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Efficient adaptive compensation of I/Q imbalances using spectral coherence with monobit kernel

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ABSTRACT

This paper shows a new algorithm to improve the performance of IQ demodulators and frequency converters exhibiting gain and phase imbalances between their branches. This algorithm does not require any input calibration signal and is independent of the input signal level. It exploits the spectral coherence (SC) concept using a monobit kernel to achieve optimization targets with minimum time to convergence, low computational load, and a wide range of input levels. Its effectiveness is shown through a low-IF receiver that improves its image rejection ratio (IRR) from 30 dB to 60 dB.

1. Introduction

IQ (Cartesian) demodulators and frequency converters are currently widely used in wireless communication systems and circuits, such as low-IF (low intermediate frequency) receivers and power amplifier linearization loops [1,2].

Low-IF receiver technology provides important radio receiver hardware simplification advantages compared to other technologies and architectures. However, the most important drawback of low-IF technology is its inherent low image rejection ratio (IRR) performance as a consequence, in some way, of imperfections of the IQ down-converters used in these systems.

The effects of branch imbalances of IQ demodulators and down-converters on image band interference in radio receivers are shown in Fig. 1. After down-conversion, part of the signal located in the image band of the receiver overlaps the signal located in the desired band.

The relation of radio receiver IIR to imperfections of its frequency down-converter is given by [3]

$$\mathrm{IIR} = \frac{1 + (\epsilon^2/4 - 1)\sin^2(\Delta/2)}{1 + (\epsilon^2/4 - 1)\cos^2(\Delta/2)} \tag{1}$$

where ϵ is the gain imbalance and Δ, given in radians, is the phase imbalance between the IQ demodulator or frequency down-converter branches of this device. For instance, a 3% gain imbalance and a 0.1 rad phase imbalance between the converter branches provides an IRR of only 25.6 dB, which is clearly insufficient to fulfill the needs of present-day receivers and communication systems. As this effect cannot be easily compensated by hardware improvements because of technology limitations and cost constraints, new digital signal processing algorithms could be useful for this task.

During the past few years, significant research efforts have been carried out to achieve a solution for low IRR figures caused by IQ demodulators and frequency down-converter imbalances (especially in low-IF receiver applications). Some authors have used adaptive techniques to try to solve this problem. For example, the system proposed in [4] treats the image signal as an interference and tries to reduce it by means of an adaptive interference canceler. In this case, the system performance is degraded by the presence of the desired signal in the image band. The system proposed in [5] uses time correlation techniques to improve IRR, but convergence of this algorithm strongly depends on the level of the input signal. Other solutions, such as the one proposed in [6], need the system to be calibrated with a reference input signal while the algorithm converges.

In this paper a new adaptive algorithm to improve the performance of IQ demodulators and frequency converters is proposed that does not need any input calibration signal, is independent of the input signal level, and requires only modest computational resources. It is based on the compensation cell shown in [6]. However, in this paper, the parameters of the compensation cell are updated by estimating the correlation between desired and image bands, using the spectral coherence (SC) concept, and consequently is independent of input level because of the SC mathematical properties. Also, the use of the monobit kernel [7] in the estimation of the SC reduces the computational load of the proposed algorithm.

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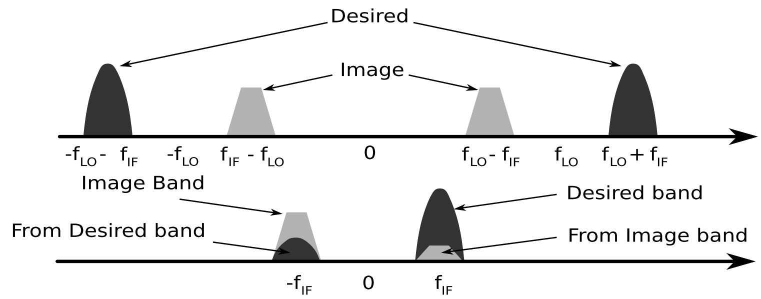

Fig. 1. Image band interference in a low-IF receiver caused by IQ frequency down-converter imbalances.

2. Algorithm strategy

As mentioned in Section 1, the algorithm proposed in this paper uses the SC estimator and a monobit kernel. The SC is defined as [8]:

$$C_{xy} = \frac{\gamma_{xy}(f)}{\sqrt{\gamma_x(f)\gamma_y(f)}} \tag{2}$$

where $\gamma_{xy}(f)$ is the cross spectral density at frequency f between $x(t)$ and $y(t)$ signals with auto spectra $\gamma_x(f)$ and $\gamma_y(f)$. The SC is a complex number that measures correlation between two signals in the frequency domain. It ranges from 0 to 1 and is level-independent [8].

