An Approach for Compensating Reciprocal Mixing and Close-In Phase Noise Distortion

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Abstract — Local oscillator phase noise is one of the primary factors that limits the achievable performance in modern communication receivers. There are two mechanisms by which the phase noise degrades the desired signal – the close-in and reciprocal mixing distortion. Although an extensive body of literature exists in compensating for close-in phase noise distortion, there is very little work on compensating for the effects of reciprocal mixing with no practical solution to date. This paper presents an approach to jointly and seamlessly compensate for both the close-in and reciprocal mixing phase noise distortion. A prototype receiver using commercial off-the-shelf components has been built and a 9.4 dB SINR improvement demonstrated.

Keywords — receivers, phase noise, reciprocal mixing, equalizer

I. INTRODUCTION

Local oscillator (LO) phase noise is a major impairment that degrades the performance of communication receivers. There are two mechanisms by which the phase noise distorts the desired signal. In the first, the LO mixing introduces random phase modulation to the desired signal. This phase noise distortion, which we subsequently refer to as close-in phase noise distortion, is especially problematic in high data rate communication systems that employ large constellation sizes. An extensive body of research exists in compensating for close-in phase noise distortion, most of which rely on the presence of pilot signals [1], [2], [3].

When the received signal is accompanied by a large interferer, the desired signal not only suffers from close-in phase noise distortion but also reciprocal mixing distortion, which is the spreading of the interfering spectrum by the LO phase noise to the desired signal band. Compared to close-in phase noise distortion, there is very little existing work on compensating for the effects of reciprocal mixing, primarily because of its difficulty. In [4], [5], which represent the only prior work we are aware of, reciprocal mixing induced distortion compensation is achieved by exploiting the symmetrical properties of the phase noise spectrum by subtracting the signal at twice the blocker offset, which the authors refer to as the phase noise image signal. The practical shortcomings of this approach are the assumption that phase noise image band is unoccupied, which is generally not the case, and the difficulty of identifying and downconverting the image band, especially if multiple blockers are present. Consequently, this approach is quite limited in its applicability, and to our knowledge, no practical solution to compensating the effects of reciprocal mixing exists to date.

II. PROPOSED PHASE NOISE DISTORTION COMPENSATION

The core idea of the proposed compensation approach is to architect the receiver front-end with additional degrees of freedom so that the desired signal spans a different vector space than the subspace spanned by reciprocal mixing and close-in phase noise distortion. The phase noise induced distortions are then removed by simply projecting to their null space. The proposed approach requires processing of only the downconverted desired signal band samples, and it operates seamlessly regardless of the blocker spectrum, number of interferers, and phase noise spectrum.

Consider the received RF signal

\[ r_{\text{RF}}(t) = \Re\left\{ s(t) e^{j2\pi M f_{1F} t} + \sum_{i} I_i(t) e^{j2\pi f_i t} + n(t) \right\} \] (1)

where \( \Re\{\cdot\} \) represents the real operation, \( s(t) \) is the baseband equivalent desired signal that is bandlimited over the interval \([-f_s/2, f_s/2]\), \( f_{1F}(t) \) is the receiver IF, \( M \) is a scaling factor such that \( M f_{1F}(t) \) represents the desired signal carrier frequency, \( I_i(t) \) is the baseband equivalent ith interferer signal centered at frequency \( f_i \), and \( n(t) \) is the input-referred additive white Gaussian noise (AWGN).

A simplified block diagram of the proposed receiver is shown in Fig. 1. The received signal \( r_{\text{RF}}(t) \) is first downconverted to an IF at \( f_{1F} \) then to baseband. In the proposed front-end, one receive chain is a high-side and the other a low-side injection IF receiver. In particular, the LOs are given by

\[ m_1(t) = \Re\left\{ e^{j2\pi (M-1) f_{01} t + (M-1) \theta(t) + \phi_1} \right\} \] (2)

\[ m_2(t) = \Re\left\{ e^{j2\pi (M+1) f_{02} t + (M+1) \theta(t) + \phi_2} \right\} \] (3)

where \( \theta(t) \) is the phase noise, and \( \phi_1 \) and \( \phi_2 \) are the phase offsets of two LOs. For each LO, the phase noise is scaled...
linearly with frequency, i.e., the phase noise of the tone at frequency $(M - 1) f_b$ is $(M - 1) \theta(t)$ while the tone at $(M + 1) f_b$ is $(M+1) \theta(t)$. Such LOs can be generated using different frequency multipliers as described in the following section. This proportional phase noise scaling provides the necessary degrees of freedom to decouple the desired signal $s(t)$ from both reciprocal mixing and close-in phase noise distortion.

After mixing with $m_1(t)$ and $m_2(t)$, the resulting IF signals $r_w(t)$ and $r_b(t)$ are quadrature mixed by $q(t) = \exp\left[-j2\pi f_b t\right]$ to generate complex baseband signals (denoted by bold lines in Fig. 1) then lowpass filtered by $h(t)$, whose bandwidth of $f_b/2$ passes only the desired signal $s(t)$. In our overview description of the proposed receiver, we assume that the phase noise of the IF LO $q(t)$ is negligible compared to that of the RF LOs $m_1(t)$ and $m_2(t)$, since the phase noise power spectral density is proportional to the square of the carrier frequency. Additional power can also be expended to further reduce the phase noise of the IF LO, which is much easier to achieve than the RF LO because of the reduced operating frequency.

1) System Modeling and Baseband Processing

Applying $e^{j(M \pm 1)\theta(t)} \approx 1 + j(M \pm 1)\theta(t)$ since $|\theta(t)| \ll 1$ and ignoring additive noise terms, the complex baseband signal $b_1(t)$ (see Fig. 1) can be approximated as

$$b_1(t) \approx e^{-j\phi_1} \left[ s(t) - j(M - 1)v_s(t) - j(M - 1)v_1(t) \right]$$  \hspace{1cm} (4)

where $v_s(t)$ and $v_1(t)$ represent the close-in and reciprocal mixing distortion terms, respectively, i.e.,

$$v_s(t) = s(t)\theta(t) * h(t)$$

$$v_1(t) = \left( \sum_i I_i(t)e^{2\pi i(Mf_i-f_b)t} \right)\theta(t) * h(t).$$

Similarly, the approximation of $b_2(t)$ becomes

$$b_2(t) \approx e^{j\phi_2} \left[ s^*(t) + j(M + 1)v_s^*(t) + j(M + 1)v_1^*(t) \right]$$ \hspace{1cm} (5)

Note that the gain coefficients of $v_s(t)$ and $v_1(t)$ in (4) (as well as their conjugates $v_s^*(t)$ and $v_1^*(t)$ in (5)) are the same, suggesting that the desired signal distortion from close-in phase noise and reciprocal mixing are indistinguishable and can be combined as a single noise source.

To appreciate the benefit of performing both a low-side and high-side IF mixing with RF LOs $m_1(t)$ and $m_2(t)$, the baseband signals $b_1(t)$ and $b_2(t)$ can be represented in matrix form as

$$\begin{bmatrix} b_1(t) \\ b_2(t) \end{bmatrix} = \begin{bmatrix} e^{-j\phi_1} & -je^{-j\phi_1}(M - 1) \\ e^{-j\phi_2} & -je^{-j\phi_2}(M + 1) \end{bmatrix} \begin{bmatrix} s(t) \\ v(t) \end{bmatrix} + \begin{bmatrix} n_1(t) \\ n_2(t) \end{bmatrix}$$ \hspace{1cm} (6)

where

$$v(t) = v_s(t) + v_1(t)$$ \hspace{1cm} (7)

is the combined phase noise induced distortion term, and $n_1(t)$ and $n_2(t)$ represent the equivalent additive noise.

