A Harmonic Rejection Strategy for 25% Duty-Cycle IQ-Mixers Using Digital-to-Time Converters

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***Abstract*—Due to the use of wide-band analog front-ends in order to cover the many different frequency bands used in Long Term Evolution (LTE), blocker signals received by the antenna may reach the mixer input. As a consequence, square-wave mixer implementations, as, e.g., the 25% duty-cycle current-driven passive mixer, may lead to the down-conversion of these blocker signals by their harmonic response. This contribution describes a harmonic rejection (HR) strategy, which modifies the local oscillator (LO) waveform of a 25% duty-cycle current-driven passive mixer, to suppress specific harmonics. Two HR control signal waveforms are proposed and their performance is evaluated by circuit simulations using a 28 nm CMOS technology. By applying the proposed concept, the down-conversion of a blocker by the mixer's 3rd order harmonic response can be suppressed by more than 30 dB, and the receiver noise figure (NF) is improved by 21.1 dB for a blocker power of −5 dBm at the low-noise amplifier (LNA) input.**

***Index Terms*—Mixer, receiver, harmonics, digital-to-time converter, wireless communications.**

I. Introduction

IN RADIO frequency (RF) transceivers mixers are used to shift the desired RF channel to the baseband (BB) or an intermediate-frequency, where the wanted signal is digitized for further digital signal processing. Frequency translation may be realized by a nonlinear operation, or, as in modern RF transceivers, with linear time-variant systems. In order to cover the high frequency range used by current communication standards as, e.g., in LTE, a wide-band frequency synthesizer is needed. In modern mobile transceivers, switched square-wave systems, as the 25% duty-cycle mixer architecture, are preferably used in direct-conversion in-phase (I) and quadrature-phase (Q) receivers [1]. It consists of four mixer switches (transistors) I_+, I_-, Q_+ and Q_- which are switched ON and OFF by a 25% duty-cycle scheme, such that at any moment only one switch is turned ON. The switch control signals $p_{I+}(\varphi)$, $p_{Q+}(\varphi)$, $p_{I-}(\varphi)$ and $p_{Q-}(\varphi)$ are visualized in Fig. 1 (a), where the variable substitution $p_{I+}(\varphi) = \tilde{p}_{I+}(\varphi T_{LO}/(2\pi))$ is used, with T_{LO} being the LO period.

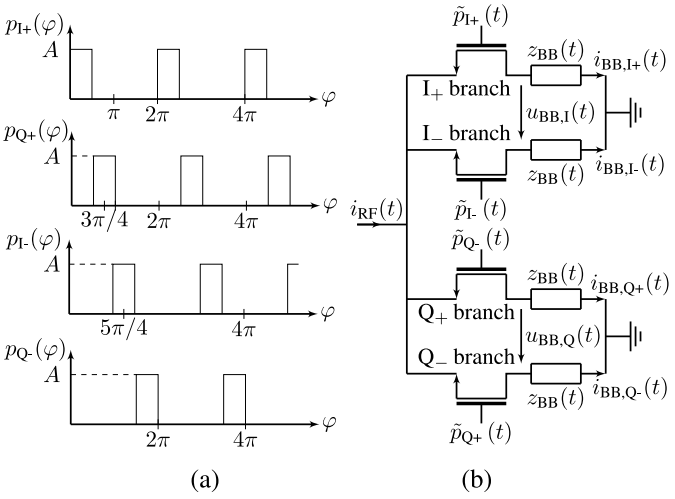

Fig. 1. Control signals (a) and analog circuit (b) of the 25% duty-cycle current-driven passive mixer.

The equivalent analog circuit of the 25% duty-cycle current-driven passive IQ mixer is depicted in Fig. 1 (b). The LNA output current $i_{RF}(t)$ flows into the four switching branches of the I- and Q-path. Each path contains a transimpedance amplifier (TIA) to generate the differential voltages $u_{BB,I}$ and $u_{BB,Q}$ which are digitized and combined to the complex IQ BB signal $u_{BB}[n] = u_{BB,I}[n] + ju_{BB,Q}[n]$. Traditional designs using 50% duty-cycle LO signals suffer by an IQ cross-talk problem due to the overlapping ON-states of the switches in the I- and Q-branch. This overlapping furthermore causes a reduced conversion gain because the RF input current is shared between two branches [2]. The 25% duty-cycle scheme offers a 3 dB higher conversion gain and a reduced noise-figure compared to 50% duty-cycle mixers [1], [3]. A well-known issue in square-wave mixers is the generation of harmonics in the mixer branches [4]–[6] which leads to the down-conversion

