COMPENSATION OF TRANSMITTER IQ IMBALANCE FOR OFDM SYSTEMS.

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Abstract—**Zero-IF transceivers are gaining interest because of their potential to enable low-cost OFDM terminals. However, the Zero-IF architecture introduces IQ imbalance which may have a huge impact on the performance. Rather than increasing component cost to decrease the IQ imbalance, an alternative is to tolerate the IQ imbalance and compensate it digitally. Current solutions require extra analog hardware at the transmitter. In this paper, we analyze the transmit IQ imbalance estimation and propose a low-cost, highly effective estimation scheme, which is fully digital and located at the receiver. Performance analysis shows that this scheme can provide up to 4 dB gain while meeting the IEEE 802.11a constellation accuracy specification and more if larger IQ imbalance is present in the transmitter. It therefore enables the design of low-cost, low-complexity OFDM modems.**

I. INTRODUCTION

OFDM is a widely recognized and standardized modulation technique [1], [2]. Unfortunately, OFDM is also sensitive to front-end non-idealities [3]. This sensitivity leads either to heavy front-end specifications and thus an expensive front-end or to large performance degradations. IQ imbalance has been identified as a key front-end effect for OFDM systems.

Therefore, we investigate the IQ imbalance estimation and compensation and introduce a low-complexity compensation scheme to combat the IQ imbalance. Current transmitter IQ imbalance solutions use calibration, such as in [4], [5] and the references listed there, require extra analog hardware, which needs to be carefully designed in order not to introduce any IQ imbalance itself. This is done to meet the constellation accuracy specification e.g. in the WLAN standards. However, even if these accuracy specs are met, we will show that the performance can suffer a 4 dB degradation.

In this paper, we propose an all-digital, low-complexity transmitter IQ imbalance compensation which converges within 1 OFDM training symbol. The compensation is done at the receiver and is fully digital. Thus it does not require any additional analog hardware, contrary to existing solutions.

II. IQ IMBALANCE

This section introduces the IQ imbalance model and its impact on OFDM.

A. Effect/Model

IQ imbalance can be characterized by 2 parameters: the amplitude imbalance ϵ between the I and Q branch, and the phase orthogonality mismatch $\Delta\phi$. The complex baseband equation for the IQ imbalance effect on the ideal time domain signal $\mathbf{r}$ is given by [6] as

$$\begin{aligned}\mathbf{r_{iq}} &= (1+\epsilon)\cos\Delta\phi\Re\{\mathbf{r}\} - (1+\epsilon)\sin\Delta\phi\Im\{\mathbf{r}\} \\ &+ \jmath[(1-\epsilon)\cos\Delta\phi\Im\{\mathbf{r}\} - (1-\epsilon)\sin\Delta\phi\Re\{\mathbf{r}\}] \qquad (1) \\ &= (\cos\Delta\phi + \jmath\epsilon\sin\Delta\phi)\cdot\mathbf{r} + (\epsilon\cos\Delta\phi - \jmath\sin\Delta\phi)\cdot\mathbf{r}^* \\ &= \alpha\cdot\mathbf{r} + \beta\cdot\mathbf{r}^* \qquad (2)\end{aligned}$$

with $\mathbf{r_{iq}}$ the time domain signal with IQ imbalance, $\Re()$ denotes the real part, $\Im()$ the imaginary part and $()^*$ the complex conjugate and

$$\alpha = \cos\Delta\phi + \jmath\epsilon\sin\Delta\phi \qquad (3)$$
$$\beta = \epsilon\cos\Delta\phi - \jmath\sin\Delta\phi \qquad (4)$$

Frequency domain signals are underscored, while time domain signals are not. Signals are indicated in bold and scalar parameters in normal font.

Throughout the rest of the paper, the term IQ parameters refers to α and β for calculations and estimations ; to indicate physical parameters, however, we use the more direct ϵ and $\Delta\phi$.

