A NOVEL ADAPTIVE TECHNIQUE FOR DIGITAL I/Q IMBALANCE COMPENSATION IN OFDM RECEIVERS

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ABSTRACT

One of the most important impairment introduced by the analog Radio Frequency (RF) front-end in wireless digital communication receivers is the imbalance between the I and Q arms of a practical quadrature demodulator. In this paper we present a novel adaptive mixed time-frequency domain technique for digital I/Q imbalance compensation where the compensation is performed in time domain while the adaptation is performed in frequency domain. This method is suitable for Orthogonal Frequency Division Multiplexing (OFDM) receivers in the presence of Carrier Frequency Offset (CFO) and is applied during the receiver's normal operation without a need for an off-line calibration phase and training signals. Theoretical analysis and simulation results that prove the achievable performances and the robustness of the approach are provided.

1. INTRODUCTION

The performance of a wireless Orthogonal Frequency Division Multiplexing (OFDM) [1] receiver is dictated by the degradation introduced in the analog Radio Frequency (RF) front-end. One of the critical non-ideality is the gain and phase imbalance between the I and Q arms of a quadrature demodulator. These errors can vary considerably with temperature, channel frequency and with oscillator drive power [2]. The presence of the I/Q imbalance can result in a poor image frequency signal rejection, which degrades the received Signal to Noise Ratio (SNR). We propose a novel adaptive mixed time-frequency domain technique for digital I/Q imbalance compensation in a quadrature demodulator in wireless OFDM receivers. It is applicable to several RF architectures such as heterodyne, Low-IF and Zero-IF. The proposed technique performs compensation in the time domain, using a single-tap filter in a manner similar to the compensation structure of the simple interference canceller described in [3]. However in the proposed method, the adaptation of the single-tap filter coefficient is performed in the frequency domain using the Least Mean Square (LMS) algorithm. The mixed time-frequency domain approach has the advantage of robust performance in the presence of Carrier Frequency Offset (CFO). Moreover the adaptive technique ensures tracking capability of imbalance variations. We present the method in the context of a Zero-IF RF architecture for an OFDM receiver. A system model based on the Wireless Local Area Network (WLAN) IEEE 802.11a standard [4] that employs OFDM modulation is used to carry out all simulations.

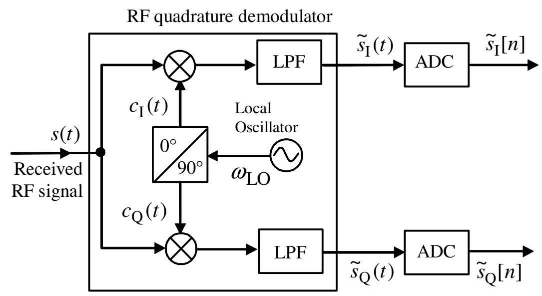

Fig. 1. Zero-IF quadrature RF receiver.

2. RECEIVER MODEL

2.1. I/Q imbalance model

A Zero-IF receiver architecture is presented in Fig. 1. The received RF signal (transmitted signal convolved with the channel) can be represented as

$$s(t) = \Re\left\{(s_I(t) + js_Q(t))e^{j\omega_0 t}\right\} = \Re\left\{\underline{s}(t)e^{j\omega_0 t}\right\} \quad (1)$$

where ω_0 is the transmit RF carrier, $s_I(t)$ and $s_Q(t)$ the digital base band quadrature signals and $\underline{s}(t) = s_I(t) + js_Q(t)$ the complex envelope. Hereafter an undersigned variable e.g. $\underline{x}$ represents the complex envelope of a time domain signal.

I/Q imbalance results from the component mismatches in the two arms of a quadrature receiver. This imbalance is distributed in the analog receiver over several blocks including mixers, low-pass filters (LPF) and analog to digital converters (ADC). For signal analysis, without loss of generality, we confine the imbalance to the Local Oscillator (LO) quadrature signals $c_I(t)$ and $c_Q(t)$. Assuming a symmetrically distributed imbalance and denoting ϵ as the

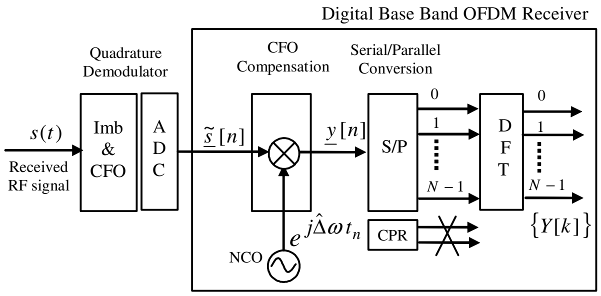

Fig. 2. Simplified OFDM receiver.

