5.3 A Phase-Noise and Spur Filtering Technique Using Reciprocal-Mixing Cancellation

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Recent passive-mixer-based architectures, such as [1], have shown that blockers as large as 0dBm can be tolerated without excessive *gain compression*. However, even in a *perfectly linear* receiver, *reciprocal mixing* of the blocker with LO phase noise deposits additive noise on the wanted signal, as shown in Fig. 5.3.1. This is an inevitable limitation to any *mixer-based* receiver. Assuming that the blocker experiences no passive RF filtering, the noise figure of such a receiver in the presence of a given blocker is only lowered *through improving the LO phase noise*. To overcome this challenge, most wireless receivers use *LC-oscillators* that despite their superior phase noise to ring oscillators [2], still consume a large portion of the radio power. As the quality factor of on-chip resonators does not scale with technology, the phase noise of an LC-oscillator can only be improved by consuming more power, while the benefits of circuit innovation are fundamentally limited [3]. Although increasing current helps, it does come at a cost, and is ultimately limited by the maximum allowable amplitude, and how reliably small inductor values can be fabricated without Q-degradation. In this paper, we propose a mixed-signal reciprocal-mixing cancellation technique that leads to a substantial reciprocal-mixing noise figure improvement *independent of the LO phase noise*.

The proposed technique makes use of phase noise symmetry around the carrier, a property of *narrowband phase modulation*. As shown in Fig. 5.3.1, an LO with the *symmetrical phase noise* profile mixed with a narrowband blocker, yields a symmetrical reciprocal-mixing profile around the blocker beat frequency Δf_b. The reciprocal-mixing tail around DC is in-band, and indistinguishable from the desired signal. By contrast, if the image of the in-band reciprocal-mixing located at $2\Delta f_b$ is frequency-shifted by $2\Delta f_b$ and properly scaled to match the in-band reciprocal-mixing, a simple subtraction results in perfect cancellation. To protect the desired signal, an auxiliary path is needed to create the reciprocal-mixing replica. After proper subtraction, reciprocal-mixing is replaced by the noise of the auxiliary path which *is independent of blocker power and offset, as well as LO phase noise*.

A block diagram of the main and auxiliary paths is shown in Fig. 5.3.2. The output of the receiver downconversion mixer consists of the desired signal at DC, the blocker at Δf_b, any in-band spur on top of the desired signal, and its image around $2\Delta f_b$. A low-pass/high-pass arrangement splits the spectrum at the mixer output between the main and auxiliary paths. The low-frequency spur and the desired signal pass through the main path while the blocker and the spur image are diverted to the auxiliary path. As illustrated in Fig. 5.3.2, the spur image can be treated as a single-sideband signal with respect to the downconverted blocker, which can then be represented as the sum of *equal AM sidebands and PM* sidebands. When applied to *a limiter*, the AM sidebands are rejected, but the PM sidebands are preserved. Therefore, at the limiter output, the in-band spur, less the receive signal has been restored. The limiter is followed by an LPF to reject the blocker and the spur image. The output is properly scaled with a variable gain stage before the final subtraction. In practice, the scaling factor G in Fig. 5.3.2 should be *complex* to compensate for phase shifts from HPF and LPF. This is conveniently implemented in digital by combining the quadrature baseband signals with the appropriate weights. The weights are adaptively computed to minimize the cross-correlation between the auxiliary path output and the final output after subtraction.

Although the example in Fig. 5.3.2 is considered reciprocal-mixing due to a *spur*, the result is equally valid for reciprocal-mixing caused by *phase noise*. The only difference in the case of phase-noise generated reciprocal-mixing is the folding of higher-order images caused by limiter odd-order nonlinearities. As shown in Fig. 5.3.3, the reciprocal-mixing spectrum after HPF is applied differentially to the limiter. The output of the limiter has the fundamental reciprocal-mixing around Δf_b with the in-band reciprocal-mixing restored from the image at $2\Delta f_b$, and its replicas at odd harmonics. The reciprocal-mixing skirts of the higher-order replicas cause an in-band distortion which is not correlated with reciprocal-mixing of the main path. This harmonic distortion limits the cancellation to about 9.5 dB mainly dominated by the third harmonic. Since most blocker-tolerant receivers inevitably use an 8-phase architecture [1], this issue is mitigated by exploiting the 8-phase mixer along with a harmonic recombination circuit (Fig. 5.3.3). Consequently the 3rd and 5th harmonics are cancelled leaving the 7th as the first dominant one, which in theory is 16.9dB below the in-band reciprocal-mixing because of oscillator $1/f^2$ roll-off, but in practice 29dB rejection is measured due to filtering at the mixer output and the internal nodes of the limiter.

