FULL-DUPLEX SELF-INTERFERENCE MITIGATION ANALYSIS FOR DIRECT CONVERSION RF NONLINEAR MIMO CHANNEL MODELS WITH IQ MISMATCH

Daniel W. Bliss, IEEE Fellow

Center of Wireless Information Systems and Computational Architectures
School of Electrical, Computer, and Energy Engineering
Arizona State University

ABSTRACT

In this paper, the self-interference mitigation performance of in-band full-duplex multiple-input multiple-output (MIMO) nodes is considered in the context of models for realistic hardware characteristics in which antennas are reused to transmit and receive simultaneously. The use of MIMO indicates a self-interference channel with spatially diverse inputs and outputs. Consequently, there are both MIMO channels for self interference and MIMO channels between the intended transmit array and receive array. Furthermore, physical transceivers suffer from nonlinearities and other nonidealities including IQ mismatch associated with direct conversion RF. Approaches to address self-interference mitigation under this model are presented and performances are detailed.

***Index Terms*—** Full-Duplex, MIMO, Wireless, Communications, Nonlinearity, IQ-Mismatch

1. INTRODUCTION

In this discussion, we assume that full-duplex communications indicates that on a given radio the allocated bandwidth is being used simultaneously to transmit and receive [1]. Furthermore, for this discussion, we also assume that the same antenna is being used to transmit and receive, although portions of this analysis can be extended for radios with isolated transmit and receive antennas.

1.1. Contribution

The primary contribution of this paper is the investigation of the formulation and effects of nonlinearity on the self-interference multiple-input multiple-output (MIMO) channel under the assumption of soft nonlinearities and IQ mismatch. In general, self-interference mitigation is a special case of adaptive interference mitigation [2]. We develop a simple formulation for temporal mitigation [3], expressed as a projection operator, when in the presence of nonlinearities in both the transmit and receive chains. Specifically, in this effort we extend an earlier investigation in which we assumed that IQ mismatch could be ignored [4].

1.2. Background

There is a significant and quickly growing literature on in-band full-duplex self-interference mitigation (for example, see the special issue [5]). Both MIMO channels and nonlinear effects have been considered previously. We considered MIMO self-interference channels in [3], channel estimation errors in [6], and nonlinear effects on MIMO self-interference adaptive transmit protection in [7, 4].

Given the numerous efforts in this field, it would be unreasonable to attempt to identify all of the significant contributions here. We will, however, attempt to identify a representative set of papers that have connection to our efforts. Experimental full-duplex demonstrations have been discussed [3, 8, 9]. Motivation for and limitations of full-duplex MIMO radios have been considered [10, 11, 12, 13]. The effects of hard nonlinearities on MIMO full-duplex systems, such as those introduced by analog-to-digital converters, have been investigated [14]. The effects of nonlinearities on full-duplex communications, in more generality, have been considered [15, 7, 16, 17, 18, 4].

2. MODEL

We consider a symmetric MIMO full-duplex system with N antennas that are used to transmit and receive simultaneously in the same band as depicted in Figure 1. For this analysis, we assume that the channel is spectrally flat. This assumption is clearly unrealistic; we plan to address this in future efforts. In this analysis, we include the effects of IQ mismatch that is typically present in direct conversion radios.

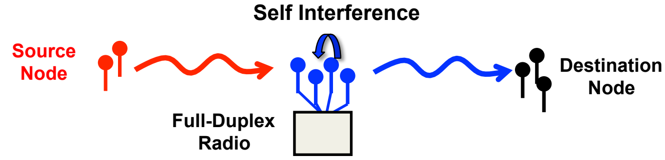

Fig. 1. MIMO full-duplex radio example.

2.1. Nonidealities

In general, many issues limit the performance of self-interference mitigation. These many issues include receive noise, transmit noise, phase noise, channel estimation errors, dispersion, and nonlinearities. Here we focus on a simplified model that ignores all adverse effects except *soft* nonlinearities and IQ mismatch.

We focus on soft nonlinearities in this work, which are associated with low order terms in the perturbative expansion of the system. In the absence of memory effects, this simplifies to a Taylor series representation. Because of typical system symmetries, the dominant nonlinearity is often the 3^{rd}-order term, and we limit the expansion to this order in this effort.

