the operation has to be performed for half the total number of sub-bands.

Additionally, the IQ mismatch compensation architecture can be applied to a wideband system irrespective of the FFT size of its sub-banding operation. The only constraint is that this architecture would be effective only if the performance degradation is actually being caused by the IQ mismatch. As discussed in the previous section, the frequency selectivity of the channel

Table 4.1: Simulation Parameters

Parameter	Value
IQ Gain mismatch (α)	0.1 dB
IQ Phase mismatch (ϕ)	0.5 deg
FFT Size (N)	32
OLA Step (D)	8
Window Type	Hamming

causes the eigen modes of the interference spatial covariance matrix to spread, and this spreading is more as the bandwidth increases or the number of sub-bands in the FFT decrease. Therefore, when the effective channel in each sub-band is not flat, the frequency selectivity of the channel itself limits the interference suppression ability and not the IQ mismatch. In this case we do not expect to see any gain by introducing IQ mismatch compensation.

4.4 Simulation Results

Sample Matrix Inversion based spatial interference suppression [16] was used as the frontend filter for an 802.11n compliant MIMO-OFDM transceiver with four antennas and complete synchronization and packet detection functionality [29]. The sub-banding operation in the interference suppression architecture was implemented using an Overlap and Add approach, with a Hamming window. The simulations were carried out using a frequency selective channel model from the 802.11n standard [30] - Channel C, which has a rms delay spread of 30ns. This is typical of indoor environments. As a representative case, the system is simulated using MCS 10 [29] (QPSK modulation with rate $\frac{3}{4}$ FEC and 2 spatial streams) in the presence of wideband gaussian interference. The IQ mismatch and sub-banding parameters are given in Table 4.1.

Fig. 4.4 illustrates the SINR gain for different FFT sizes in the sub-banding operation. For each curve, the OLA step is maintained to be a quarter of the FFT size. The first point to note is that the SINR gain saturates for frequency selective channels even in the absence of IQ

(a) SINR Gain vs SIR for sub-banding with FFT Size = 32, OLA Step = 8 (2 signal spatial streams, 1 interferer, 25dB SNR)

(b) SINR Gain vs SIR for sub-banding with FFT Size = 16, OLA Step = 4 (2 signal spatial streams, 1 interferer, 25dB SNR)

(c) SINR Gain vs SIR for sub-banding with FFT Size = 8, OLA Step = 2 (2 signal spatial streams, 1 interferer, 25dB SNR)

Figure 4.4: Limited interference suppression due to IQ mismatch in frequency selective channels as a function of sub-band size (Channel C)

mismatch. Moreover, this saturation occurs earlier for smaller FFT sizes, and therefore larger sub-bands. This is expected, since the spreading of the eigen modes due to multipath limits the interference suppression. Moreover, IQ mismatch causes a larger degradation in the interference suppression ability of the system with larger FFT sizes, and therefore smaller sub-bands. As the sub-bands become smaller, the eigen mode spread due to multipath is suppressed more so that limit on interference suppression ability now comes from IQ mismatch.

Fig. 4.5a indicates the additional SINR gain obtained by using the IQ mismatch compensation architecture proposed in this chapter. As one might observe, it is possible to mitigate the effect of the IQ mismatch almost completely.

Fig. 4.5b illustrates the packet error rate degradation due to IQ mismatch and the gain due to compensation. The IQ mismatch case in Fig. 4.5b indicates the PER when $\mathbf{W}_{\text{SMI}}$ is used as the frontend filter, and the MIMO decoder for the receiver is based on the IQ mismatch compensation scheme presented in [28]. We see that the performance degrades despite IQ mismatch compensation at the decoder. This reinforces the need for IQ mismatch compensation in the interference suppression architecture in order to ensure packet detection and synchronization functionality. The compensation case in Fig. 4.5b indicates the PER when the interference suppression architecture presented in this chapter is used for the frontend filter along with the MIMO decoder from [28]. It is again worth noting here that this MIMO decoder does give rise to a performance loss, depicted by the gap between the curves with compensation and with no IQ mismatch. This is because we now have to estimate 4x more channel coefficients, and therefore there is a larger estimation error. Nevertheless, the performance gain due to IQ mismatch compensation is still around 12dB at 10% PER.

Fig. 4.6 illustrates the contribution of the Packet Detection Error Rate (PDER) to the Packet Error Rate. We see here that IQ mismatch causes a degradation to the PDER, which again can be undone by using the interference suppression architecture proposed in this chapter.