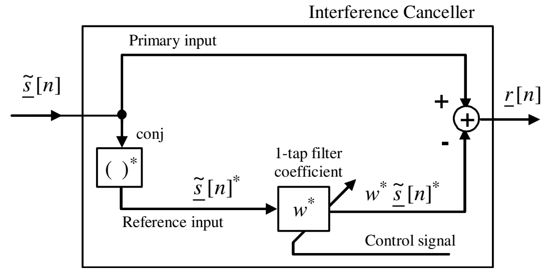

Fig. 3. Adaptive interference cancellation structure.

gain imbalance, $\Delta\phi$ the phase imbalance and $\Delta\omega$ the CFO in radians, the LO quadrature signals are

$$\begin{aligned} c_{\mathrm{I}}(t) &= 2(1-\epsilon/2)\cos((\omega_0+\Delta\omega)t-\Delta\phi/2) \\ c_{\mathrm{Q}}(t) &= -2(1+\epsilon/2)\cos((\omega_0+\Delta\omega)t+\Delta\phi/2) \end{aligned} \quad (2)$$

Assuming ideal filtering and no additive noise the demodulated signal can be represented as

$$\widetilde{\underline{s}}(t) = \mu\underline{s}'(t) + \nu\underline{s}'(t)^* \quad (3)$$

where

$$\begin{aligned} \mu &= \cos(\Delta\phi/2) - j(\epsilon/2)\sin(\Delta\phi/2) \\ \nu &= -(\epsilon/2)\cos(\Delta\phi/2) - j\sin(\Delta\phi/2) \end{aligned} \quad (4)$$

and $\underline{s}'(t) = \underline{s}(t)e^{-j\Delta\omega t}$. In (3) $\underline{s}'(t)$ and $\widetilde{\underline{s}}(t)$ refer to the complex envelope of the received signal including the CFO and the received signal with CFO and I/Q imbalance effects respectively. Moreover the first term on the right-hand side of (3) indicates the useful signal while the second term represents the interference from the image signal. Using (3) and (4) the Signal to Interference Ratio (SIR) due to a given phase and gain imbalance can be derived as $\mathrm{SIR} = \frac{|\mu|^2}{|\nu|^2}$.

2.2. OFDM receiver

The complex envelope at the input of the receiver for one OFDM symbol can be written as [1]

$$\underline{s}(t) = \sum_{k=0}^{N-1} H[k]X[k]e^{2\pi k\Delta f_c t}, 0 \leq t \leq N-1 \quad (5)$$

where Δf_c is the inter-carrier spacing, T_s is the sampling period, $H[k]$ is the channel frequency response and $X[k]$ is the transmitted M-QAM constellation point on the sub-carrier k. The parameter N indicates the size of the Discrete Fourier Transform (DFT).

Fig. 2 depicts a simplified digital base band OFDM receiver. The CFO is compensated using a Numerically Controlled Oscillator (NCO) that generates a complex sine of angular frequency $\widehat{\Delta\omega}$, which is estimated during the receiver's training phase. In the following analysis we make the assumption that the estimated CFO is close to the actual CFO. Under this hypothesis, which is only for simplification of the analysis, the sampled signal after CFO compensation is

$$\underline{y}[n] = \widetilde{\underline{s}}[n] = e^{j\widehat{\Delta\omega}t_n} \cong \mu\underline{s}[n] + \nu\underline{s}[n]^* e^{j2\Delta\omega t_n} \quad (6)$$

The OFDM receiver after serial to parallel conversion of the signal $\underline{y}[n]$ and cyclic prefix removal (CPR) performs a DFT. The resultant signal in the frequency domain is given by

$$Y[k] = \mu H[k]X[k] + \nu\sum_{l=0}^{N-1} H[l]^* X[l]^* I(k,l,\Delta\omega) \quad (7)$$

The term I in (7) that represents the Inter-Carrier Interference (ICI) in the presence of I/Q imbalance and CFO is

$$I(k,l,\Delta\omega) = \frac{1}{N}\sum_{n=0}^{N-1} e^{-j\frac{2\pi}{N}(k+l)n + 2\Delta\omega t_n} \quad (8)$$

From (8) it can be seen that for $\Delta\omega \neq 0$ all the carriers contribute to the interference. Hence the algorithm described in [5] that assumes crosstalk between symmetrical sub-carriers pairs only is not applicable for I/Q imbalance compensation in the presence of digitally compensated CFO.

3. MIXED TIME-FREQUENCY DOMAIN I/Q IMBALANCE COMPENSATION

3.1. Interference cancellation structure

The proposed adaptive interference cancellation structure that performs I/Q imbalance compensation is shown in Fig. 3. This block implements a simple linear combination of the *primary input* i.e. the received signal $\widetilde{\underline{s}}[n]$ and the *reference input* i.e. the corresponding complex conjugate value $\widetilde{\underline{s}}[n]^*$, and provides the following output

$$\underline{r}[n] = \widetilde{\underline{s}}[n] - w^*\widetilde{\underline{s}}[n]^* \quad (9)$$

This method of compensation in the time domain avoids the problem related to a frequency domain technique that is subject to ICI interference resulting from I/Q imbalance and CFO.

It can be easily derived that the optimum coefficient w for this simple one tap compensation filter is given by $w_{opt} = \nu^*/\mu$ where ν and μ are defined in (4). The optimum value achieves perfect imbalance compensation resulting in the complete suppression of the image signal.

3.2. Procedure for coefficient adaptation

The structure depicted in Fig. 3 is based on an Adaptive Noise Cancellation (ANC) technique [6]. If the classical ANC algorithm is employed either in the context of a Zero-IF or a Low-IF architecture, SIR improvement at the output of the compensator is limited by the SIR of the reference input [3][7]. This limitation is overcome by the proposed coefficient adaptation method depicted in Fig. 4. The output of the DFT for the k-th sub-carrier in an OFDM symbol m can be written as

$$Y_m[k] = \beta_m[k] - w^*\alpha_m[k] \tag{10}$$

where

$$\begin{aligned} \beta_m[k] &= DFT_k\left\{\underline{\tilde{s}}_m[n]e^{j\widehat{\Delta\omega}t_{m,n}}\right\} \\ \alpha_m[k] &= DFT_k\left\{\underline{\tilde{s}}_m[n]^*e^{j\widehat{\Delta\omega}t_{m,n}}\right\} \end{aligned} \tag{11}$$

Denoting $\hat{H}_m[k]$ and $\hat{X}_m[k]$ as the estimated channel and the slicer output respectively for the sub-carrier k in the m-th OFDM symbol, we form the reference signal $\hat{Y}_m[k] = \hat{H}_m[k]\hat{X}_m[k]$ that is free of ICI due to imbalance. Defining the error signal as $e_m[k] = \hat{Y}_m[k] - Y_m[k]$, the cost function is used to determine the optimum value of the compensator coefficient w. Minimizing the cost function $J(w) = E\{|e_m[h]|^2\}$ leads to the classical Wiener-Hopf solution that is given by

$$\nabla J(w) = 2w\xi + 2\gamma = 0 \Rightarrow w_{opt} = -\frac{\xi}{\gamma} \tag{12}$$

where

$$\begin{aligned} \xi &= E\left\{\alpha_m[k]\alpha_m[k]^*\right\} \\ \gamma &= E\left\{\alpha_m[k](\hat{Y}_m[k] - \beta_m[k])^*\right\} \end{aligned} \tag{13}$$

The classical Wiener-Hopf solution in (12) can be obtained using the iterative method of the steepest descent leading to the update equation

$$w_{m+1} = w_m - \delta\nabla J(w_m) \tag{14}$$

However due to difficulties in accurate measurement of the gradient $\nabla J(w_m)$, an iterative procedure based on the LMS

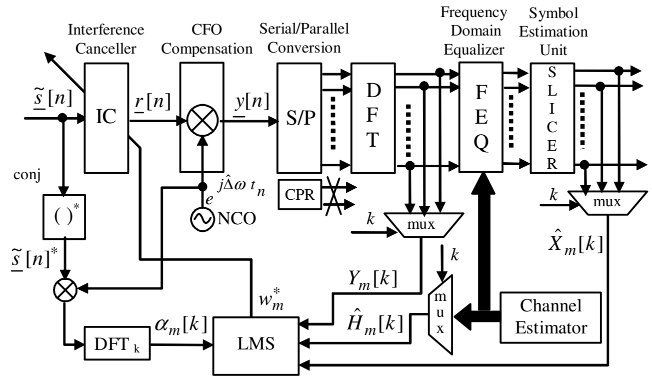

Fig. 4. Mixed time-frequency domain I/Q imbalance cancellation scheme.