In modern radio systems, direct conversion is often employed to move between complex baseband and passband representations, as displayed for a receive channel in Figure 2. Consequently, there is a separate channel between the in-phase and quadrature channels, as displayed in the simplified scalar example representation for the complex baseband signal $z(t)$ given by

$$z(t) = \Re\{h\, x(t)\} + i\,\Im\{h\, x(t)\}\,(1+\varepsilon) - \frac{\alpha_{\Re}}{\sigma_{\rm rx}^2}(\Re\{h\, x(t)\})^3 - i\,\frac{\alpha_{\Im}}{\sigma_{\rm rx}^2}(\Im\{h\, x(t)\})^3 - \mathcal{O}(h\, x(t))^5 + n(t)\,, \qquad (1)$$

where h is the channel, and $x(t)$ is the transmitted signal as a function of time t. The terms $\alpha_{\Re}$ (and its imaginary counterparts) are the normalized unitless coefficients of the 3^{rd}-order nonlinearity. The term $\sigma_{\rm rx}^2$ is the average receive power. Under the assumption of small IQ mismatch, we encoded this effect in the complex variable ε. The real part of ε produces scale mismatch and the imaginary component breaks IQ orthogonality. Because the effect is small, we assume product of IQ mismatch and nonlinearity are negligible.

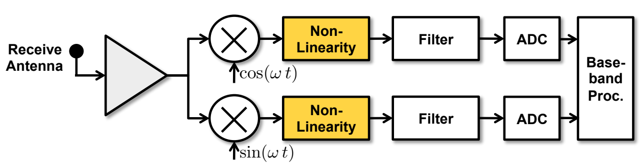

Fig. 2. Simplified example of the receive chain with independent nonlinearity for the in-phase and quadrature channels.

2.2. Direct Conversion RF MIMO Nonlinear Model

We present a simple model of the MIMO $N \times N$ antenna-reuse full-duplex radio. It is common for these systems to assume a circulator or similar technology to reduce direct transmit-to-receive chain signal strength. While this isolation is helpful, it is far from sufficient. We consider a simple model of transmit nonidealities caused by direct conversion. We assume a normalized transmit sequence $\mathbf{s}(t) \in \mathbb{C}^{N\times 1}$, such that $\langle \|\mathbf{s}(t)\|^2 \rangle = N$. The intended transmitted signal is given by $\mathbf{x}(t) = \sigma_{\rm tx}\, \mathbf{s}(t)/\sqrt{N}$, where $\sigma_{\rm tx}^2$ is the total transmit power across the multiple antennas. We assume for this analysis that the nonlinearity coefficients ($\boldsymbol{\alpha}$ for transmit and $\boldsymbol{\beta}$ for receive) are the same for the real and imaginary components. The effect of the transmit nonlinearity represented at complex baseband is given by

$$\mathbf{x}_{\rm NL}(t) = \mathbf{x}(t) + i\,\Im\{\mathbf{x}(t)\} \odot \boldsymbol{\varepsilon}_{\rm tx} - \frac{1}{\sigma_{\rm tx}^2}\boldsymbol{\alpha} \odot \left(\Re\{\mathbf{x}(t)\}^{(3)} + i\,\Im\{\mathbf{x}(t)\}^{(3)}\right), \qquad (2)$$

where $\cdot\odot\cdot$ indicates Hadamard product, and $\cdot^{(3)}$ indicates raising each of the vector or matrix elements to the third power. If we consider the linear (in $x(t)$) term of the transmit signal, one notices that real and imaginary components of $x(t)$ effectively see different channels because of the IQ mismatch. We have used the real channel as the reference channel in Equation (2). We can rewrite the linear component as having two different complex channels associated with the real and imaginary components $\mathbf{H}^{\Re}$ and $\mathbf{H}^{\Im} \in \mathbb{C}^{N\times N}$ of the transmit up-conversion chain, given by

$$\mathbf{H}\,\mathbf{x}(t) + i\,\mathbf{H}\,\Im\{\mathbf{x}(t)\} \odot \boldsymbol{\varepsilon}_{\rm tx} = \mathbf{H}^{\Re}\,\Re\{\mathbf{x}(t)\} + i\,\mathbf{H}^{\Im}\,\Im\{\mathbf{x}(t)\}\,. \qquad (3)$$

The effective channel is further modified by the IQ mismatch of the receive direct conversion RF chain. Consequently, in processing and channel estimation we must attempt to incorporate this effect.