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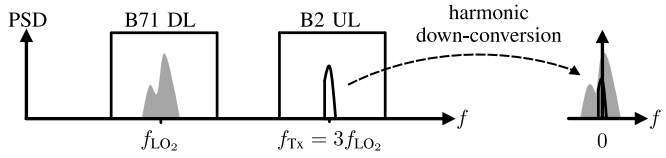

Fig. 2. Inter-band CA scenario with the down-conversion of the leaked Tx signal by the 3rd order harmonic response of the SCC mixer.

of unwanted spectral components contained in the RF input current $i_{RF}(t)$. To avoid this harmonic down-conversion by the harmonic response of the mixer, several approaches can be found in the existing literature. In [7], a 33% duty-cycle mixer for the rejection of the 3rd order harmonic response of the mixer is proposed. The HR mixers introduced in [5], [6] use an LO signal with suppressed harmonics by approximating the ideal sine wave by a 3-bit amplitude-quantized signal. The waveform is composed of three 45° shifted square-wave signals which are weighted and added. While this LO signal rejects the 3rd and 5th harmonic response simultaneously, amplitude- and phase mismatches result in a lowered suppression. In [4], this HR concept is used in IQ-transmitters, and in [8], [9] for receivers. Higher order harmonics may be rejected by using more than three square waves [6] or by using an 8-path HR mixer with 12.5% duty-cycle [10]. This brief presents an HR strategy which modifies the switch control waveforms of the 25% duty-cycle mixer to suppress specific harmonics. By inserting gaps and adding pulses at particular positions within the 25% duty-cycle waveform harmonics are suppressed. At the same time the beneficial property of non-overlapping ON states of the switches is maintained which minimizes IQ cross-talk. Another advantage of this approach is that no weighting and summation of RF signals as in [6], [11] is needed to realize the rejection of harmonics. Furthermore, only one transistor is needed in each of the mixer branches I_+, I_-, Q_+ and Q_-.

II. Problem Statement

Due to the use of wide-band analog front-ends in order to cover the many different frequency bands used in LTE, blocker signals received by the antenna may reach the mixer input. These blocker signals may be down-converted to the receive (Rx) BB by the harmonic response of the mixer, thereby disturbing the reception of the wanted signal. Furthermore, Fig. 2 depicts a frequency division duplex (FDD) scenario with finite duplexer Tx-to-Rx stop-band isolation, where the Tx signal which leaks through the duplexer into the receiver is down-converted by the harmonic response of the mixer. This transmitter induced self-interference may, e.g., occur in an LTE CA mode with two receivers and one transmitter. Assuming an LTE inter-band CA mode with the uplink (UL) primary component carrier (PCC) in band 2 at $f_{Tx} = 1890$ MHz, and the downlink (DL) SCC in band 71 at $f_{LO_2} = 630$ MHz, the 3rd order harmonic response of the SCC Rx mixer occurs at 1890 MHz, which down-converts the unwanted leaked Tx signal to the Rx BB. By rejecting the 3rd order harmonic response of the mixer, the harmonic down-conversion would be suppressed.

III. System Model of the 25% Duty-Cycle Mixer

To analyze the 25% duty-cycle mixer, the complex Fourier series representation of the four mixer control signals is derived. The Fourier coefficients of the I+ branch control signal are

$$\begin{aligned} c_k &= \frac{1}{T_{LO}} \int_0^{T_{LO}} \tilde{p}_{I+}(t) e^{-jk2\pi f_{LO} t} dt \\ &= \frac{1}{2\pi} \int_0^{2\pi} p_{I+}(\varphi) e^{-jk\varphi} d\varphi \\ &= c'_k \, e^{-jk\frac{\pi}{4}}, \end{aligned} \tag{1}$$

where $c'_k = \frac{A}{4}\text{sinc}(k\frac{\pi}{4})$. The Fourier coefficients of the other phases are obtained by rotating the I+ coefficients c_k by the angle $k\frac{\pi}{2}$ where k is the harmonic index [1]. The resulting mixer phase signals can be represented by the Fourier series

