Improving Harmonic Rejection for Spectrum Sensing using Crosscorrelation

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Abstract—Harmonic downmixing may cause a spectrum sensing device to erroneously flag parts of the spectrum as occupied. Receivers employing harmonic rejection (HR) rarely obtain more than 60 dB of HR, which may not be enough for spectrum sensing for dynamic spectrum access. We improve HR by employing two RF-receivers, integrated in a single 65nm CMOS IC, and using a small offset in their LO-frequencies. The offset is corrected in the digital baseband, after which the final receiver outputs are crosscorrelated. Our measurements show that the HR can be significantly improved without calibration or time penalty, even for high initial HR, but that the improvement is limited by crosstalk between the receivers.

Index Terms—cognitive radio, crosscorrelation, energy detection, harmonic rejection, image rejection, spectrum analyzer, spectrum sensing

I. INTRODUCTION

The current practice of static allocation of frequency bands in the EM-spectrum is very inefficient, as most of these allocated bands are unused at any give time and location [1]. This can be solved by dynamic spectrum access to allow the opportunistic use of available frequencies. To maximize the detection of unused spectrum, wideband operation and accurate sensing are required. Most integrated wideband receivers use a zero-IF architecture with digital local oscillators (LOs) and switching mixers for wideband frequency generation and high linearity. Such mixers also downconvert signals at a multiple of the LO-frequency, which is known as harmonic downmixing.

Several solutions to improve the harmonic rejection (HR) have been presented in literature. RF tracking filters can suppress signals present at those harmonics, but are bulky, lossy and noisy. A HR-mixer suppresses several harmonics, but typically only by 40 dB due to mismatch [2]. More has been achieved by two-stage HR, obtaining 60 dB HR [3]. Digital algorithms have shown to improve the suppression of some individual harmonics up to 80 dB [3], [4]. This paper proposes a technique to realize additional HR for *all* harmonics by employing two receivers in combination with a frequency offset and crosscorrelation (xc) [5], [6].

II. IMPROVED HARMONIC REJECTION

Fig. 1 is used to explain the concept to improve HR for spectrum sensing. The RF input signal $s_{\text{RF}}(t)$ is assumed wide-sense stationary and may contain signals over a very wide range of frequencies. It is processed by two identical receivers, each of which performs amplification (not shown in Fig. 1),

Fig. 1. System architecture to improve HR by xc.

mixing, filtering and AD-conversion. The digital outputs of the receivers are combined with some digital signal processing (DSP) to obtain a spectrum estimate.

In the upper receiver, the LO has a frequency f_1 , while in the lower receiver, the LO has frequency $f_2 = f_1 + \Delta f$. For now, we focus on the first (upper) receiver. An IF output signal $s_{\text{IF}_1}(t)$ is obtained at the output of the mixer. Passive mixers are often used for their high linearity, but they require a square-wave LO-signal $s_{\text{LO}_1}(t)$, which does not only contain the fundamental frequency f_1 , but also higher harmonics:

$$s_{\text{LO}_1}(t) = \sum_{k=-\infty}^{\infty} c_k e^{-j2\pi k f_1 t} \quad (1)$$

where c_1 is the desired coefficient (without loss of generality we define $c_1 = 1$). The other c_k are the weight factors for the harmonics and lead to harmonic downmixing, as shown at the output of the ADC in Fig. 1. Particularly, $c_{-1} \neq 0$ leads to finite image rejection (IR).

For convenience, we define $s_{\text{RF}}(t)$ as a superposition of non-overlapping baseband-equivalent signals $z_k(t)$, centered around the LO-harmonics (if $s_{\text{RF}}(t)$ is real, $z_{-k}(t) = \overline{z_k(t)}$):

$$s_{\text{RF}}(t) = \sum_{k=-\infty}^{\infty} z_k(t) e^{j2\pi k f_1 t} \quad (2)$$

with $P_k = \mathbb{E}[|z_k(t)|^2]$ the power of $z_k(t)$, and $\hat{P}_k$ the estimated power of $z_k(t)$. $\hat{P}_k$ and P_k will differ due to the mea-