The effects of direct conversion receive nonlinearities at complex baseband $\mathbf{z}(t) \in \mathbb{C}^{N\times 1}$ is given by

$$\mathbf{z}(t) = \mathbf{H}\,\mathbf{x}_{\rm NL}(t) + i\,\Im\{\mathbf{H}\,\mathbf{x}_{\rm NL}(t)\} \odot \boldsymbol{\varepsilon}_{\rm rx} + \mathbf{n}(t) - \frac{1}{\sigma_{\rm rx}^2}\boldsymbol{\beta} \odot \left(\Re\{\mathbf{H}\,\mathbf{x}_{\rm NL}(t)\}^{(3)} + i\,\Im\{\mathbf{H}\,\mathbf{x}_{\rm NL}(t)\}^{(3)}\right), \qquad (4)$$

where $\mathbf{H} \in \mathbb{C}^{N\times N}$ is the self-interference channel representing the complex propagation attenuation between each antenna and the circulators for the same antenna, and $\mathbf{n}(t) \in \mathbb{C}^{N\times 1}$ is the thermal noise. Expanding the self-interference signal in powers of nonlinearities, the complex baseband signal is given by

$$\begin{aligned}\mathbf{z}(t) \approx\ & \mathbf{H}\,\mathbf{x}(t) + i\,\mathbf{H}\,\Im\{\mathbf{x}(t)\} \odot \boldsymbol{\varepsilon}_{\rm tx} + i\,\Im\{\mathbf{H}\,\mathbf{x}(t)\} \odot \boldsymbol{\varepsilon}_{\rm rx} \\ & - \frac{1}{\sigma_{\rm tx}^2}\mathbf{H}\,\boldsymbol{\alpha} \odot \left(\Re\{\mathbf{x}(t)\}^{(3)} + i\,\Im\{\mathbf{x}(t)\}^{(3)}\right) \\ & - \frac{1}{\sigma_{\rm rx}^2}\boldsymbol{\beta} \odot \left(\Re\{\mathbf{H}\,\mathbf{x}(t)\}^{(3)} + i\,\Im\{\mathbf{H}\,\mathbf{x}(t)\}^{(3)}\right) + \mathbf{n}(t)\,.\end{aligned} \qquad (5)$$

2.3. Matrix Representation of Sampled Signals

At this point, it will be useful to employ matrix representations of sampled data. As an example, the intended transmit signal is represented by

$$\mathbf{S} = (\mathbf{s}(T0) \quad \mathbf{s}(T1) \cdots \mathbf{s}(T[n_s-1])) \in \mathbb{C}^{N\times n_s}\,, \qquad (6)$$

where T is the sample period and n_s is the number of samples. Similarly, $\mathbf{X} \in \mathbb{C}^{N \times n_s}$ is constructed from $\mathbf{x}(t)$, and $\mathbf{Z} \in \mathbb{C}^{N \times n_s}$ is constructed from $\mathbf{z}(t)$.

3. IQ-SPACE SELF-INTERFERENCE MITIGATION

To address the issue of mismatched IQ channels one can construct an interference mitigation approach that increases the dimensionality by introducing a real space of parameters. To mitigate the effect of the self interference including IQ mismatch and low-order nonlinearities we employ the temporal projection operation given by

$$\begin{pmatrix} \Re\{\check{\mathbf{Z}}\} \\ \Im\{\check{\mathbf{Z}}\} \end{pmatrix} = \begin{pmatrix} \Re\{\mathbf{Z}\} \\ \Im\{\mathbf{Z}\} \end{pmatrix} \mathbf{P}^{\perp}_{\check{\mathbf{S}}} \tag{7}$$

$$\mathbf{P}^{\perp}_{\check{\mathbf{S}}} = \mathbf{I} - \check{\mathbf{S}}^{\dagger} \, (\check{\mathbf{S}} \, \check{\mathbf{S}}^{\dagger})^{-1} \, \check{\mathbf{S}} \,, \tag{8}$$

where the temporal space is now spanned by

$$\check{\mathbf{S}} = \begin{pmatrix} \Re\{\mathbf{S}\} \\ \Im\{\mathbf{S}\} \\ (\Re\{\mathbf{S}\})^{(3)} \\ (\Im\{\mathbf{S}\})^{(3)} \\ (\Re\{[\mathbf{H}^{\Re} \, \Re\{\mathbf{S}\} + i \, \mathbf{H}^{\Im} \, \Im\{\mathbf{S}\}]\})^{(3)} \\ (\Im\{[\mathbf{H}^{\Re} \, \Re\{\mathbf{S}\} + i \, \mathbf{H}^{\Im} \, \Im\{\mathbf{S}\}]\})^{(3)} \end{pmatrix} \in \mathbb{R}^{(6N) \times n_s} \,. \tag{9}$$

To perform this operation, we need to have an accurate estimate of the self-interference channel. Of course, we do not know the channel $\mathbf{H}$.