$$\begin{aligned} \tilde{p}_{I+}(t) &= \sum_{k=-\infty}^{\infty} c_k e^{jk2\pi f_{LO} t} \\ \tilde{p}_{I-}(t) &= \sum_{k=-\infty}^{\infty} (-1)^k c_k e^{jk2\pi f_{LO} t} \\ \tilde{p}_{Q+}(t) &= \sum_{k=-\infty}^{\infty} e^{-jk\frac{\pi}{2}} c_k e^{jk2\pi f_{LO} t} \\ \tilde{p}_{Q-}(t) &= \sum_{k=-\infty}^{\infty} e^{jk\frac{\pi}{2}} c_k e^{jk2\pi f_{LO} t}, \end{aligned} \tag{2}$$

and the in-phase differential voltage at the output of the differential amplifier becomes

$$\begin{aligned} u_{BB,I}(t) &= i_{BB,I+}(t) * z_{BB}(t) - i_{BB,I-}(t) * z_{BB}(t) \\ &= i_{RF}(t)\tilde{p}_{I+}(t) * z_{BB}(t) - i_{RF}(t)\tilde{p}_{I-}(t) * z_{BB}(t) \\ &= \left[4\, i_{RF}(t) \sum_{\substack{k=1 \\ k\,\text{odd}}}^{\infty} c'_k \cos\!\left(k2\pi f_{LO} t - k\frac{\pi}{4}\right) \right] * z_{BB}(t), \end{aligned} \tag{3}$$

where $z_{BB}(t)$ is the low-pass BB impulse response. Similarly the quadrature-phase differential voltage becomes

$$\begin{aligned} u_{BB,Q}(t) &= i_{BB,Q+}(t) * z_{BB}(t) - i_{BB,Q-}(t) * z_{BB}(t) \\ &= i_{RF}(t)\tilde{p}_{Q-}(t) * z_{BB}(t) - i_{RF}(t)\tilde{p}_{Q+}(t) * z_{BB}(t) \\ &= \left[4\, i_{RF}(t) \sum_{\substack{k=1 \\ k\,\text{odd}}}^{\infty} \sigma_k c'_k \sin\!\left(k2\pi f_{LO} t - k\frac{\pi}{4}\right) \right] * z_{BB}(t), \end{aligned} \tag{4}$$

where $\sigma_k = -1$ for $k = 1, 5, 9, 13, \ldots$, and $\sigma_k = 1$ otherwise. Assuming a direct-conversion receiver with $f_{LO} = f_{Rx}$, and an LNA output current which contains the wanted Rx signal and a blocker (BL) signal around the 3rd order harmonic response of the mixer ($3f_{LO} = f_{BL}$)

$$i_{RF}(t) = \Re\left\{ i_{BB}^{Rx}(t) e^{j2\pi f_{Rx} t} \right\} + \Re\left\{ i_{BB}^{BL}(t) e^{j2\pi f_{BL} t} \right\}, \tag{5}$$

the received complex BB voltage becomes

$$\begin{aligned} u_{BB}(t) &= u_{BB,I}(t) + j u_{BB,Q}(t) \\ &= \left[4\, i_{RF}(t) \sum_{\substack{k=1 \\ k\,\text{odd}}}^{\infty} c'_k e^{\sigma_k j(k2\pi f_{LO} t - k\frac{\pi}{4})} \right] * z_{BB}(t) \end{aligned}$$

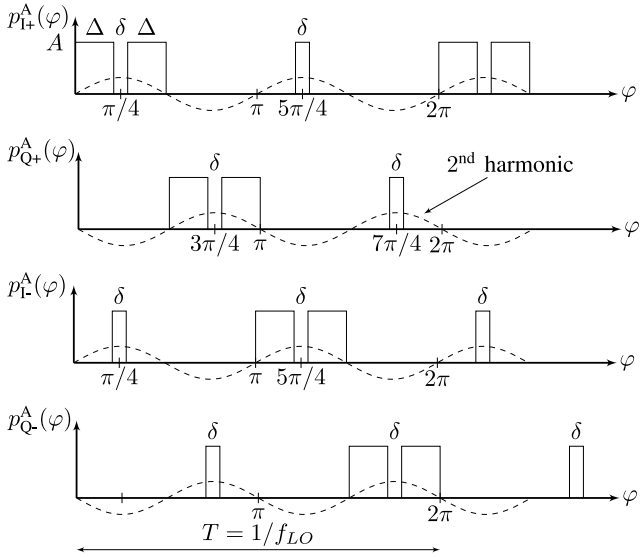

Fig. 3. Mixer control signals using the proposed waveform A in a 25% duty-cycle mixer to reject specific harmonics.

$$= \left[2c_1' i_{\mathrm{BB}}^{\mathrm{Rx}}(t) e^{+j\frac{\pi}{4}} + \underbrace{2c_3' i_{\mathrm{BB}}^{\mathrm{BL}*}(t) e^{-j\frac{3\pi}{4}}}_{\text{down-converted blocker}} \right] Z_{\mathrm{BB}}, \quad (6)$$

where Z_{BB} is the BB impedance. It can be observed, that the wanted Rx signal is down-converted by the fundamental Fourier coefficient. The unwanted blocker signal is down-converted by the 3rd order harmonic response of the mixer which disturbs the wanted signal.

IV. Proposed Harmonic Rejection Control Signals

The proposed harmonic rejection strategy suppresses specific harmonics by inserting gaps and pulses into the control signal while maintaining the constraint that at each time only one of the four branch switches is turned ON. Thereby no IQ crosstalk occurs. Due to the absence of any reverse isolation in passive mixers, the low-Q BB impedance is transformed into a high-Q RF band-pass filter seen from the RF input side [3]. Because of this reason, at any time at least one switch should be turned ON in order to provide a constant impedance seen from the RF input. By maintaining these constraints, the two proposed switch control waveforms A and B are developed.

A. Proposed Waveform A

In the proposed waveform A (Fig. 3), a gap in the middle of the 25% duty-cycle pulse, and a pulse at the offset of π with the same width δ is inserted. With careful selection of the width δ, a specific odd harmonic in the control signal may be rejected. The Fourier coefficients of waveform A can be composed by using the Fourier coefficients

$$c_k^p(\varphi_c, \Delta) = \frac{1}{2\pi} \int_{\varphi_c - \frac{\Delta}{2}}^{\varphi_c + \frac{\Delta}{2}} A e^{-jk\varphi} d\varphi = \frac{A\Delta}{2\pi} e^{-jk\varphi_c} \mathrm{sinc}\left(k\frac{\Delta}{2}\right) \quad (7)$$

of a single periodic prototype pulse centered around the angle φ_c with the width Δ and the constant amplitude A. Thereby, the Fourier coefficients of the I+ control signal $p^{\mathrm{A}}_{\mathrm{I+}}(\varphi)$ using waveform A with $\Delta = \frac{\pi}{4} - \frac{\delta}{2}$, $\varphi_{c,1} = \frac{\Delta}{2}$, $\varphi_{c,2} = \frac{\pi}{2} - \frac{\Delta}{2}$ and $\varphi_{c,3} = \frac{5\pi}{4}$ are

$$\begin{aligned} c_k^{\mathrm{A}}(\delta) &= \frac{1}{2\pi} \int_0^{2\pi} p^{\mathrm{A}}_{\mathrm{I+}}(\varphi) e^{-jk\varphi} d\varphi \\ &= c_k^p(\varphi_{c,1}, \Delta) + c_k^p(\varphi_{c,2}, \Delta) + c_k^p(\varphi_{c,3}, \delta) \\ &= \frac{A}{2\pi} e^{-jk\frac{\pi}{4}} \left[2\Delta \cos\left(k\left(\frac{\pi}{4} - \frac{\Delta}{2}\right)\right) \mathrm{sinc}\left(k\frac{\Delta}{2}\right) \right. \\ &\qquad \left. + \delta(-1)^k \mathrm{sinc}\left(k\frac{\delta}{2}\right) \right]. \end{aligned} \quad (8)$$

To reject a specific harmonic r in waveform A, the nonlinear equation $|c_r^{\mathrm{A}}(\delta)| = 0$ needs to be solved which leads to

$$2\cos\left(r\left(\frac{\pi}{4} - \frac{\Delta}{2}\right)\right) \sin\left(r\frac{\Delta}{2}\right) + (-1)^r \sin\left(r\frac{\delta}{2}\right) = 0. \quad (9)$$

The required gap/pulse width δ can be derived by using the trigonometric identity

$$\cos(\alpha)\sin(\beta) = \frac{1}{2}[\sin(\alpha + \beta) - \sin(\alpha - \beta)] \quad (10)$$

in (9) which leads to the nonlinear relationship

$$\sin\left(r\frac{\delta}{2}\right) = \frac{\sin\left(r\frac{\pi}{4}\right)}{1 - (-1)^r}. \quad (11)$$

The differential implementation of the mixers requires only the suppression of odd harmonics which leads to $1 - (-1)^r = 2$. Due to the periodicity of the sine function, solving (11) via

$$\delta_r^{\mathrm{A}} = \frac{2}{r}\left[(-1)^m \sin^{-1}\left(\frac{1}{2}\sin\left(r\frac{\pi}{4}\right)\right) + m\pi\right] \quad (12)$$

has multiple solutions for $m \in \mathbb{Z}$, where only the solutions with $0 < \delta < 25\%$ are of interest. The gap/pulse width δ_r^{A} to reject the odd harmonic r in each of the four mixer control signals is summarized in Table I. The down-conversion of the wanted Rx signal occurs with the fundamental $c_1^{\mathrm{A}}(\delta_r^{\mathrm{A}})$ which may be reduced by using the proposed HR strategy. The relative reduction of the fundamental Fourier coefficient compared to the ordinary 25% duty-cycle scheme is included in Table I. The resulting complex valued BB voltage

$$\begin{aligned} u_{\mathrm{BB}}(t) &= u_{\mathrm{BB,I}}(t) + j u_{\mathrm{BB,Q}}(t) \\ &= 4\, i_{\mathrm{RF}}(t) \sum_{\substack{k=1 \\ k\,\mathrm{odd} \\ k \neq r}}^{\infty} c_k' e^{\sigma_k j\left(k2\pi f_{\mathrm{LO}} t - k\frac{\pi}{4}\right)} * z_{\mathrm{BB}}(t) \end{aligned} \quad (13)$$

does not contain the harmonic response of the harmonic r anymore. As a consequence of that, the harmonic response at the harmonic r of the mixer is removed and the down-conversion of the unwanted spectral component within $i_{\mathrm{RF}}(t)$ at the frequency $r \cdot f_{\mathrm{LO}}$ is suppressed.

B. Proposed Waveform B

To avoid narrow pulses and gaps which may be limited by the rise- and fall-time of the control signal edges, waveform B depicted in Fig. 4 is proposed. During the period where none of the switches is turned ON, a fifth switch is needed in the

TABLE I
HARMONIC REJECTION WITH WAVEFORM A

Rejected harmonic r	δ_r^{A} in % duty-cycle	$\frac{c_1^{\mathrm{A}}(\delta_r^{\mathrm{A}})}{c_1}$ (dB)
3rd	3.8 %	−3.6 dB
5th	22.3 %	−1.7 dB
7th	15.9 %	−8.9 dB
9th	9.8 % / 23.5 %	−17.0 dB / −0.9 dB
13th	14.5 %	−12.2 dB
15th	20.8 %	−2.9 dB

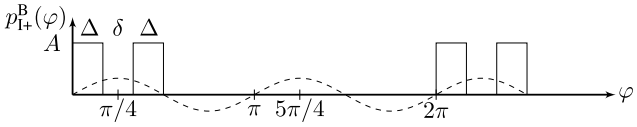

Fig. 4. Proposed waveform B without narrow pulse.

TABLE II
HARMONIC REJECTION WITH WAVEFORM B

Rejected harmonic r	δ in % duty-cycle	$\frac{c_1^{\mathrm{B}}(\delta_r^{\mathrm{B}})}{c_1}$ (dB)
3rd	8.3 %	−4.0 dB
5th	n.a.	n.a.
7th	17.9 %	−12.1 dB
9th	2.8 % / 8.3 %	−1.1 dB / −4.0 dB
13th	9.6 %	−4.7 dB
15th	8.3 %	−4.0 dB

IQ mixer which terminates the LNA output current to the BB impedance. This is needed to maintain a constant mixer input impedance seen from the LNA output current. The Fourier coefficients of waveform B are

$$\begin{aligned} c_k^{\mathrm{B}}(\delta) &= \frac{1}{2\pi}\int_0^{2\pi} p_{\mathrm{I+}}^{\mathrm{B}}(\varphi)e^{-jk\varphi}d\varphi \\ &= c_k^p(\varphi_{c,1},\Delta) + c_k^p(\varphi_{c,2},\Delta) \\ &= \frac{A\Delta}{\pi}e^{-jk\frac{\pi}{4}}\cos\left(k\left(\frac{\pi}{4}-\frac{\Delta}{2}\right)\right)\mathrm{sinc}\left(k\frac{\Delta}{2}\right), \end{aligned} \quad (14)$$

and by setting $|c_r^{\mathrm{B}}(\delta)| = 0$ for a specific harmonic r the nonlinear equation $\sin(r\frac{\delta}{2}) = \sin(r\frac{\pi}{4})$ is obtained. Solving this nonlinear relationship leads to the multiple solutions

$$\delta_r^{\mathrm{B}} = \frac{2}{r}\left[(-1)^m r\frac{\pi}{4} + m\pi\right] \quad (15)$$

for $m \in \mathbb{Z}$, where only the solutions with $0 < \delta < 25\%$ are of interest. The required gap widths to reject specific harmonics using waveform B are summarized in Table II. With the duty-cycle of 8.33%, the 3rd, 9th and 15th harmonic are suppressed simultaneously. Table I and Table II show that for both waveforms the fundamental amplitude is reduced by about 4 dB when the 3rd order harmonic response is rejected.

C. Waveform Generation Using Digital-to-Time Converters

The proposed pulse generator (PGEN) circuit to generate the pulse with the width δ at the desired phase location is shown in Fig. 5. The circuit consists of two digitally-controlled edge interpolator (DCEI) based digital-to-time converters (DTCs) [12] with the three input phases φ_{1-3} which

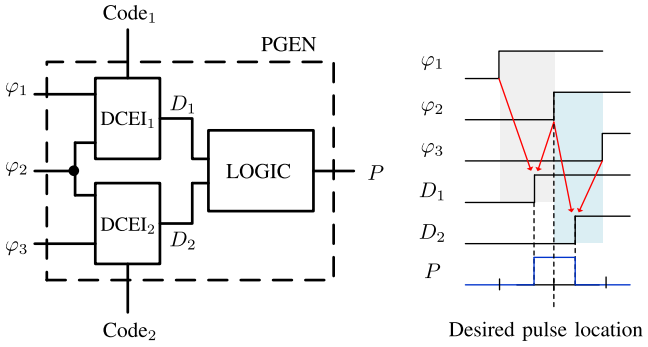

Fig. 5. Pulse generation circuit.

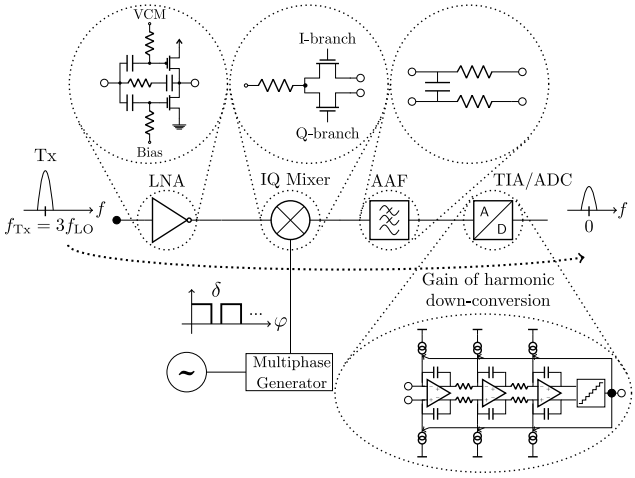

Fig. 6. Circuit simulation setup using waveform A and B.

have a fixed phase relationship of $\pi/4$ between each other. The two DCEIs interpolate the edges between the input phases corresponding to the code words to generate the rising edges D_1 and D_2. For identical code words, the pulse P which is generated by combining the signals D_1 and D_2 using an XOR logic is centered around the rising edge of φ_2. By combining the signal P and $p_{\mathrm{I+}}(\varphi)$ by an XOR logic, the desired signal $p_{\mathrm{I+}}^{\mathrm{A}}(\varphi)$ with the gap width δ within the 25% duty-cycle pulse is generated. Using the DTC based approach described in [12], the edges D_1 and D_2 may be generated with a time resolution of 244 fs at an LO frequency of 2 GHz. The final pulse-width resolution becomes 488 fs corresponding to less than 0.1% of a duty-cycle.