surement variance, but will converge for infinite measurement time (neglecting noise contributions of the receiver). Assuming direct-conversion, the mixer output is

$$s_{\text{IF}_1}(t) = \sum_{k=-\infty}^{\infty} c_k z_k(t) + \text{higher frequency components} \quad (3)$$

where the higher frequency components are filtered out by the filter. After the ADC, the power in the received band would be estimated as (in Fig. 1, this is calculated when $\Delta f = 0$)

$$\begin{aligned} \hat{P}_{\text{1rev}} &= \frac{1}{N} \sum_{n=0}^{N-1} |s_{\text{IF}_1}[n]|^2 = \frac{1}{N} \sum_{n=0}^{N-1} \sum_{k=-\infty}^{\infty} |c_k z_k[n]|^2 \\ &= \sum_{k=-\infty}^{\infty} |c_k|^2 \hat{P}_k \end{aligned} \quad (4)$$

with N the number of samples (per receiver). Here we have assumed that all z_k are uncorrelated and thus add in power. The estimated power highly depends on signals present at harmonics of the LO. Filtering to reduce z_k before down-conversion will help, but is usually insufficient and difficult to implement for wideband RF systems. Therefore, receivers aim to reduce c_k through a differential implementation ($c_k = 0$ for all even k) and HR-mixers ($c_k = 0$ for some odd k). Due to mismatches, the odd harmonics are typically only suppressed by around 40 dB ($|c_k/c_1| \approx 0.01$), denoted as $\text{HR}_k = 40$ dB.

Using two receivers and a frequency offset can improve this. At the output of the second mixer, we find the signal $\sum_{k=-\infty}^{\infty} c_k z_k(t) e^{-j2\pi k \Delta f t}$. Multiplying this by $e^{j2\pi \Delta f t}$ (implemented in the digital domain, as it should not suffer from additional harmonic downconversion, see Fig. 1) results in the output of the second receiver

$$s_{\text{IF}_2}(t) = \sum_{k=-\infty}^{\infty} c_k z_k(t) e^{-j2\pi(k-1)\Delta f t}. \quad (5)$$

The power in the received band can be estimated by the xc of the two receiver outputs (note $t = n/f_s$) [7]

$$\begin{aligned} \hat{P}_{\text{xc}} &= \left| \frac{1}{N} \sum_{n=0}^{N-1} s_{\text{IF}_1}[n] \overline{s_{\text{IF}_2}[n]} \right| \\ &= \left| \frac{1}{N} \sum_{n=0}^{N-1} \sum_{k=-\infty}^{\infty} |c_k z_k[n]|^2 e^{j2\pi(k-1)\Delta f \frac{n}{f_s}} \right| \\ &= \hat{P}_1 + \frac{1}{N} \sum_{k=-\infty, k \neq 1}^{\infty} |c_k|^2 \hat{P}_k \left| \sum_{n=0}^{N-1} e^{j2\pi(k-1)\Delta f \frac{n}{f_s}} \right| \\ &\triangleq \hat{P}_1 + \epsilon \end{aligned} \quad (6)$$

where ϵ is the error made in the estimation, with

$$|\epsilon| \leq \frac{1}{N} \sum_{k=-\infty, k \neq 1}^{\infty} |c_k|^2 \hat{P}_k. \quad (7)$$

For large N , $\epsilon \rightarrow 0$, so $\hat{P}_{\text{xc}} \rightarrow \hat{P}_1$, and thus $\hat{P}_{\text{xc}}$ can be used as an estimate for P_1 . This means, the power of $z_1(t)$ can be found without contamination from harmonic downmixing.

Fig. 2. Efficient DSP of xc to improve HR using a frequency offset. In this example it is assumed that $\Delta f = B/M$ (1 FFT-bin).

Note that we have assumed that the bandwidth of $z_1(t)$ is less than $B - |\Delta f|$, with B the bandwidth of the filter + ADCs, such that after the frequency shift in the digital domain, z_1 from the two receiver paths can still fully overlap.