We also analyze the effect of the IQ imbalance in the frequency domain. If $\underline{\mathbf{r}} = FFT\{\mathbf{r}\}$, then applying the IQ imbalance (2) on $\mathbf{r}$ and transforming the time domain signal to the frequency domain leads to

$$\begin{aligned}\underline{\mathbf{r}}_{\mathbf{iq}} &= FFT\{\alpha\cdot IFFT(\underline{\mathbf{r}}) + \beta\cdot[IFFT(\underline{\mathbf{r}})]^*\} \\ &= \alpha\cdot\underline{\mathbf{r}} + \beta\cdot\underline{\mathbf{r}}^*_{\mathbf{m}} \qquad (5)\end{aligned}$$

where $\underline{\mathbf{r}}_{\mathbf{iq}}$ is the OFDM symbol with IQ imbalance and $\underline{\mathbf{r}}_{\mathbf{m}}$ the OFDM symbol, mirrored over the carriers: $(\underline{\mathbf{r}})_m(i) = (\underline{\mathbf{r}})(mod(N_{sc}-i+2, N_{sc}))$, with N_{sc} the number of sub-carriers in the OFDM symbol, $1 \leq i \leq N$ and mod the modulo operation. Carrier 1 is the DC carrier.

In this paper we focus on IQ compensation for bursty communication, for which channel estimation is performed on the basis of a known training symbol. Both IEEE802.11a [1] and HIPERLAN-II [2] provide such a Long Training Symbol (LTS) (BPSK symbol $\underline{\mathbf{t}}$) in the preamble. The effect of IQ imbalance on channel estimation can be calculated based on (5)

$$\begin{aligned}\underline{\mathbf{h}} &= \underline{\mathbf{t}}\cdot[\underline{\mathbf{c}}(\alpha\underline{\mathbf{t}} + \beta\underline{\mathbf{t}}^*_{\mathbf{m}}) + \underline{\mathbf{n}}] \\ &= \underline{\mathbf{c}}[\alpha + \beta\cdot\underline{\mathbf{t}}'] + \underline{\mathbf{t}}\cdot\underline{\mathbf{n}} \qquad (6)\end{aligned}$$

where $\underline{\mathbf{h}}$ is the channel estimate calculated from the LTS, $\underline{\mathbf{c}}$ is the exact channel vector, $\underline{\mathbf{n}}$ the AWGN noise vector and $\underline{\mathbf{t}}' = \underline{\mathbf{t}}\cdot\underline{\mathbf{t}}_{\mathbf{m}}$.

B. Estimation

From (6) the corrected channel response is easily derived as (making abstraction of the noise)

$$\hat{\underline{\mathbf{c}}} = \frac{\underline{\mathbf{h}}}{\alpha + \beta\underline{\mathbf{t}'}} \qquad (7)$$

This equation allows us to compute the corrected channel response $\hat{\underline{\mathbf{c}}}$ based on the measured channel response $\underline{\mathbf{h}}$ and the IQ imbalance parameters (α,β).

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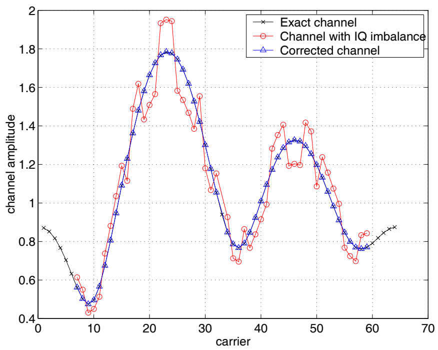

Fig. 1. The effect of IQ imbalance and correction on channel estimation.

The estimation of α and β is based on the information that the corrected channel response should have a smooth channel characteristic: since the coherence bandwidth of the channel is (a lot) larger than the inter-carrier-spacing in a WLAN system, the channel response does not change substantially between successive frequency taps (the x-line in figure 1). With IQ imbalance, sharp transitions occur in the measured channel response $\underline{\mathbf{h}}$ due to the β degradation term (the o-line in figure 1). Thus, correcting the IQ imbalance means making the channel response 'smooth' again. Therefore, we select the set of IQ parameters (α, β) which renders the corrected channel response $\underline{\hat{\mathbf{c}}}$ as smooth as possible ; in other words we minimize the Mean Square Error (MSE) between consecutive channel coefficients

$$MSE = \sum_l |\underline{\hat{\mathbf{c}}}_{l+1} - \underline{\hat{\mathbf{c}}}_l|^2 \tag{8}$$