[6] for computation of the optimum filter coefficient is adopted. The recursive update equation is given by

$$w_{m+1} = w_m - 2\delta\alpha_m[k]e_m[k]^* \tag{15}$$

where the step-size δ decides the rate of convergence and stability. The condition to ensure convergence is

$$0 < \delta < \xi^{-1} \tag{16}$$

In a burst transmission system w is updated every symbol where the initial coefficient of the current burst corresponds to the last update of w in the previous burst. As it has been shown, the proposed method for coefficient adaptation is based on a sub-carrier k that needs to be selected by an appropriate mechanism. Since we are considering the receiver operation for frequency selective channels, the criteria adopted for the sub-carrier selection is

$$k = \arg\left\{\max_{l\in[0\ldots N-1]} |H[l]|\right\} \tag{17}$$

4. SIMULATION RESULTS

Simulations were performed with a system model of the IEEE 802.11a standard [4] compliant physical layer. Performance analysis was done for two channel models, A and G, where A is a typical multipath indoor office environment with non-line-of-sight propagation and 50 ns of rms delay spread [8] while G represents an environment with non multipath line-of-sight propagation. All simulations included Additive White Gaussian Noise (AWGN), CFO and I/Q phase and gain imbalance impairments. A non-trivial RF design target $\Delta\phi = 4^o$ of phase imbalance and of $\epsilon = 12\%$ of gain imbalance, resulting in an SIR of 23.5, were employed in the simulations. The maximum CFO of 40 ppm as specified in the standard was used and digitally compensated using the estimation performed during the preamble of the burst. All the simulation results correspond to

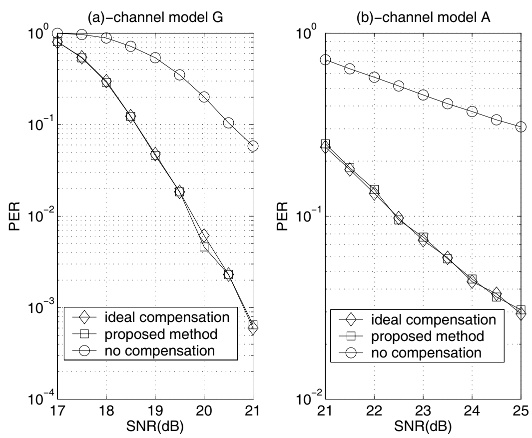

Fig. 5. PER vs. SNR for channel model G and A.

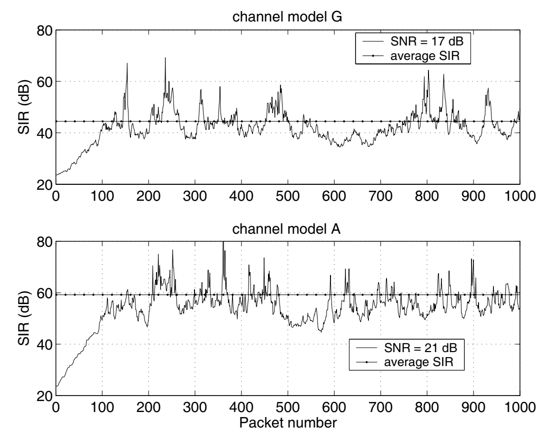

Fig. 6. SIR after compensation for channel G and A.

54 Mbit/s mode (64-QAM rate 3/4) with a packet length of 1000 bytes. Simulations with I/Q imbalance compensation using the proposed method, no compensation and ideal imbalance compensation were performed for both channels. An adaptation step-size $\delta = 0.01\xi^{-1}$ was adopted for all simulations.

Fig. 5 show the Packet Error Rate (PER) vs. SNR for channel G and A. For both channels the proposed method achieves a performance almost identical to that of an ideal compensator. Considering a reference PER of 10% as specified in [4], we can observe from Fig. 5(a) that for channel G the performance gain using the proposed method is approximately 1.9 dB. From Fig. 5(b) that refers to multipath channel, the advantage given by the technique can be estimated, by extrapolation, in the range of 6-7 dB.

Fig. 6 depicts the convergence of the proposed method for channel G and A, for two different SNRs. The plots show that even at low SNR convergence is achieved within 100 packets, which corresponds to nearly 15 ms of data transmission/reception time. For channel G with an SNR as low as 17 dB, upon convergence the minimum and average obtained SIR are 35 dB and 44 dB respectively. Similarly for channel A and an SNR of 21 dB the minimum and average obtained SIR are 45 dB and 56 dB. Hence for both the channel types the algorithm convergence ensures that the residual interference is well below the AWGN level.

5. CONCLUSIONS

In this paper we have presented a mixed time-frequency domain technique for digital I/Q imbalance compensation in the presence of CFO. With the proposed method a performance almost identical to that of an ideal compensator can be achieved even in the presence of carrier frequency offset and frequency selective channels. This digital scheme enhances the image rejection and can lead to relaxed RF requirements and increased receiver sensitivity.

6. REFERENCES

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