For the proof of concept, the receiver in Fig. 5.3.4 was implemented in 40nm CMOS. The main path consists of an inverter-based TIA with a feedback resistor. The output of the TIA is filtered using an on-chip passive RC filter, followed by digitally controlled variable-gain transconductors, whose output currents are combined with those of the auxiliary path to perform both harmonic rejection and reciprocal-mixing cancellation. At the input of the main path TIA a grounded bypass capacitor is commonly used to attenuate the out-of-band blockers. In our architecture, instead of steering the out-of-band current to ground, it is directed towards the auxiliary path. A TIA at the input of the auxiliary path converts the input current into voltage, followed by a hard-limiting inverter stage to ensure proper restoration of the in-band reciprocal mixing. The output is then low-pass filtered to attenuate the blocker, before it is applied to a digitally-controlled variable gain transconductor, which is a replica of the one used in the main path.

The P and N transistors in the inverter-based auxiliary path TIA are sized properly to equalize the transconductance of both devices and hence minimizes the even-order nonlinearity, while the current is optimized for noise performance The value of the auxiliary path TIA feedback resistor is critical for proper operation, it should be large enough to maximize the gain and relax the noise requirement of subsequent stages, but not too large to increase the TIA input impedance or limit the TIA bandwidth, which should be wide enough to pass the reciprocal-mixing image at $2\Delta f_b$. The auxiliary path including the TIAs and the limiters consumes a total of 8mA. The total active area of the LNA, mixers, main paths and auxiliary paths is $1.4mm^2$ (Fig. 5.3.7).

To evaluate the performance, the receiver is tuned to 2.3GHz and a CW blocker is applied at 20MHz away, while a spur at 20.2MHz is injected to the LO. The output spectrum of the receiver with and without cancellation is shown in Fig. 5.3.5. The spur is effectively filtered by over 30dB. Using the same setup, the receiver noise figure versus blocker power is plotted in Fig. 5.3.6. The post-cancellation NF improves by up to 19dB. The receiver small signal NF is measured at 2.4dB, which increases to 9.5dB in the presence of a -15dBm blocker after cancellation, limited by the noise of the auxiliary path. To the first order, this noise figure is *independent of the blocker offset frequency or LO phase noise*. In practice the blocker offset can be as low as 10MHz in this design, set by the size of the HPF capacitor. This performance translates to an *inductorless VCO* with an FoM of 188dB, over 20dB higher than a typical ring oscillator, and comparable to an LC-based VCO. Alternatively, when a high quality LC VCO is used, this technique enhances the RX blocker tolerance substantially at a modest increase in its current. In the absence of blockers the auxiliary path can be turned off to save current without perturbing the LO phase continuity, which would be a consequence of reducing VCO power.

References:

[1] D. Murphy et al., “A Blocker-Tolerant Wideband Noise-Cancelling Receiver with a 2dB Noise Figure,” *ISSCC Dig. Tech. Papers,* pp.74-76, 19-23 Feb. 2012.
[2] A. Abidi, “Phase Noise and Jitter in CMOS Ring Oscillators,” *IEEE J. Solid-State Circuits*, vol.41, no.8, pp.1803-1816, Aug. 2006.
[3] A. Mazzanti and P. Andreani, “Class-C Harmonic CMOS VCOs, With a General Result on Phase Noise,” *IEEE* J. *Solid-State Circuits,* vol.43, no.12, pp. 2716-2729, Dec. 2008.

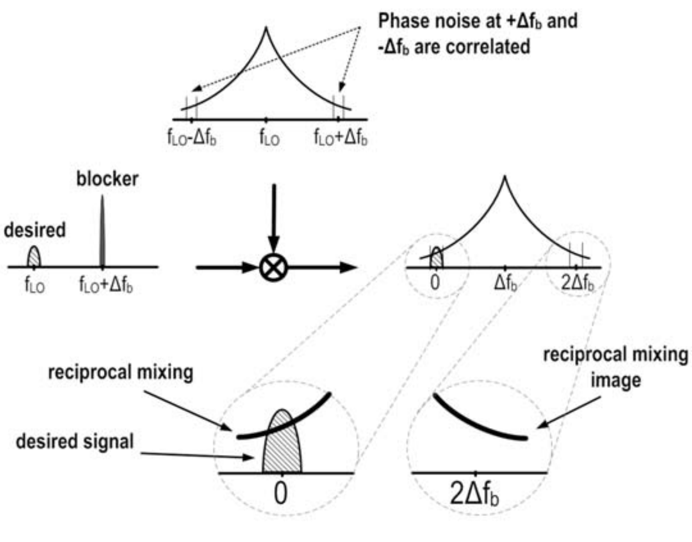

Figure 5.3.1: Generation of reciprocal-mixing and its image.

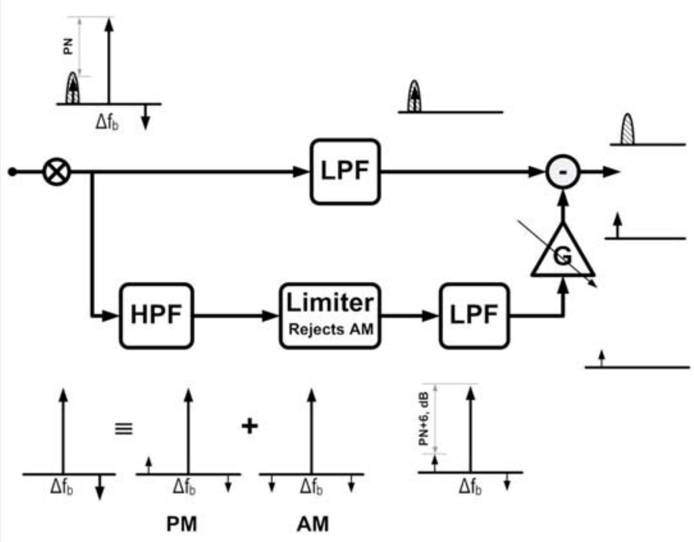

Figure 5.3.2: Proposed architecture for reciprocal-mixing cancellation.

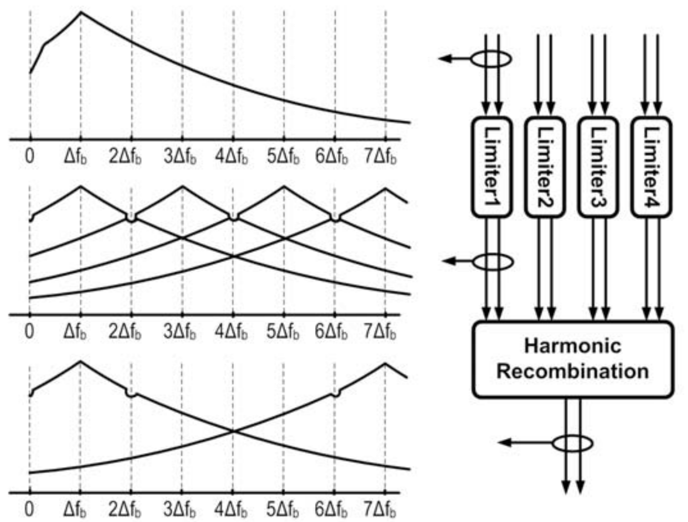

Figure 5.3.3: Rejection of limiter 3rd- and 5th- order nonlinearities using 8 phases.

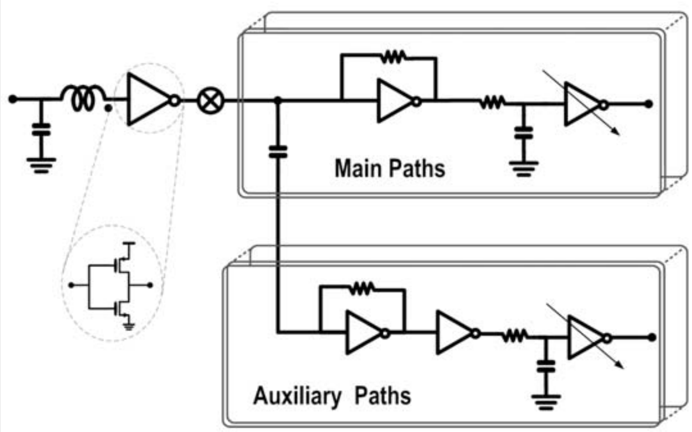

Figure 5.3.4: Receiver architecture with reciprocal-mixing cancellation.

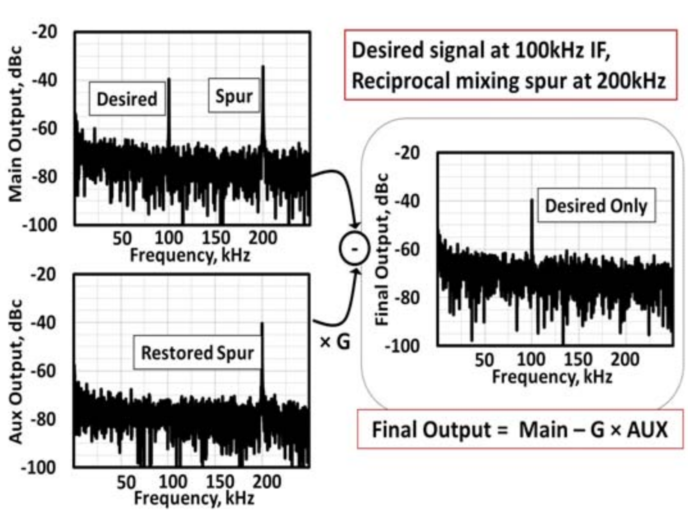

Figure 5.3.5: Measured output spectrum with and without cancellation.

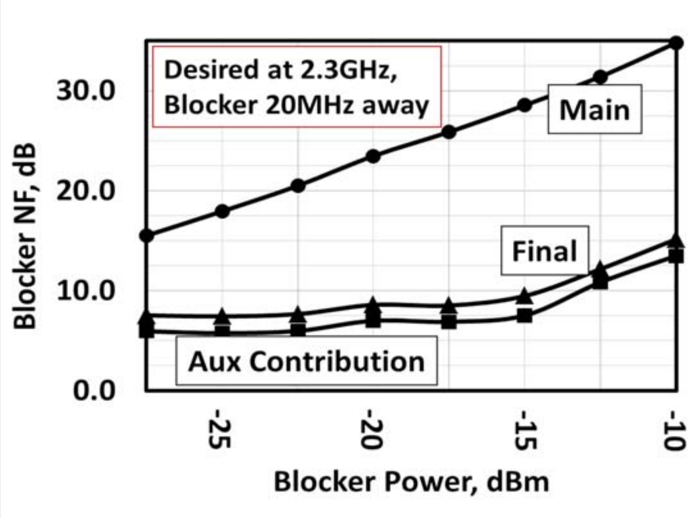

Figure 5.3.6: Measured noise figure versus blocker power.

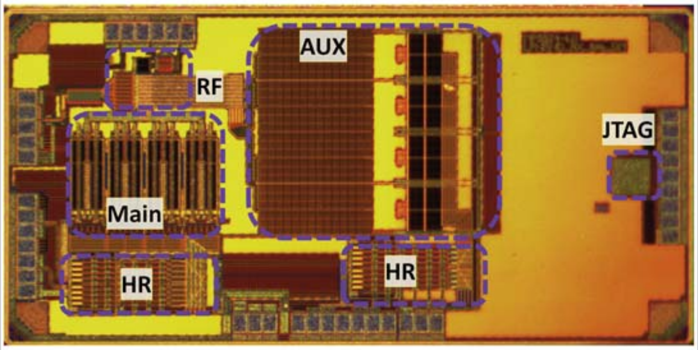

Figure 5.3.7: Die Micrograph.

