Zhang, C., Laughlin, L., Beach, M. A., Morris, K. A., & Haine, J. L. (2018). Phase Noise in Full-Duplex Radios Using Off-the-Shelf Oscillators. *IEEE Wireless Communications Letters*, *7*(2), 206-209. https://doi.org/10.1109/LWC.2017.2764905

Peer reviewed version

Phase Noise in Full-Duplex Radios using off-the-shelf Oscillators

Chunqing Zhang, Leo Laughlin, Mark A. Beach, Kevin A. Morris, and John L. Haine.

***Abstract*—This paper analyses the effect of phase noise (PN) on digital non-linear self-interference (SI) cancellation for low delay spread non-linear SI channels typical of small form factor in-band full-duplex (IBFD) radios in indoor environments. Use of a shared local oscillator (LO) between the transmit and receive radios is assumed, and it is shown that, in theory, un-cancelled SI due to PN can be reduced by optimising the relative delay between Rx LO path and SI channel. Simulations and measurements from hardware IBFD transceivers using realistic LO phase noise characteristics demonstrate that, for these devices/environments, in practice LO phase noise does not limit digital cancellation when using shared local oscillators.**

***Index Terms*—In-band full-duplex, shared local oscillator, phase noise, non-linearity.**

I. Introduction

In band full-duplex (IBFD) which allows simultaneous transmit (Tx) and receive (Rx) on the same frequency, has the potential to increase spectral efficiency, reduce latency, and solve the hidden node problem [1]. However, transmitting and receiving simultaneously gives rise to the problem of self-interference (SI), whereby the Tx signal causes catastrophic interference at the receiver, preventing the Rx signal from being decoded. IBFD operation therefore requires high transmit-to-receive (Tx-Rx) isolation in the full-duplex transceiver to mitigate the SI, allowing the receiver to achieve an acceptable signal-to-interference-and-noise-ratio (SINR).

Various SI cancellation techniques have been presented in the literature, for example antenna separation and/or isolation, radio frequency (RF) cancellation, and digital baseband cancellation [1]. RF imperfections have been identified as the primary factor which limits the performance these systems [2]. This paper investigates the effect of local oscillator (LO) phase noise (PN) on digital baseband cancellation, providing experimental results for IBFD transceivers with different hardware characteristics and multipath SI channels. In particular this paper addresses transceivers in which the same LO signal is used for up and down conversion, as may be possible in small devices, and considers the effect of delay mismatch between the SI and LO signal paths. Some previous research [2]–[6] has already addressed the effect of phase noise on self-interference cancellation. In [3], [4] the authors report theoretical and simulated results for the level of digital cancellation (DC) acheived in the presence of phase noise and multipath. However, [3] and [4] model only linear SI, and use a simulated multipath channel which models outdoor relay devices, therefore these results may not applicable to smaller devices and indoor scenarios. In [5] and [6] SI channel delay was considered, but these works did not combine this analysis with multipath or non-linear SI channels. In [2] it is shown that PN can be a limiting factor in IBFD transceivers employing linear active RF cancellation, however this work did not analyse the dependence on the LO characteristics or determine the achievable performance for systems with typical phase noise and non-linearity.

Manuscript received July 9, 2017; revised September 14, 2017 and October 16, 2017; accepted October 16, 2017. Date of publication xxxx xx, 2017; date of current version xxxx xx, 2017. The associate editor coordinating the review of this paper and approving it for publication was xxx.

This work is supported by the UK Engineering and Physical Sciences Research Council through the Full-duplex for scalable dense networks (SENSE) project - EP/P002978/1), and the CDT Equipment Award (EP/K035746/1).

The authors are with the Communication Systems & Networks Laboratory, University of Bristol, UK, BS8 1UB. e-mail: jack.zhang@bristol.ac.uk.

Digital Object Identifier xxxx.

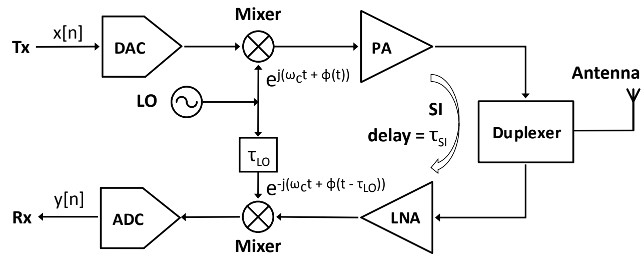

Fig. 1. Block diagram of a one antenna Homodyne IBFD transceiver sharing the same LO for Tx and Rx.