To minimize the MSE we derive (8) towards β, as follows

$$\begin{aligned}
\frac{\partial MSE}{\partial \beta} &= \frac{\partial \sum_l |\underline{\hat{\mathbf{c}}}_{l+1} - \underline{\hat{\mathbf{c}}}_l|^2}{\partial \beta} \\
&= \sum_l \frac{\partial \left| \frac{\underline{\mathbf{h}}_{l+1}}{\alpha+\beta\cdot\underline{\mathbf{t}}'_{l+1}} - \frac{\underline{\mathbf{h}}_l}{\alpha+\beta\cdot\underline{\mathbf{t}}'_l} \right|^2}{\partial \beta} \\
&= \sum_l \frac{\partial \left| \frac{\alpha(\overbrace{\underline{\mathbf{h}}_{l+1} - \underline{\mathbf{h}}_l}^{\mathbf{u_l}}) + \beta(\overbrace{\underline{\mathbf{t}}'_l\underline{\mathbf{h}}_{l+1} - \underline{\mathbf{t}}'_{l+1}\underline{\mathbf{h}}_l}^{\mathbf{v_l}})}{(\alpha+\beta\cdot\underline{\mathbf{t}}'_{l+1})(\alpha+\beta\cdot\underline{\mathbf{t}}'_l)} \right|^2}{\partial \beta}
\end{aligned}$$

The denumerator in the optimization will not vary much as a function of β, allowing it to be considered as a constant when minimizing towards β. Thus

$$\begin{aligned}
\frac{\partial MSE}{\partial \beta} &\approx \sum_l \frac{\partial |\alpha\mathbf{u}_l - \beta\mathbf{v}_l|^2}{\partial \beta} \\
&= \sum_l \Big(\frac{\partial(|\alpha|^2|\mathbf{u}_l|^2)}{\partial \beta} + \frac{\partial(\beta^*\alpha\mathbf{u}_l\mathbf{v}_l^*)}{\partial \beta} \\
&+ \frac{\partial(\alpha^*\beta\mathbf{u}_l^*\mathbf{v}_l)}{\partial \beta} + \frac{\partial(|\beta|^2|\mathbf{v}_l|^2)}{\partial \beta} \Big)
\end{aligned} \tag{9}$$

At this point in the derivation we assume α and β to be independent. Since α and β both depend on ϵ and $\Delta\phi$ (3 and 4) this estimation is not completely optimal as it ignores this extra information. The validity of this approximation is verified through the simulations.

For the derivative to β to exist, we need to solve the Cauchy-Riemann equations. As the MSE is always real, the equations reduce to

$$\left\{ \begin{array}{rcl} \sum_l \frac{\partial |\alpha|^2|\mathbf{u}_l|^2 + \beta^*\alpha\mathbf{u}_l\mathbf{v}_l^* + \alpha^*\beta\mathbf{u}_l^*\mathbf{v}_l + |\beta|^2|\mathbf{v}_l|^2}{\partial \beta_r} & = & 0 \\ \sum_l -\jmath \frac{\partial |\alpha|^2|\mathbf{u}_l|^2 + \beta^*\alpha\mathbf{u}_l\mathbf{v}_l^* + \alpha^*\beta\mathbf{u}_l^*\mathbf{v}_l + |\beta|^2|\mathbf{v}_l|^2}{\partial \beta_i} & = & 0 \end{array} \right. \tag{10}$$

Solving both equations for β_r and β_i ($\beta = \beta_r + \jmath\beta_i$) leads to

$$\left. \begin{array}{rcl} \hat{\beta}_r & = & -\frac{\sum_l \Re\{\alpha\mathbf{u}_l^*\mathbf{v}_l\}}{\sum_l |\mathbf{v}_l|^2} \\ \hat{\beta}_i & = & \frac{\sum_l \Im\{\alpha\mathbf{u}_l^*\mathbf{v}_l\}}{\sum_l |\mathbf{v}_l|^2} \end{array} \right\} \Rightarrow \hat{\beta} = \hat{\beta}_r + \jmath\hat{\beta}_i = -\frac{\sum_l \alpha^*\mathbf{u}_l\mathbf{v}_l^*}{\sum_l |\mathbf{v}_l|^2} \tag{11}$$

Approximating $\alpha^* \approx 1$ and resubstituting $\mathbf{u_l}$ and $\mathbf{v_l}$ gives us the MMSE estimate of β

$$\hat{\beta} = -\frac{\sum_l (\underline{\mathbf{h}}_{l+1} - \underline{\mathbf{h}}_l)(\underline{\mathbf{t}}'_l\underline{\mathbf{h}}_{l+1} - \underline{\mathbf{t}}'_{l+1}\underline{\mathbf{h}}_l)^*}{\sum_l |\underline{\mathbf{t}}'_l\underline{\mathbf{h}}_{l+1} - \underline{\mathbf{t}}'_{l+1}\underline{\mathbf{h}}_l|^2} \tag{12}$$

An estimate of α is derived based on the estimate of β (12). This can be done because both complex IQ parameters depend on the scalar IQ parameters ϵ and $\Delta\phi$. Therefore, once we obtain an estimate of the complex parameter β, its real and imaginary part contain sufficient information to calculate ϵ and $\Delta\phi$, or equivalently, the real and imaginary part of α.

From (3) and (4) we derive

$$\begin{aligned}
\Re\{\alpha\} &= \cos\Delta\phi \\
\Im\{\alpha\} &= \epsilon\sin\Delta\phi \\
\Re\{\beta\} &= \epsilon\cos\Delta\phi \\
\Im\{\beta\} &= -\sin\Delta\phi
\end{aligned}$$

This means

$$\Re\{\alpha\}\Im\{\alpha\} = -\Re\{\beta\}\Im\{\beta\} \tag{13}$$

$$\Im^2\{\beta\} + \Re^2\{\alpha\} = 1 \tag{14}$$

Solving equations (13) and (14) for $\Re\{\alpha\}$ and $\Im\{\alpha\}$ leads to

$$\begin{aligned}
\Re\{\alpha\} &= \sqrt{1 - \Im^2\{\beta\}} \\
\Im\{\alpha\} &= -\frac{\Re\{\beta\}\Im\{\beta\}}{\sqrt{1 - \Im^2\{\beta\}}}
\end{aligned}$$

and thus

$$\hat{\alpha} = \sqrt{1 - \Im^2\{\hat{\beta}\}} - \jmath\frac{\Re\{\hat{\beta}\}\Im\{\hat{\beta}\}}{\sqrt{1 - \Im^2\{\hat{\beta}\}}} \tag{15}$$

Figure 1 shows that we can correct the influence of the IQ imbalance on the channel estimate extremely well using (7)-(12)-(15): the corrected channel response (the Δ-line) coincides (almost) perfectly with the exact channel response (the x-line).

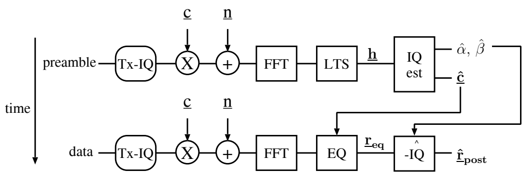

Fig. 2. Overview of the IQ estimation and post-compensation.

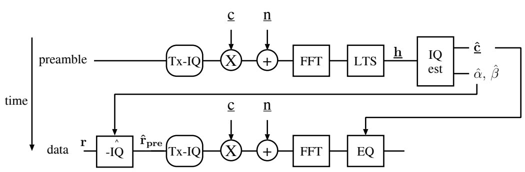

Fig. 3. Overview of the IQ estimation and pre-compensation.

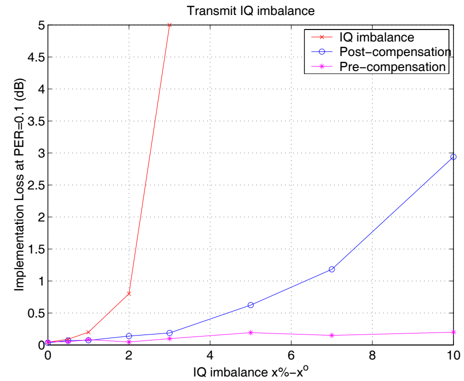

Fig. 4. Performance of the IQ estimation and compensation scheme.

C. Compensation

This estimation algorithm provides us with a corrected channel response and an estimate of the IQ parameters α and β (or equivalently ϵ and $\Delta\phi$). Since ϵ and $\Delta\phi$ and thus α and β are typically static over many symbols, we can use their estimates from the channel correction also for the correction of the IQ imbalance on the data.

At the receiver, we first correct the channel impact by equalizing the data. To compensate the IQ imbalance on the equalized data $\underline{\mathbf{r}}_{\mathbf{eq}}$, we need to derive a frequency-domain IQ imbalance correction. This can be found by solving equation 5 for $\underline{\mathbf{r}}$ as

$$\hat{\underline{\mathbf{r}}}_{\mathbf{post}} = \frac{\hat{\alpha}^* \cdot \underline{\mathbf{r}}_{\mathbf{eq}} - \hat{\beta} \cdot (\underline{\mathbf{r}}_{\mathbf{eq}})^*_{\mathbf{m}}}{|\hat{\alpha}|^2 - |\hat{\beta}|^2} \tag{16}$$

The block diagram of this scheme is shown in figure 2. To assess the performance of the IQ imbalance estimation/compensation scheme, we performed simulations for coded (R=3/4 from the IEEE 802.11a standard) 64QAM in a multi-path environment. The multi-path channel consists of 4 independent equal power Rayleigh fading taps.

The performance of this post-compensation scheme is shown in figure 4. The impact of transmit IQ imbalance remains below 1 dB implementation loss at a Packet Error Rate (PER) of 10% for IQ imbalances below (2%,2^o), but rises quickly for higher IQ imbalance as indicated by the 'x'-line. The post-compensation schemes (shown by the 'o'-line) reduces the remaining degradation below 1 dB for IQ imbalances up to (6%,6^o) and compensates IQ imbalance of (10%,10^o) with an implementation loss of 3 dB. The increasing implementation loss with increasing IQ imbalance comes from the fact that the IQ imbalance is compensated based on the equalized data stream. As OFDM uses zero-forcing equalization, it suffers from noise enhancement. Therefore, also the IQ imbalance post-compensation gets distorted by this noise enhancement.

An alternative solution to avoid the performance degradation of the post-compensation scheme for high IQ imbalance is to pre-compensate the IQ imbalance at the transmitter, by feeding back the IQ imbalance estimates to the transmitter as shown in figure 3.

To pre-compensate the IQ imbalance on the time domain data signal $\mathbf{r}$ at the transmitter, we need to apply a time domain correction, equivalent to the frequency domain correction in equation 16

$$\hat{\mathbf{r}}_{\mathbf{pre}} = \frac{\hat{\alpha}^* \cdot \mathbf{r} - \hat{\beta} \cdot \mathbf{r}^*}{|\hat{\alpha}|^2 - |\hat{\beta}|^2} \tag{17}$$

After pre-compensation, only the difference between the estimated IQ imbalance and the actual analog IQ imbalance appears at the output of the transmitter. As our estimation is very accurate this error is very small and thus the remaining degradation as well. Moreover, as the compensation is applied before the noise is added to the link, the IQ imbalance pre-compensation does not suffer from the noise enhancement.

The drawback of the pre-compensation method is the delay caused by the feedback of the IQ imbalance estimations ($\hat{\alpha}$,$\hat{\beta}$) to the transmitter: as the parameters are estimated at the receiver based on the preamble of the data burst, the IQ imbalance cannot be precompensated on that burst, but only starting from the following burst. Therefore, the first burst will most likely be received erroneously resulting in a throughput penalty. However, this first burst loss can be overcome by applying post-compensation on the first burst and pre-compensation on all the following bursts.

D. Interpretation of the results

For 64QAM the most stringent constellation accuracy requirement applies [1]. However, this still allows up to (2.8%,2.8^o) IQ imbalance. As can be seen from figure 4, this causes a extra PER performance degradation of 4 dB, which can be reduced even by the post-compensation scheme to 0.2 dB. Therefore, our scheme can bring up to 4 dB in performance gain for standard-compliant modems and even more if the constellation accuracy requirement is relaxed.

The low remaining degradation after the pre-compensation indicates that the estimation accuracy is near-optimal.

The IQ estimation/compensation algorithm requires no extra analog hardware and a very small additional digital complexity. The IQ imbalance may occur anywhere in the receiver, because the RF, LO and baseband contributions are jointly estimated and compensated.

Since the IQ estimation only requires a known Training Symbol, this scheme is applicable to any system which uses a Training Symbol to estimate the channel (e.g. Wireless LAN [2], [1] and Broadband Fixed Wireless Access [7]). Moreover, the scheme is also independent of the data that follows the Training Symbol. This means it is applicable to any constellation size and to OFDM as well as Single-Carrier with Frequency-Domain Processing.

III. Conclusions

In this paper, we introduced a low-complexity estimation/compensation scheme that tackles the IQ imbalance caused by Zero-IF transmitters. The estimation scheme is located at the receiver and is fully digital. It converges on one OFDM training symbol. For standard-compliant modems, it can improve performance by 4 dB without any overhead and relax the IQ imbalance design specs. For larger transmitter IQ imbalance, the gains are increasing rapidly still without any additional overhead and even more if some limited feedback to the transmitter is possible. In conclusion, our estimation/compensation scheme enables the design of low-cost, low-complexity OFDM modems.

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