In this paper, the SC is estimated using:

$$\tilde{C}_{xy} = \frac{\sum_{n=1}^{N} X_n(f)Y_n^*(f)}{\sqrt{\sum_{n=1}^{N} |X_n(f)| \sum_{n=1}^{N} |Y_n(f)|}} \tag{3}$$

where $X_n(f)$ and $Y_n(f)$ are spectral estimates of the input signals, which can be obtained using the windowed fast Fourier transform (FFT) or a bank of filters. Note that (3) is similar but not equal to the magnitude-squared coherence (MSC) [8], which is the most commonly used SC estimator, and is defined as $|\tilde{C}_{xy}(f)|^2$.

As formulated in (3), the SC estimates the correlation between the spectral components of two signals. These spectral components must be at the same frequency in order to estimate the SC; for other cases, the cyclic-correlation should be obtained [9].

Given that the signal at the output of the converter is complex, the SC concept is used to estimate the correlation between the positive and negative frequency range of the input signal spectrum. Because after conversion, the desired and image bands of the signal are symmetrical about zero frequency, as shown in Fig. 1, it is not necessary to extend the SC estimation to the whole frequency range, but can be reduced to just the desired and image bands.

The use of a filter bank (such as the DFT) to estimate the SC can result in degraded system behavior since the filter frequencies are governed by the sampling frequency and the number of points of the DFT. In this situation, it is unlikely that the signal is centered within a filter, spreading its energy over two or more filters. To overcome this problem, two bandpass finite impulse response (FIR) filters tuned at the desired and image bands will be used instead of a filter bank implemented using the DFT. This provides an important computational load reduction compared to previously published algorithms [5,6]. It must be noted that the number of filter taps determines the bandwidth used to estimate SC. This bandwidth must be equal to the bandwidth of the desired signal to reduce the output noise. The SC estimator can be rewritten as

$$\tilde{C}(f_{IF})_x = \frac{\sum_{n=1}^{N} X_{desired}(f_{IF})X_{image}^*(-f_{IF})}{\sqrt{\sum_{n=1}^{N} |X_{desired}(f_{IF})| \sum_{n=1}^{N} |X_{image}(-f_{IF})|}}. \tag{4}$$

The quality of the SC estimate and the algorithm convergence speed will be reduced for low signal-to-noise ratio (SNR). This problem can be solved increasing N, the number of averaged estimates in (4), as is done for other spectral estimation algorithms. Increasing N improves the SC estimation quality but also increases the convergence time of the algorithm. Even though a trade-off between SC estimation quality and convergence time is required for most situations. N can be set to a minimal value large enough to have suitable SC quality, while simultaneously being small enough to achieve short convergence time for moderate-to-high SNR.

The computational load imposed by this algorithm can be further reduced if the coefficients of the FIR filters mentioned above are quantized using a monobit kernel, which was proposed by J.B. Tsui [7] to simplify the hardware of a digital channelized receiver. Given that the coefficients of the FFT are complex, they cannot be represented by a 1-bit real number. It is necessary to use 1-bit for the real part and 1-bit for the imaginary part, and this 2-bit representation is called a *monobit*. The possible values are $+1, -1, +j$ and $-j$, and the quantized coefficient depends on the true coefficient argument. This can be written as [7]:

$$Q\left[\arg\left(e^{\frac{-j2\pi kn}{N}}\right)\right] = \begin{cases} 1, & -\pi/4 \leqslant \frac{-j2\pi kn}{N} < \pi/4, \\ j, & \pi/4 \leqslant \frac{-j2\pi kn}{N} < 3\pi/4, \\ -1, & 3\pi/4 \leqslant \frac{-j2\pi kn}{N} < 5\pi/4, \\ -j, & 5\pi/4 \leqslant \frac{-j2\pi kn}{N} < 7\pi/4. \end{cases} \tag{5}$$

With these simplified quantization values, the multiplications required by the FFT computation are not necessary and only sums are needed. This helps to simplify the hardware of the system and reduce its computational load. In this paper, the monobit kernel is used to quantize the FIR filter coefficients, which noticeably reduces the computational burden of the algorithm.

The FIR filters used by the proposed algorithm implement a rectangular window. The low-pass response of this filter is converted to bandpass form by multiplying its coefficients by a complex exponential function whose frequency is equal to the center of the desired and image bands. Thus,

$$w[n] = \begin{cases} e^{j\omega_0 n}, & 0 \leqslant n \leqslant M, \\ 0, & \text{otherwise} \end{cases} \tag{6}$$

where ω_0 is the center frequency of the filter and M is the number of filter taps. Then applying (5) to the coefficients obtained by (6), these are converted to a monobit kernel representation, avoiding the use of multiplications.

3. Algorithm description

A block diagram of the proposed algorithm is shown in Fig. 2. The input signal $s(t)$ to the IQ demodulator is down-converted, low-pass filtered, and converted from analog to digital in Block I. The IQ (Cartesian) signal at the output of Block I is fed to the compensation cell of Block II. This compensation cell implements the equations [6]: