The influence of DC offsets on the digital cancellation of second-order TX intermodulation distortions in homodyne receivers

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I. Introduction

Intermodulation distortions are produced when a modulated blocker signal with a nonconstant envelope gets processed by a stage with a nonlinear characteristic. Especially in homodyne receivers (RX), where the RX signal is directly downconverted into the baseband, even-order intermodulations might overlap the wanted signal in its baseband. In Frequency Division Duplex (FDD) systems, where RX and transmitter (TX) are active at the same time, the own TX is the source of major intermodulation distortions, since the TX signal power can be 130dB larger than that of the RX signal [1]. Usually the isolation of the duplexers, which are used to seperate TX and RX path at the antenna, is limited. As a consequence parts of the TX signal are leaking into the RX path. Due to nonlinearities, especially of the downconverter, intermodulation distortions are produced, degrading the performance of the RX. The quadratic term in the transfer functions of those nonlinear stages produces the major distortions, the so called second-order intermodulation distortions (IMD2). One approach to avoid IMD2 is to attenuate the TX leakage before the downconversion in the RX takes place. This would require analog filters which cannot be integrated into the radio frequency (RF) transceiver chip, increasing the production costs of the wireless communication device. Furthermore, in multiband RX, separate bandpass filters are needed for different signal bands.

Since the TX signal is known, an estimation and compensation of the part leaking into the RX signal is possible. In [2] an adaptive filter in the RF domain is used to cancel the TX leakage. This method is very sensitive to time-variations between the TX signal and its leakage in the RX path, since the compensation is done in the RF domain. The sensitivity to time-variations can be relaxed if the cancellation is performed after the down- and analog-/digital converter in the digital domain. Using this approach it is necessary to estimate and compensate the IMD2 because the TX leakage has already been processed by the nonlinear downconverter. In [3] a single adaptive filter is used to compensate the IMD2 for the I- and Q-component of the RX baseband signal in common. A matched I- and Q-path or a linear time-invariant relation between them, which might be determined in the calibration process of the transceiver chip, is assumed. Whereas, in [4], it has been shown that the IMD2 compensation can be performed independently using two separate adaptive filters for the I- and the Q-path, without any assumptions on the I/Q mismatch. In the present paper the latter approach has been used for the cancellation. For this approach, the theoretical dependence of the cancellation performance on the DC voltage in the RX signal will be derived. The result of this theoretical derivation will be verified with signals measured at the digital front end (DFE) output of a Long Term Evolution (LTE) transceiver chip.

Starting with the discussion of intermodulation distortions in section II, the noise canceller used to estimate and furthermore eliminate the IMD2 is discussed in section III. The actual derivation of the cancellation performance for different DC voltages can be found in section III-A. In section IV the remaining IMD2 of measured and postprocessed signals will be compared to the theoretical limits calculated in the preceding section.

II. Intermodulation distortions

In this section the mechanism which produces intermodulation distortions will be discussed. Suppose a real passband signal

$$s(t) = I(t)\cos(\omega_c t) - Q(t)\sin(\omega_c t) \tag{1}$$

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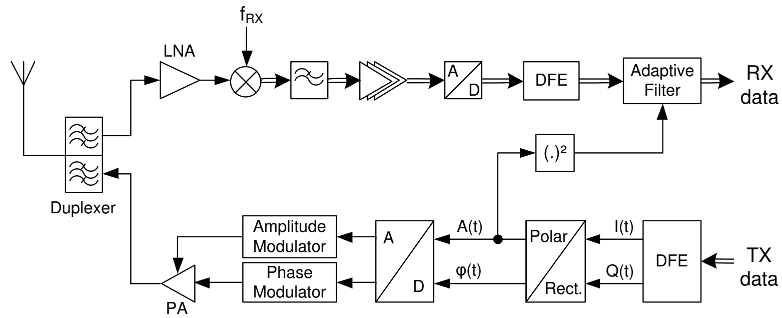

Fig. 1. Block diagram of a transceiver using polar modulation for transmission

is being received. Here $I(t)$ represents the in-phase, $Q(t)$ represents the quadrature-phase component of the passband signal and ω_c is the angular frequency of the carrier. The envelope-phase representation of (1) is

$$s(t) = A(t)\cos(\omega_c t + \varphi(t)) \tag{2}$$

where the envelope is given by

$$A(t) = \sqrt{I^2(t) + Q^2(t)}. \tag{3}$$

If the signal (2) passes a nonlinear device with the following characteristic

$$y(t) = \alpha_1 s(t) + \alpha_2 s^2(t) + \alpha_3 s^3(t) + ... \tag{4}$$

the following signal components will be produced by the quadratic term of the characteristic (4):

$$\frac{1}{2}\alpha_2 A^2(t)\left[1 + \cos(2\omega_c t + 2\varphi(t))\right] \tag{5}$$

High frequency components around $2\omega_c$ can easily be attenuated with the channel select filter after the downconversion, since they are far away from the baseband. Nevertheless the component

$$IMD2_{BB}(t) = \alpha_2 \frac{1}{2}\left(I^2(t) + Q^2(t)\right) \tag{6}$$

is falling around 0 Hz and is therefore distorting the wanted RX baseband signal. From (6) it can be followed that this kind of distortion is linear dependent on the squared envelope of the blocker which results in a quadratic increase of the noise power with the power of the blocker signal.

III. Cancellation of the second-order intermodulation distortions

The used cancellation method is based on an adaptive filter used as interference canceller [5]. Here an adaptive filter is used to estimate the linear dependency between a reference and the distorted signal. After subtraction of the estimated distortion from the distorted signal, the remaining signal should be free of distortions which are linear dependent on the reference signal. From (6) it can be followed that the squared envelope of the TX signal should be taken as reference for the IMD2 canceller. In case of a polar transmitter, the envelope of the TX signal is already available and does only have to be squared as shown in Fig. 1.

A. DC offset considerations

In the case of IMD2 cancellation the distorting signal is proportional to the squared magnitude of the TX signal, which can be separated into its zero-mean component $u[k]$ and its mean μ_u. The adaptive filter is trying to minimize the Mean Square Error (MSE) by estimating common filter coefficients $w[k]$ for the zero-mean component as well as the DC component of the reference signal, as shown in (7).

$$\hat{y}[k] + \mu_{\hat{y}} = w[k] * (u[k] + \mu_u) \tag{7}$$

The estimated zero-mean $\hat{y}[k]$ component and the estimated mean $\mu_{\hat{y}}$ of the IMD2 are subtracted from the IMD2 $(d[k] + \mu_d)$ as can be seen in Fig. 2. If the signal in the RX path consists of DC voltages additional to the one from the IMD2, the IMD2 cannot be cancelled completely. To avoid this problem, the DC offset can be compensated in the RX as well as in the reference path separately before the IMD2 cancellation takes place. While a DC offset compensation is usually done in the RX path to compensate all DC sources, an additional compensation of the DC in the reference path might be too costly for practical implementations. Therefore the influence of unmatched DC components in the RX and reference signals have to be analyzed. Matching in this case means that the filter coefficients are appropriate to compensate the DC as well as the

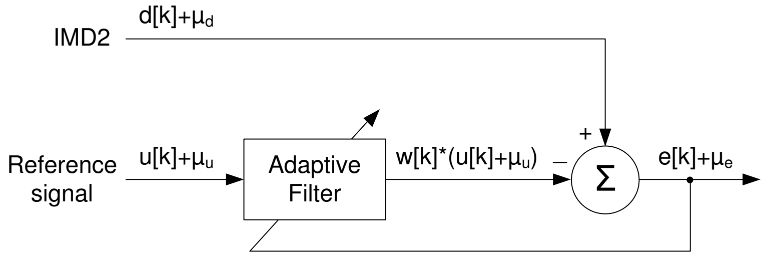

Fig. 2. Adaptive filter used as noise canceller [5]

zero-mean component of the IMD2 at the same time. If the reference signal and the IMD2 are perfectly time aligned and its DC components are matched, a single filter coefficient would be sufficient to estimate the dependency of those signals. Due to the fact that the TX signal is processed by several analog stages in the RF domain before it causes intermodulation distortions in the RX baseband signal, the time delay between the reference signal and the IMD2 is usually a fractional part of the sampling period in the digital domain of the RX. This is one reason that more than one filter coefficient is used for the IMD2 cancellation. A more detailed description on this issue can be found in [6].

To show the influence of unmatched DC components between the reference and IMD2 on the cancellation performance, the Minimum Mean Square Error (MMSE) of the estimation will be derived. For this derivation the IMD2 canceller can be thought as in Fig. 2, getting the IMD2 as desired signal and producing an error signal equal to zero and furthermore an MMSE of zero in the ideal case. Other signal components in the RX signal, like the actually wanted signal, are statistically independent from the reference signal and can therefore be neglected for the following considerations.

The MSE of the estimation error can be calculated as

$$J = E(|e[k] + \mu_e|^2) = E(|d[k] + \mu_d - \mathbf{w}^T(\mathbf{u}[k] + \boldsymbol{\mu}_u)|^2) \tag{8}$$

Since $d[k]$ and $u[k]$ are zero-mean and uncorrelated, J reduces to

$$\begin{aligned} J = \sigma_d^2 + \mu_d^2 - \mathbf{w}^T(\mathbf{p} + \boldsymbol{\mu}_u\mu_d) - (\mathbf{p}^T + \mu_d\boldsymbol{\mu}_u^T)\mathbf{w} \\ +\mathbf{w}^T(\mathbf{R} + \boldsymbol{\mu}_u\boldsymbol{\mu}_u^T)\mathbf{w} \end{aligned} \tag{9}$$

where

$$\mathbf{p} = E(\mathbf{u}[k]d[k]) \tag{10}$$

is the cross-covariance vector between the reference signal (input of the filter) and the IMD2.

$$\mathbf{R} = E(\mathbf{u}[k]\mathbf{u}^T[k]) \tag{11}$$

represents the autocovariance matrix of the reference signal, σ_d^2 represents the zero-mean and μ_d^2 the DC power of the IMD2. Equation (9) can be expanded into the following quadratic form

$$\begin{aligned} J = \sigma_d^2 + \mu_d^2 - (\mathbf{p} + \boldsymbol{\mu}_u\mu_d)^T(\mathbf{R} + \boldsymbol{\mu}_u\boldsymbol{\mu}_u^T)^{-1}(\mathbf{p} + \boldsymbol{\mu}_u\mu_d) \\ +(\mathbf{w}^T - (\mathbf{R} + \boldsymbol{\mu}_u\boldsymbol{\mu}_u^T)^{-1}(\mathbf{p} + \boldsymbol{\mu}_u\mu_d))^T \\ (\mathbf{R} + \boldsymbol{\mu}_u\boldsymbol{\mu}_u^T)(\mathbf{w} - (\mathbf{R} + \boldsymbol{\mu}_u\boldsymbol{\mu}_u^T)^{-1}(\mathbf{p} + \boldsymbol{\mu}_u\mu_d)). \end{aligned} \tag{12}$$

Using optimal filter coefficients $\mathbf{w}$ [5], equation (12) reduces to

$$\begin{aligned} J_{min} = \sigma_d^2 + \mu_d^2 - (\mathbf{p} + \boldsymbol{\mu}_u\mu_d)^T(\mathbf{R} + \boldsymbol{\mu}_u\boldsymbol{\mu}_u^T)^{-1}(\mathbf{p} + \boldsymbol{\mu}_u\mu_d) = \\ \frac{1}{N\mu_u^2 + \sigma_u^2}(\mu_d\sigma_u - \sigma_d\mu_u)^2 \end{aligned} \tag{13}$$

which represents the MMSE. From equation (13) it can be followed that the MMSE is decreasing if the number of coefficients N is being increased.