The resolution bandwidth (RBW) can be easily changed in the digital domain by using a (windowed) M -point fast Fourier transform (FFT) to divide the bandwidth B into M subbands of B/M Hz, see Fig. 2, allowing low-cost flexible spectrum sensing. If Δf is chosen as kB/M , with $k \in \mathbb{Z}$, $|k| < M$, the multiplication with $e^{j2\pi \Delta f t}$ can be implemented as a shift in FFT-bins with negligible processing overhead.

When not used for spectrum sensing, the second receiver can simply be turned off, used to simultaneously receive a second band, or used for HR-schemes that are compatible with demodulation, such as proposed in e.g. [8]. Note that many MIMO systems already have multiple receivers, in which case the system can be reconfigured for spectrum sensing without adding hardware.

III. IMPLEMENTATION

To demonstrate the feasibility of integration of this HR-technique using xc, a prototype was fabricated. Integration of two receivers operating at different frequencies introduces crosstalk, and this is most likely to occur at RF and via the LOs. Therefore (and to allow more experimental freedom at IF), only the RF-parts of the two receiver frontends are integrated in one IC, see Fig. 3. It was presented in [9], but without the concept of frequency offset.

The RF-frontend consists of two identical channels, with each channel based on [3]; the complete frontend is detailed in [9]. A low-noise transconductance amplifier (LNTA) transforms the input voltage into a 2:3:2 ratio of output currents, which are mixed down by a passive mixer. The LNTA can switch between 50Ω and 100Ω input impedance to provide matching in single-receiver mode as well. A discrete-step attenuator precedes the LNTA for higher linearity, but is left unused here. The mixer is driven by an 8-phase LO with a duty cycle of $\frac{1}{8}$ [3]. A transimpedance amplifier (TIA), implemented externally with a TI-THS4130 opamp with RC-feedback, provides voltage gain simultaneously with low-pass filtering. This introduces the first voltage amplification at IF, where large voltage swings can be better handled. The signal is further amplified by broadband amplifiers to interface with 5th-order passive filters and 14-bit differential ADCs (PMC66-

Fig. 3. Implementation overview, with (a) system diagram and (b) chip micrograph

14HSA14) sampling at 10 MS/s. The DSP is performed in software on a PC using double precision. At the output of the multiply-accumulates (MACs), the magnitude is taken for robustness to phase mismatch between the receivers [7].

The LO-generating circuitry employs a divide-by-8 with a frequency range of 0.3–1.0 GHz (2.4–8.0 GHz in). In combination with the 2:3:2 ratio of the LNTA-outputs, this provides 30–35 dB of HR for the 3rd and 5th harmonic for a single channel. Everything is differential to suppress even harmonics. Δf is generated externally using two locked signal generators, see Fig. 3a. The ADC sample rate could not be locked to the external signal generators, so the frequency offset correction was estimated in the DSP and then corrected for (the shift in FFT-bins as discussed in section II was not used).

IV. MEASUREMENT RESULTS

The measurement of HR_k ($k \neq 0, 1$) at certain f_{LO} was carried out with the following settings: $f_{CLK,1} = 8f_{LO}$, $f_{CLK,2} = f_{CLK,1} - 1.6$ MHz (thus $\Delta f = -200$ kHz), $f_{in,1} = f_{LO} + 1.2$ MHz ($P_1 = -90$ dBm), $f_{in,k} = kf_{LO} + 1.9$ MHz ($P_k = -20$ dBm). Flattop windows were used to accurately estimate the power of the sinusoids and to have good sidelobe suppression. The measured power at 1.2 MHz IF was set to equal the input power of -90 dBm, and the HR_k can then be found by evaluating the power of the downconverted harmonic with its original input power. For IR, only one input tone $f_{in,1} = f_{LO} + 1.2$ MHz at -50 dBm was inserted, with the image power measured at -1.2 MHz IF. In the following figures, the frequency offset correction for receiver 2 has already been applied. This means that for receiver 2, all signals have moved 200 kHz to the left.