The novel contribution of this paper is twofold: Firstly, section II and III respectively present theoretical modelling and simulations of non-linear SI in the presence of phase noise and delay mismatch, showing that the combination of PN and power amplifier (PA) non-linearity results in SI components which cannot be perfectly cancelled, and quantifying the effect of delay mismatch between the LO and SI paths on non-linear digital SI cancellation. Secondly, section IV presents measured digital cancellation results from hardware IBFD transceivers in indoor environments, with and without delay matching of the LO and SI delay. The simulations and measurements presented in this paper use the phase noise characteristics of real frequency synthesizer devices typical of low cost radio transceivers, and also compare this with the free-running oscillator model used in previous analysis [3]. Results show that for the SI channel delay spreads observed in these devices/environments are lower than for relay applications, reducing the impact of PN, and that the phase noise from practical LO devices is not a limiting factor in this application.

II. System Model and Mathematical Analysis

This section applies a well known memory polynomial model [7] to model SI in the presence of phase noise and non-linearity in the SI channel. Fig. 1 depicts an IBFD transceiver which is sharing the LO between the transmitter and receiver chain. The baseband complex waveform $x[n]$ is up converted by mixer (with LO at frequency ω_c and phase noise $\phi(t)$). The Tx signal then passes through the SI channel, which is non-linear due primarily to the power amplifier (PA), and which may include some form of cancellation in the duplexing circuitry (not shown), to arrive as SI at the receiver input. The SI (along with the desired Rx signal) is mixed down using the same LO.

For brevity this analysis is restricted to analysing the linear component and the third order intermodulation components (IM3) of the SI (however the insights gained from this analysis apply to all intermodulation components and also to the multipath SI channel). As shown in [7], the IM3 components generated by the PA are given in baseband as

$$y_3[n] = \gamma_3 \sum_{m_1=0}^{M-1} \sum_{m_2=m_1}^{M-1} \Bigg[g_{m_1,m_2}(n) * \Big(x[n]x[n-m_1]x^*[n-m_2] + x[n]x^*[n-m_1]x[n-m_2] + x^*[n]x[n-m_1]x[n-m_2] \Big) \Bigg]. \tag{1}$$

where γ_3 is a common scaling factor, M is the maximum memory length, and g_{m_1,m_2} is a filter function as determined by the Volterra kernel [7]. To investigate the theoretical interaction between this non-linearity in the SI channel and the LO phase noise, this model can be extended by considering the relative phase shifts between $x[n]$, $x[n-m_1]$, $x[n-m_2]$, and their conjugates, which result from the LO phase noise in both the up and down conversion. For example, if we represent the continuous phase noise, $\phi(t)$, in discrete baseband as $\phi[nT]$ or $\phi[n]$ in brief, where T is sampling interval and n is the n^{th} sampling point, then at Rx time n the term $x[n-m_1]$ will have been upconverted with a phase offset of $\phi[n-m_1]$, and is down converted with a phase offset of $-\phi[n-n_\tau]$, where the effective LO delay for down conversion, τ, is the difference between the mean SI channel delay and the Tx-Rx LO delay, such that $\tau = \tau_{LO} - \tau_{SI}$, and it is assumed that this can be adequately represented in the discrete as n_τ, i.e., the sampling rate is high enough (although in practice τ is not multiple of sampling interval). Therefore when considering the phase noise, the $x[n-m_1]$ component becomes $x[n-m_1]e^{j\phi[n-m_1]}e^{-j\phi[n-n_\tau]}$. Thus the memory polynomial can be extended to include phase noise by making the following substitutions

$$x[n] = x[n]e^{j\phi[n]}e^{-j\phi[n-n_\tau]} \tag{2}$$
$$x[n-m_1] = x[n-m_1]e^{j\phi[n-m_1]}e^{-j\phi[n-n_\tau]} \tag{3}$$
$$x[n-m_2] = x[n-m_2]e^{j\phi[n-m_2]}e^{-j\phi[n-n_\tau]} \tag{4}$$
$$x^*[n] = x^*[n]e^{-j\phi[n]}e^{-j\phi[n-n_\tau]} \tag{5}$$
$$x^*[n-m_1] = x^*[n-m_1]e^{-j\phi[n-m_1]}e^{-j\phi[n-n_\tau]} \tag{6}$$
$$x^*[n-m_2] = x^*[n-m_2]e^{-j\phi[n-m_2]}e^{-j\phi[n-n_\tau]}. \tag{7}$$

Substituting (2)-(7) into (1) and rearranging gives the IM3 component with phase noise as

$$y_3^{'}[n] = \gamma_3 \sum_{m_1=0}^{M-1} \sum_{m_2=m_1}^{M-1} \Bigg[g_{m_1,m_2}(n) * \Big(x[n]x[n-m_1]x^*[n-m_2]e^{j(\phi[n]+\phi[n-m_1]-\phi[n-m_2]-\phi[n-n_\tau])} + x[n]x^*[n-m_1]x[n-m_2]e^{j(\phi[n]-\phi[n-m_1]+\phi[n-m_2]-\phi[n-n_\tau])} + x^*[n]x[n-m_1]x[n-m_2]e^{j(-\phi[n]+\phi[n-m_1]+\phi[n-m_2]-\phi[n-n_\tau])} \Big) \Bigg]. \tag{8}$$