V. CIRCUIT SIMULATION RESULTS

The proposed HR strategy is evaluated by circuit simulations using a 28 nm CMOS technology. The simulation architecture is depicted in Fig. 6, where the LNA is implemented as a transconductance amplifier (voltage-to-current converter) which drives the succeeding 25% duty-cycle current driven passive mixer. The RF signal is down-converted by the mixer, filtered by the anti-aliasing-filter (AAF) and digitized by the TIA/analog-to-digital converter (ADC). For the simulation, the receiver was configured for the reception of 10 MHz LTE signals in band 71 at $f_{\mathrm{LO}} = 630$ MHz. At the input of the LNA a Tx related blocker signal at 1890 MHz

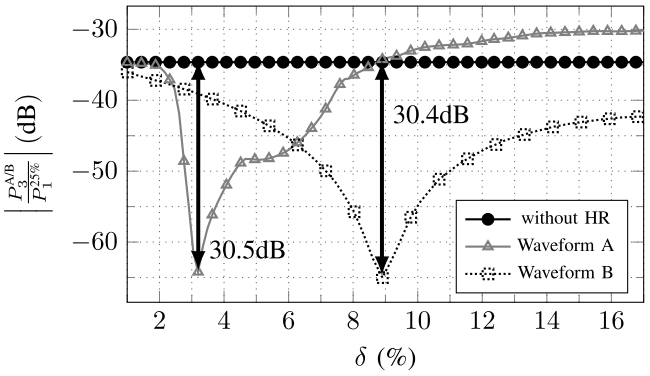

Fig. 7. Pulse/gap-width dependent BB power of the blocker signal which is down-converted by the 3rd order harmonic response of the mixer.

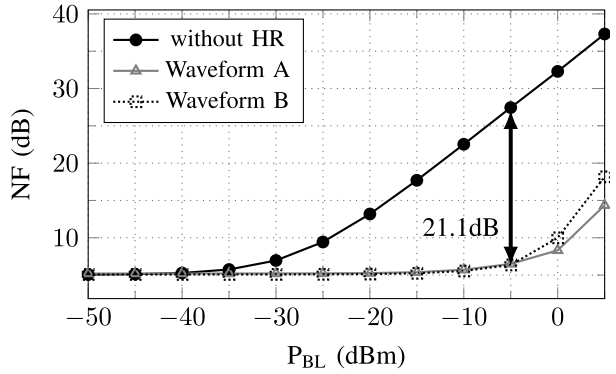

Fig. 8. Blocker power dependent NF of the receiver at the ADC output.

was inserted which is down-converted by the 3rd order harmonic response of the mixer (see Fig. 2). Waveforms A and B are configured to reject the 3rd order harmonic response of the mixer, and the resulting BB signal power with and without the proposed HR strategy is plotted in Fig. 7. The BB power $P_3^{A/B}$ using waveform A/B is a result of the harmonic down-conversion of the blocker signal with the 3rd order harmonic response of the mixer. This power is normalized to the BB power $P_1^{25\%}$ which corresponds to the down-conversion of the blocker signal at f_{LO} with the fundamental of the ordinary 25% duty-cycle mixer. Without HR, the Tx related blocker signal is suppressed by about 35 dB which includes also the front-end filtering. Using waveform A/B, an improvement of the blocker suppression by 30.5 dB and 30.4 dB is achieved. Deviations of the nominal value of δ leads to a reduced suppression of the blocker. It can be observed that waveform B is less sensitive to variations in δ. Due to the finite rise and fall times of the control signals in the circuit simulation, the optimal values of δ differ slightly from the values presented in Tables I and II. Fig. 8 shows that the NF of the receiver is significantly improved when the 3rd order harmonic response of the mixer is suppressed. Modern front-ends may include an external LNA (eLNA) with improved linearity. Assuming a Tx power of 27 dBm at the power amplifier (PA) output, 50 dB duplexer suppression and a gain of 18 dB of the eLNA, the resulting blocker power at the receiver chip input may reach −5 dBm. At the blocker power of −5 dBm, the NF can be improved by about 21 dB with both proposed waveforms.

VI. Conclusion

This brief proposed an HR strategy for 25% duty-cycle current-driven passive mixers. By inserting gaps and pulses of variable width into the mixer control waveforms, specific harmonics are suppressed. The generation of the proposed mixer control waveforms involves two DTCs and an XOR. The pulse/gap width is adjusted by the digital input code words of the DTCs which enable a dynamic selection of the suppressed harmonic. By circuit simulations using a 28 nm CMOS technology we demonstrated a suppression of the 3rd order harmonic response of more than 30 dB, and an improvement of the NF of the receiver by 21.1 dB for a blocker power level of −5 dBm at the LNA input.

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