Denoting (6) as $b(t) = Ax(t) + n(t)$, the objective is to estimate $s(t)$ based on observations $b(t)$. As the matrix $A$ is full-rank, the desired signal $s(t)$ can be extracted from the corrupted baseband signals $b(t)$. More importantly, since the elements of matrix $A$ are functions only of static receiver parameters (i.e., $M, \phi_1, \phi_2$), the proposed approach is effective regardless of the phase noise or the interferer spectrum. Desired signal $s(t)$ can be readily estimated using well-known minimum mean-squared error (MMSE) or zero forcing (ZF) equalizers.

III. PROTOTYPE IMPLEMENTATION AND MEASUREMENT

To experimentally validate the proposed phase noise compensation approach, a prototype receiver is built using commercial off-the-shelf components. A block diagram of the receiver signal path is shown in Fig. 2a). The RF signal centered at 2 GHz (generated from a vector signal generator) is split using a power splitter. The two outputs are the inputs to the two receive paths. The top path is multiplied by a 1.6 GHz $m_1(t)$ signal then bandpass filtered at 400 MHz. The resulting output IF signal is amplified and quadrature downconverted to baseband for digitization using a real-time spectrum analyzer tuned at 400 MHz. The bottom path is the same as the top path except that the split RF signal is multiplied by a 2.4 GHz $m_2(t)$ signal.

The LO signals $m_1(t)$ and $m_2(t)$ with the same phase noise but scaled proportionally with the operating frequency
are obtained based on a reference 800 MHz signal generated from a signal generator. A noisy RF LO is modeled by adding a Gaussian noise to the signal generator output followed by a limiter as illustrated in Fig. 2b). The 800 MHz noisy signal is amplified then split using a power splitter. One output passes through a frequency doubler while the other a frequency tripler to generate the corresponding 1.6 GHz and 2.4 GHz LO tones, respectively. The phase noise of the resulting LO signals is proportional as given in (2) and (3) with $M = 5$ and $\omega_{IF} = 400$ MHz. The prototype receiver built using commercial off-the-shelf components is shown in Fig. 2c).

For ease of testing, a 16-QAM signal with a 10 kHz symbol rate is transmitted at 1.9999 GHz carrier frequency. The transmit filter is a square-root raised cosine filter with a roll-off factor of 0.35. To demonstrate the robustness of the proposed approach to the effects of phase noise, the input signal power is set to be sufficiently large at approximately $-40$ dBm so that the signal-dependent phase noise becomes the dominant source of distortion and the effects of additive white Gaussian noise is negligible.

The IF spectrum of the low-side injection path in the absence of a blocker is shown in Fig. 3a). The noise floor is primarily due to the mixing of the desired signal with LO phase noise. Fig. 3b) plots the decoded constellation, which corresponds to an output SINR of 27.0 dB. Although not shown, the output SINR of the high-side injection path is 24.9 dB. The decoded constellation of the proposed receiver is shown in Fig. 3c) and achieves an output SINR of 32.1 dB, which represents an improvement of 5.1 dB and 6.2 dB compared to the low-side and high-side injection IF receivers, respectively.

A single-tone blocker is injected at 1.9998 GHz, which corresponds to a frequency offset of 100 kHz relative to the desired signal centered at 1.9999 GHz. The IF signal spectrum of the low-side injection mixer is shown in Fig. 4a). Compared to Fig. 3a), the presence of the blocker increased the noise floor by approximately 20 dB. The decoded constellation of the low-side injection path is plotted in Fig. 4b) and corresponds to an output SINR of 15.7 dB. Although not shown, the output SINR of the high-side injection path is 12.3 dB. When the proposed receiver is employed, the constellation is shown in Fig. 4c), which corresponds to an output SINR of 25.1 dB. Compared to the low-side injection path alone, which represents the higher performance path, the proposed receiver improved the output SINR by 9.4 dB.

IV. CONCLUSION

A novel and practical approach to compensating for the effects of close-in and reciprocal mixing phase noise distortion compensation is demonstrated. The proposed approach seamlessly compensates for both reciprocal mixing and close-in phase noise distortion without knowledge of the blockers or the desired signal characteristics. Despite the simplicity of the proposed phase noise compensation approach, preliminary results suggest that the resulting performance improvements are significant.

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REFERENCES