Fig. 4.7 depicts the performance degradation due to IQ mismatch when the system is simulated using the Channel D model from the 802.11n standard [30]. This channel has a rms delay spread of 40 ns. We observe that as the frequency selectivity of the channel increases, the

(a) SINR Gain vs SIR for an 802.11n compliant system using Sample Matrix Inversion based interference suppression (MCS 10, 1 interferer, 25dB SNR)

(b) Packet Error Rate vs SIR for an 802.11n compliant system using Sample Matrix Inversion based interference suppression (MCS 10, 1 interferer, 25dB SNR)

Figure 4.5: IQ mismatch compensation in frequency selective channels (Channel C)

saturation in SINR Gain (Fig. 4.7a) happens sooner, and the saturation value is also lower than that of Channel C. This is due to more spreading of eigen modes in Channel D for the same sub-banding size. The trend in the PER (Fig. 4.7b) is also similar. The IQ mismatch compensation architecture successfully mitigates the effect of IQ mismatch to a large extent in this case as well. However, since the degradation due to IQ mismatch was not as large as Channel C, the

Figure 4.6: Packet Detection Error Rate vs SIR for an 802.11n compliant system using Sample Matrix Inversion based interference suppression (MCS 10, 1 interferer, 25dB SNR, Channel C)

gain due to IQ mismatch compensation is around 7dB at 10% PER.

On real hardware, a more evident metric is the throughput of the system. This is a linear function of the Packet Error Rate, and is computed as $Th = (1 - PER) * Th_{max}$, where the maximum throughput Th_{max} is defined by the 802.11n standard [29]. For MCS 10, this value is 39 Mbps. Fig. 4.8 depicts the throughput achievable by the interference suppression system in Channel C and Channel D. IQ mismatch limits the interference suppression ability by 15 - 18 dB in Channel C, while the IQ mismatch compensation scheme presented in this chapter provides a gain of 12 - 15 dB. For Channel D, IQ mismatch limits the interference suppression ability by 11 - 15 dB, while IQ mismatch compensation can suppress an additional 7 - 13 dB of interference.

4.5 Concluding Remarks

This chapter presented an IQ mismatch compensation architecture for interference suppression in wideband systems with frequency selective channels. Since wideband systems typically perform interference suppression after sub-banding the signal, the IQ mismatch compensation scheme is applied in the frequency domain. Simulation results were shown that demonstrate 12 - 15 dB additional interference suppression in indoor environments due to this architecture.

(a) SINR Gain vs SIR for an 802.11n compliant system using Sample Matrix Inversion based interference suppression (MCS 10, 1 interferer, 25dB SNR)

(b) Packet Error Rate vs SIR for an 802.11n compliant system using Sample Matrix Inversion based interference suppression (MCS 10, 1 interferer, 25dB SNR)

Figure 4.7: IQ mismatch compensation in frequency selective channels (Channel D)

(a) Throughput achievable by an 802.11n compliant system using Sample Matrix Inversion based interference suppression (MCS 10, 1 interferer, 25dB SNR, Channel C)

(b) Throughput achievable by an 802.11n compliant system using Sample Matrix Inversion based interference suppression (MCS 10, 1 interferer, 25dB SNR, Channel D)

Figure 4.8: Additional interference suppression due to IQ mismatch compensation in wideband systems

CHAPTER 5

Conclusion

Narrowband interference suppression systems can, in principle, completely suppress as many interferers as the number of available degrees of freedom at the receiver. However, in practical systems that have implementation impairments, IQ mismatch limits this interference suppression ability due to additional interference created from the image band of the interference signal. For a packet based system, limited interference suppression due to IQ mismatch can affect packet detection and synchronization functionality even if IQ mismatch is compensated for at the decoder stage. Therefore, it is essential to compensate for IQ mismatch at the frontend before packet detection, and an interference suppression architecture is presented that compensates for IQ mismatch using the concept of expanding the received signal sub-space.

Wideband interference suppression systems, on the other hand, are limited by the frequency selectivity of the channel. IQ mismatch further limits the interference suppression ability of a wideband system, and can be compensated for using the wideband interference suppression architecture presented. This architecture combines the expanded sub-space concept with the sub-banding typically present in wideband interference suppression systems. In indoor environments, the additional interference suppression due to this architecture is shown to be of the order of 12-15 dB.

The approach of expanding the received signal sub-space to eliminate interference from the image band can also be extended to expanding the sub-space over time to counter interference due to multipath. This is the idea behind space time processing [9]. If the taps of the filter in the time domain match the power delay profile of the channel, it might be possible to mitigate the effect of multipath to a large extent. However, the complexity associated with such a scheme might be restrictive for channels with large delay spread, and needs to be investigated further.

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