3.1. MIMO Self-Interference Channel Estimation

Because the self-interference level is very high, the self-interference channel can be estimated well, in theory. However, the effects of nonlinearities can adversely affect the estimate if not addressed. Furthermore, the effects of IQ mismatch introduce a significant challenge, which are only partially addressed in this work. The standard linear least squares MIMO channel estimator [2] is given by

$$\hat{\mathbf{H}} = \mathbf{Z} \, \mathbf{X}^{\dagger} \, (\mathbf{X} \, \mathbf{X}^{\dagger})^{-1} = \sqrt{\frac{N}{\sigma^2_{\text{tx}}}} \, \mathbf{Z} \, \mathbf{S}^{\dagger} \, (\mathbf{S} \, \mathbf{S}^{\dagger})^{-1} \tag{10}$$

One could employ a nonlinear maximum-likelihood approach to jointly estimate all nonlinear coefficients explicitly in addition to the channel, but this approach is somewhat complicated. To reduce estimation complexity, it is convenient to use an estimator that is invariant to the nonlinearities. This channel estimator projects onto a basis (for both data and reference) that is orthogonal to the nonlinearities. Unfortunately, this requires knowledge of the channel. This issue can be circumvented by employing an iterative estimator developed by extending multiuser detection concepts in [2]. We initialize the orthogonal spanning matrices as

$$\mathbf{A}^{\Re}_{(1)} = \begin{pmatrix} \Im\{\mathbf{S}\} \\ (\Re\{\mathbf{S}\})^{(3)} \\ (\Im\{\mathbf{S}\})^{(3)} \end{pmatrix} , \quad \mathbf{A}^{\Im}_{(1)} = \begin{pmatrix} \Re\{\mathbf{S}\} \\ (\Re\{\mathbf{S}\})^{(3)} \\ (\Im\{\mathbf{S}\})^{(3)} \end{pmatrix} . \tag{11}$$

We then iterate with the following procedure,

repeat

$$\mathbf{P}^{\Re}_{(m)} = \mathbf{A}^{\Re\,\dagger}_{(m)} \, [\mathbf{A}^{\Re}_{(m)} \, \mathbf{A}^{\Re\,\dagger}_{(m)}]^{-1} \mathbf{A}^{\Re}_{(m)} \tag{12}$$

$$\mathbf{P}^{\Im}_{(m)} = \mathbf{A}^{\Im\,\dagger}_{(m)} \, [\mathbf{A}^{\Im}_{(m)} \, \mathbf{A}^{\Im\,\dagger}_{(m)}]^{-1} \mathbf{A}^{\Im}_{(m)} \tag{13}$$

$$\tilde{\mathbf{S}}^{\Re}_{(m)} = \Re\{\mathbf{S}\} \, (\mathbf{I} - \mathbf{P}^{\Re}_{(m)}) \tag{14}$$

$$\tilde{\mathbf{S}}^{\Im}_{(m)} = i \, \Im\{\mathbf{S}\} \, (\mathbf{I} - \mathbf{P}^{\Im}_{(m)}) \tag{15}$$

$$\hat{\mathbf{H}}^{\Re}_{(m)} = \sqrt{\frac{N}{\sigma^2_{\text{tx}}}} \, \mathbf{Z} \, (\tilde{\mathbf{S}}^{\Re}_{(m)})^{\dagger} \, [\tilde{\mathbf{S}}^{\Re}_{(m)} \, (\tilde{\mathbf{S}}^{\Re}_{(m)})^{\dagger}]^{-1} \tag{16}$$

$$\hat{\mathbf{H}}^{\Im}_{(m)} = \sqrt{\frac{N}{\sigma^2_{\text{tx}}}} \, \mathbf{Z} \, (\tilde{\mathbf{S}}^{\Im}_{(m)})^{\dagger} \, \tilde{\mathbf{S}}^{\Im}_{(m)} \, (\tilde{\mathbf{S}}^{\Im}_{(m)})^{\dagger}]^{-1} \tag{17}$$