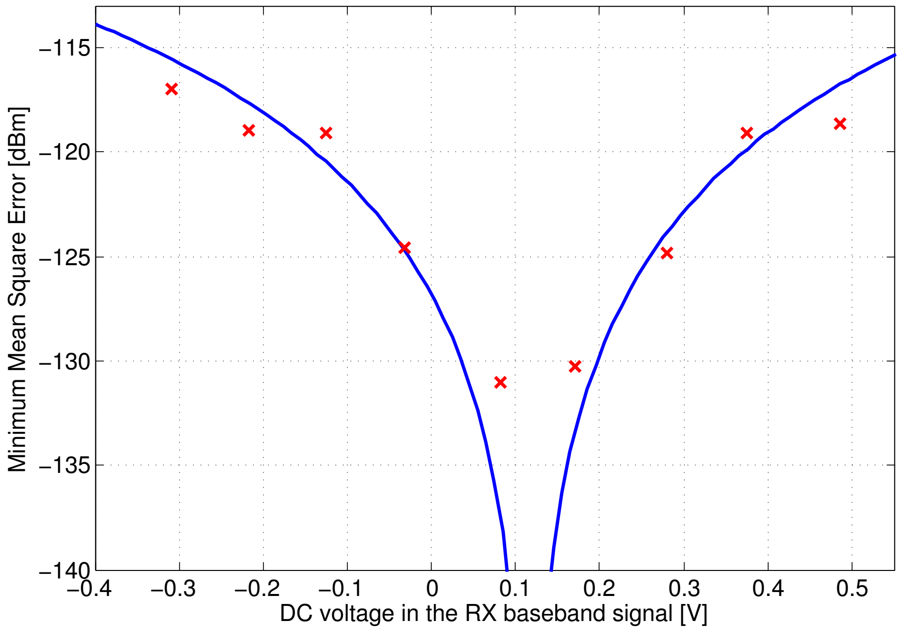

Fig. 3. DC dependent Minimum Mean Square Error (MMSE) of the cancellation

IV. Measurements and post-processing

To evaluate the DC offset dependency of the cancellation, signals measured at the DFE output of a real LTE transceiver chip with a signal bandwith of 5MHz have been postprocessed. To be independent of the convergence behavior and the misadjustment of the adaptive filter algorithm, 8 filter coefficients, represented by the vector $\mathbf{w}$, have been estimated by solving the Wiener-Hopf equation [5]. The DC voltage in the RX signal has been changed by using different bias currents in the downconverter. The solid line in Fig. 3 shows the theoretical MMSE for different DC voltage levels in the RX signal, according to (13), which also represents the power level of the remaining IMD2 after cancellation. The red points in Fig. 3 represent the results of the measured signals which have been postprocessed with the Wiener filter. The noise in the measured signals have been calculated from the Error Vector Magnitude (EVM) to a known wanted signal in the RX signal. The difference between the noise power of the signal after cancellation and a signal without IMD2 results in the power of the remaining IMD2. This is possible since the EVM calculation removes the known wanted signal and the remaining noise components of those signals are uncorrelated.

The zero in the curve of the theoretical MMSE represents the case when there are no other DC components contained in the RX signal, except the one from the IMD2. This has also been called the matched case before. The deep notch cannot be seen in the measured results, since the bias currents of the mixer could only be set in discrete steps, so it was not possible to achieve this matched case exactly. Nevertheless, it can be seen that the remaining IMD2 of the measured and postprocessed signals are closely following the shape of the theoretical MMSE.

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