An example measurement for IR is shown in Fig. 4 (only part of the usable baseband spectrum shown). Here xc obtains 48 dB IR, about 20 dB better than the individual receivers. The only signals present in the baseband of receiver 1 should be at ± 1.2 MHz (the desired signal and its image) and around DC due to DC-offset, but it can be observed that there are also (weaker) signals present at ± 1.4 MHz and ± 1.6 MHz. Similarly, the output of receiver 2 shows signals at DC and ± 1.4 MHz (desired signal and image), but also at ± 1.2 MHz. After this frequency correction, the desired signal lines up at

Fig. 4. Example measurement for IR at $f_{LO} = 400$ MHz. The image of receiver 1 is at -1.2 MHz with 30 dB rejection, the image of receiver 2 is at -1.6 MHz with 28 dB rejection, while the maximum spur due to images in the xc-spectrum is at -1.4 MHz with 48 dB rejection.

1.2 MHz. The undesired line-up of spurs at -1.4 MHz (due to crosstalk) limits the IR-improvement using xc here to 20 dB.

An example measurement for HR_3 is shown in Fig. 5. The output spectra of the individual receivers have significantly higher spurs due to harmonic downmixing than the spectrum obtained using xc. For example, the signal at the third harmonic folds down to -1.9 MHz in receiver 1 and to -2.7 MHz in receiver 2 with a power of -48 dBm. Thus, with an input power of -20 dBm, this results in $HR_3 = 28$ dB. For xc, the strongest remaining peak is -70 dBm at -2.7 MHz, resulting in $HR_3 = 50$ dB, an improvement of 22 dB. This improvement is obtained *immediately*, as we did not perform xc over more than 1 FFT. A longer measurement does not lower this highest peak as it is correlated between the receivers.

Without crosstalk, the immediate HR-improvement would be higher, and further improved by increasing the measurement time (see also (6)). The crosstalk mechanism needs further study, but we think it is caused by the low-ohmic connection of the grounds of the two receivers via the padding. Nevertheless,

Fig. 5. Example measurement for HR_3 at $f_{LO} = 300$ MHz. The signal $f_{in,3}$ of -20 dBm at 901.9 MHz mixes down to -1.9 MHz for receiver 1 and to -2.7 MHz for receiver 2 (after frequency correction), both at -48 dBm for a $HR_3 = 28$ dB. In the xc-spectrum (lowest part, with output spectra of the two individual receivers indicated in light gray), the highest spur due to harmonic downmixing is at -2.7 MHz at -70 dBm, giving $HR_3 = 50$ dB.

Fig. 6. Measured HR over frequency for a single receiver (gray lines) and with xc using two receivers (black lines) for several relevant harmonics. The measurement setup limits the RF input frequency to below 3.5 GHz, thus e.g. the 7^{th} harmonic is only measured up to 500 MHz.

even with crosstalk, a significant gain in HR is obtained.

We measured HR_k for different f_{LO} and k , the results of which are shown in Fig. 6. For a single receiver, we have taken the average in dB of the HR_k of the two individual receivers, as the HR of the two receivers matches quite well. We have consistently taken the strongest spur present to determine HR_k . The IR decreases with frequency, because the spur at -1.6 MHz (also visible in Fig. 4) increases with frequency (likely due to the crosstalk mechanism). The 4^{th} harmonic is already 70 dB rejected in the individual receivers by the differential implementation, but xc still improves HR_4

to 80 dB, *without* calibration. With xc, HR_7 (the 7^{th} harmonic is not suppressed by the HR-mixer) improves from 25 dB for the individual receivers to 45 dB, an improvement of 20 dB. For other harmonics, such as the image, the 3^{rd} and the 5^{th} , the improvement obtained using xc ranges from 5 dB to 25 dB.

V. CONCLUSIONS

In this work, a concept to improve HR for spectrum sensing is experimentally verified with two receivers integrated in one IC. The technique uses xc of the outputs of two receivers with a frequency offset to decorrelate the harmonic images that are downconverted by the mixers. Significant improvements of up to 25 dB in HR have been observed without requiring additional measurement time. The improvement in HR is limited by crosstalk due to sub-optimal layout; without the crosstalk, the improvement could be much more. Nevertheless, the technique discussed in this paper gives significant improvements without requiring any calibration, and is a step forward towards an integrated spectrum analyzer with HR.

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