$$\mathbf{B}_{(m)} = \hat{\mathbf{H}}^{\Re} \, \Re\{\mathbf{S}\} + i \, \hat{\mathbf{H}}^{\Im} \, \Im\{\mathbf{S}\}$$

$$\mathbf{A}^{\Re}_{(m+1)} = \begin{pmatrix} \Im\{\mathbf{S}\} \\ (\Re\{\mathbf{S}\})^{(3)} \\ (\Im\{\mathbf{S}\})^{(3)} \\ (\Re\{\mathbf{B}_{(m)}\})^{(3)} \\ (\Im\{\mathbf{B}_{(m)}\})^{(3)} \end{pmatrix} , \; \mathbf{A}^{\Im}_{(m+1)} = \begin{pmatrix} \Re\{\mathbf{S}\} \\ (\Re\{\mathbf{S}\})^{(3)} \\ (\Im\{\mathbf{S}\})^{(3)} \\ (\Re\{\mathbf{B}_{(m)}\})^{(3)} \\ (\Im\{\mathbf{B}_{(m)}\})^{(3)} \end{pmatrix}$$

until bored,

where the m^{th} estimate of the channel is given by $\hat{\mathbf{H}}^{\Re \text{ or } \Im}_{(m)}$, and the m^{th} estimate of the orthogonal spanning matrices is indicated by $\mathbf{A}^{\Re \text{ or } \Im}_{(m)}$. Given our relatively simple model, this estimate converges quickly as seen in Figure 3. If the IQ mismatch is large compared to the nonlinearity terms, then further iterations do not improve upon the initial estimate.

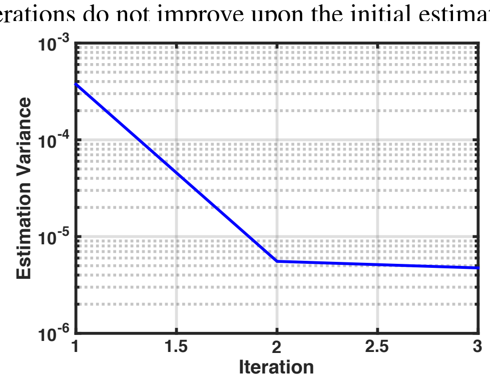

Fig. 3. Self-interference channel estimation variance for 4 antenna system with nonlinearities in the transmitter, $\boldsymbol{\alpha}$ of -30 dB, and receiver, $\boldsymbol{\beta}$ of -30 dB, and IQ-mismatch variance of -60 dB as a function of iteration, assuming 130 dB self-interference. Assumed self-interference channel $\mathbf{H}$ element variance is -20 dB.

3.2. Estimation and Subtraction

The projection operation can be interpreted as estimation and subtraction,

$$\begin{pmatrix} \Re\{\check{\mathbf{Z}}\} \\ \Im\{\check{\mathbf{Z}}\} \end{pmatrix} = \begin{pmatrix} \Re\{\mathbf{Z}\} \\ \Im\{\mathbf{Z}\} \end{pmatrix} - \underbrace{\begin{pmatrix} \Re\{\mathbf{Z}\} \\ \Im\{\mathbf{Z}\} \end{pmatrix} \check{\mathbf{S}}^{\dagger} \, (\check{\mathbf{S}}\, \check{\mathbf{S}}^{\dagger})^{-1}}_{\hat{\check{\mathbf{H}}}} \, \check{\mathbf{S}} \,, \tag{18}$$

where $\hat{\check{\mathbf{H}}} \in \mathbb{R}^{(2N)\times(6N)}$ is an estimate of the combined linear and nonlinear channel as seen at complex baseband. In this form, we observe that the nonlinear channel could be estimated at one point in time and then applied at a later time. This opens the possibility of using this approach to mitigate the self-interference including nonlinearities at passband, which would be desirable.

4. PERFORMANCE EXAMPLES

In the following plots we present the residual power to noise ratio under the assumption of temporal self-interference mitigation limited to nonlinear transmit and receive 3^{rd}-order terms, for a system with 4 antennas, transmit-to-receive-noise ratio of 130 dB, i.i.d. complex Gaussian self-interference channel with average attenuation of 20 dB, and i.i.d. complex Gaussian signals with 100 samples. The residual power is averaged over an ensemble of 1,000 channel, IQ mismatch, signal, and noise realizations. In Figures 4 and 5 the processing assumes symmetric complex processing with no IQ mismatch. This assumption is valid for Figure 4, but fails badly in Figure 5 for which each element of the i.i.d. circularly symmetric complex Gaussian IQ match parameters $\varepsilon_{\mathrm{tx}}$ and $\varepsilon_{\mathrm{rx}}$ has variance -60 dB. In Figure 6, we use the approach developed within this paper. Despite the IQ mismatch, relatively good self mitigation performance is achieved. It is worth noting that when the IQ mismatch is larger, channel estimation performance suffers, and consequently the self-interference mitigation performance suffers.

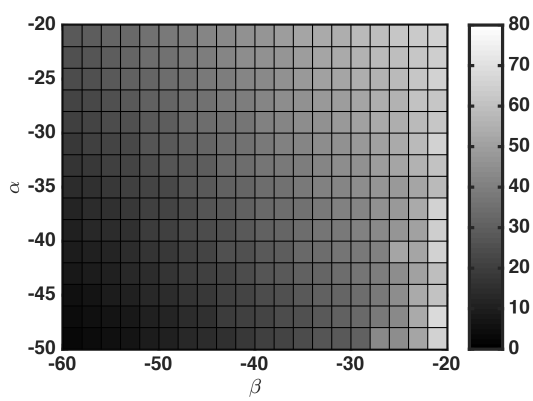

Fig. 4. Residual power to noise ratio (dB) after complex symmetric nonlinear mitigation approach correctly assuming IQ match, with $\varepsilon_{\mathrm{tx}} = \varepsilon_{\mathrm{rx}} = 0$.

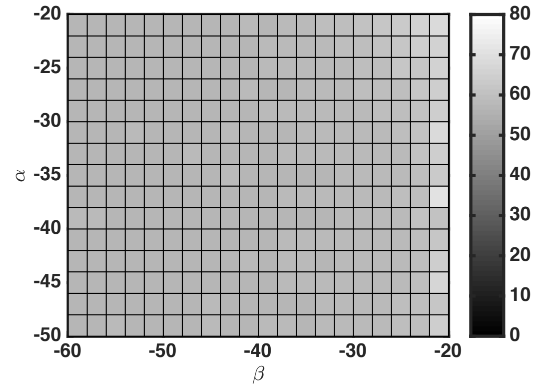

Fig. 5. Residual power to noise ratio (dB) after complex symmetric nonlinear mitigation approach incorrectly assuming IQ match, with $\varepsilon_{\mathrm{tx}}$ and $\varepsilon_{\mathrm{rx}}$ variance -60 dB.

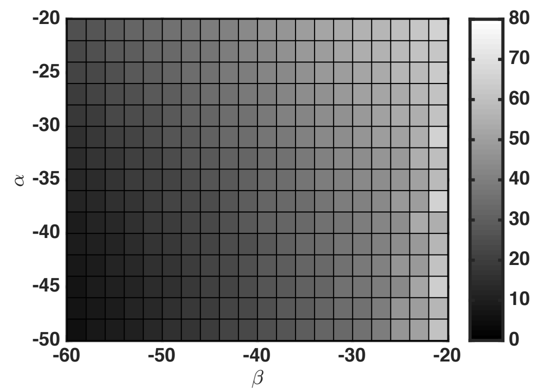

Fig. 6. Residual power to noise ratio (dB) employing nonlinear mitigation approach introduced in this paper, with $\varepsilon_{\mathrm{tx}}$ and $\varepsilon_{\mathrm{rx}}$ variance -60 dB.

5. CONCLUSION

We have developed a simple formalism for temporal mitigation of self-interference for in-band full-duplex MIMO radios with antenna reuse. We considered a reduced complexity problem formalism by assuming flat-fading channels. We introduced an invariance motivated channel estimation approach and provided simulated performance results as a function of the nonlinear contributions that incorporated the effects of IQ mismatch. With useful parameter sets, we can mitigate the self-interference to near the noise floor; however, channel estimation, and thus mitigation performance, is challenged in the large IQ-mismatch regime.

We would like to thank Richard Gutierrez, Bryan Paul, Andrew Herschfelt, and Yu Rong of ASU for their comments.

6. REFERENCES

[1] A. Sabharwal, P. Schniter, Dongning Guo, D.W. Bliss, S. Rangarajan, and R. Wichman, "In-band full-duplex wireless: Challenges and opportunities," *IEEE Journal on Selected Areas in Communications*, vol. 32, no. 9, pp. 1637–1652, Sept. 2014.

[2] Daniel W. Bliss and Siddhartan Govindasamy, *Adaptive Wireless Communications: MIMO Channels and Networks*, Cambridge University Press, Cambridge, 2013.

[3] D. W. Bliss, P. A. Parker, and A. R. Margetts, "Simultaneous transmission and reception for improved wireless network performance," *Conference Proceedings of the IEEE Statistical Signal Processing Workshop*, Aug. 2007.

[4] D. W. Bliss and Y. Rong, "Full-duplex self-interference mitigation performance in nonlinear channels," in *IEEE Asilomar Conference on Signals, Systems and Computers*, Nov 2014, pp. 1696–1700.

[5] A. Sabharwal, P. Schniter, D. Guo, D.W. Bliss, S. Rangarajan, and R. Wichman, "Guest editorial: In-band full-duplex wireless communications and networks," *IEEE Journal on Selected Areas in Communications*, vol. 32, no. 9, pp. 1633–1636, Sept. 2014.

[6] D. W. Bliss and Y. Rong, "Effects of channel estimation errors on in-band full-duplex MIMO radios using adaptive transmit spatial mitigation," *Conference Record of the IEEE Asilomar Conference on Signals, Systems & Computers*, Nov. 2013.

[7] D. W. Bliss, T. M. Hancock, and P. Schniter, "Hardware phenomenological effects on cochannel full-duplex MIMO relay performance," *Conference Record of the IEEE Asilomar Conference on Signals, Systems & Computers*, Nov. 2012.

[8] M. Duarte and A. Sabharwal, "Full-duplex wireless communications using off-the-shelf radios: Feasibility and first results," in *IEEE Asilomar Conference on Signals, Systems and Computers*, Nov. 2010, pp. 1558–1562.

[9] J. I. Choi, M. Jain, K. Srinivasan, P. Levis, and S. Katti, "Achieving single channel, full duplex wireless communications," in *Proceedings of ACM Mobicom*, 2010.

[10] Y. Y. Kang and J. H. Cho, "Capacity of MIMO wireless channel with full-duplex amplify-and-forward relay," in *IEEE 20th International Symposium on Personal, Indoor and Mobile Radio Communications*, Sept. 2009, pp. 117–121.

[11] S. Simoens, O. Munoz-Medina, J. Vidal, and A. del Coso, "On the Gaussian MIMO relay channel with full channel state information," *IEEE Transactions on Signal Processing*, vol. 57, no. 9, pp. 3588–3599, Sept. 2009.

[12] Hyungsik Ju, Eunsung Oh, and Daesik Hong, "Improving efficiency of resource usage in two-hop full duplex relay systems based on resource sharing and interference cancellation," *IEEE Transactions on Wireless Communications*, vol. 8, no. 8, pp. 3933–3938, Aug. 2009.

[13] Chinmay S Vaze and Mahesh K Varanasi, "The degrees of freedom of MIMO networks with full-duplex receiver cooperation but no CSIT," *arXiv preprint arXiv:1209.1291*, 2012.

[14] B. P. Day, A. R. Margetts, D. W. Bliss, and P. Schniter, "Full-duplex bidirectional MIMO: Achievable rates under limited dynamic range," *IEEE Transactions on Signal Processing*, vol. 60, no. 7, pp. 3702–3713, July 2012.

[15] M. Omer, R. Rimini, P. Heidmann, and J.S. Kenney, "A compensation scheme to allow full duplex operation in the presence of highly nonlinear microwave components for 4G systems," in *Microwave Symposium Digest (MTT), 2011 IEEE MTT-S International*, June 2011, pp. 1–4.

[16] Elsayed Ahmed, Ahmed M. Eltawil, and Ashutosh Sabharwal, "Self-interference cancellation with nonlinear distortion suppression for full-duplex systems.," *CoRR*, vol. abs/1307.3796, 2013.

[17] A. Kiayani, L. Anttila, and M. Valkama, "Digital suppression of power amplifier spurious emissions at receiver band in FDD transceivers," *Signal Processing Letters, IEEE*, vol. 21, no. 1, pp. 69–73, Jan. 2014.

[18] D. Korpi, T. Riihonen, V. Syrjala, L. Anttila, M. Valkama, and R. Wichman, "Full-duplex transceiver system calculations: Analysis of ADC and linearity challenges," *IEEE Transactions on Wireless Communications*, vol. 13, no. 7, pp. 3821–3836, July 2014.