Similarly, the linear SI component with PN is given by

$$y_1^{'}[n] = \gamma_1 g_0 * x[n]e^{j(\phi[n]-\phi[n-n_\tau])}. \tag{9}$$

We observe from (9) that when the delay in the LO and SI channel is matched (i.e., $n_\tau = 0$), that for the linear component, the PN is the same in the upconversion and downconversion steps, and therefore cancels out. However, from (8), even with delay matching the non-linear SI components still depend on the PN, and therefore in theory these cannot be completely cancelled. Observing the effect of delay mismatch in (9), due to the small delay in in-door scenarios, i.e., n_τ is small, it may be noted that the PN change between adjacent discrete LO signal is small and will be largely influenced by the higher frequency offset PN components, i.e., PN floor. This will be proved in section III.

The combination of multipath propagation and PN in the SI channel has been shown to degrade the DC performance [3]. Similarly, if we include multipath propagation in the above model, then both (8) and (9) would result in more SI components which depend on the phase noise and therefore cannot be cancelled, further degrading performance. The digital cancellation which can be achieved in practice depends on hardware imperfections and multipath coupling in real systems (i.e., PN, PA non-linearity, and delay spread) and is evaluated in the following sections.

III. Simulated Performance

A. Simulation Setup

To allow the interaction between PN and non-linearity to be observed, this simulation assumes PA non-linearity and PN are the only transceiver impairments (no receiver thermal noise is included). The simulation uses a simplified Volterra series memory polynomial PA model [8], and uses the same PA parameters as given in [8, example 2].

Normally oscillator PN is specified as a piecewise or discrete spectrum in dBc/Hz at certain frequency offset and the PN is highly related to the operating frequency [9], therefore, this paper focus the discussion on sub-6 GHz bands where full-duplex is likely to be used. In these simulations, the PN waveforms are generated according to the PSD specification of off-the-shelf Qorvo RF2051 wideband RF phase-locked loop (PLL) frequency synthesizer [10], or by using the free-running oscillator mode as used in [3]. Two LOs are simulated: a closed-loop Qorvo LO, with a 3-dB BW of 30 Hz, and a free-running LO, also with a 30 Hz 3-dB BW (see Table I for PSD data). Throughout this paper the DC algorithm proposed in [11] is used; this algorithm cancels SI due to linear and non-linear SI, including the effects of IQ imbalance.

The simulation generates Tx waveforms, adds the upconversion phase noise, applies the SI channel model (including the PA non-linearity and assumed analog domain isolation), adds the down conversion phase noise (subject to delay matching parameter n_τ), and then performs digital cancellation. The simulation has been run using a two tone waveform with tone spacing of 4 MHz centred about the carrier frequency, and also using a 20-MHz bandwidth LTE uplink waveform, with these waveforms giving essentially the same result. The simulation

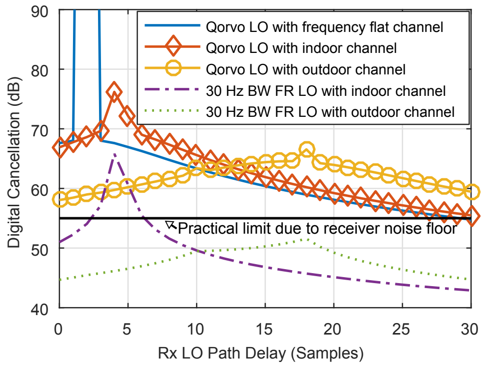

Fig. 2. Simulated linear SI Channel DC performance as a function of Rx LO path delay. (The linear PA has a delay of 2 samples. FR=free-running.)

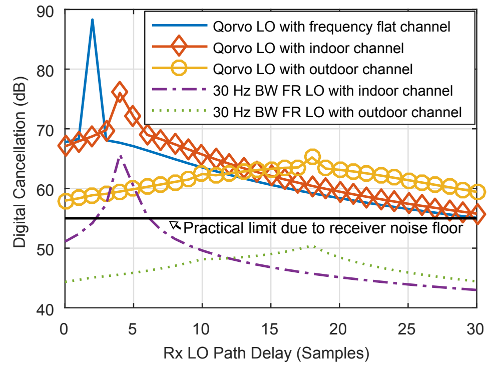

Fig. 3. Simulated non-linear SI Channel DC performance as a function of Rx LO path delay. (FR=free-running.)

sampling rate is 120 MS/s, and thus each sample delay in the LO path corresponds to a delay of 8.3 ns. The Tx power is 20 dBm, and 60 dB of analog isolation is assumed, i.e., -40 dBm Rx SI power at the receiver input, which is realistic according to the literature [12]. Two multipath channels are used: the same outdoor relay multipath SI channel, as used in [3] and [4], and an indoor multipath channel based on measurements from the dual antenna hardware IBFD transceiver described below in Section IV. For comparison, simulations are run with and without the non-linear PA model. Also for comparison, a frequency flat SI channel is also used in both the linear and non-linear simulations. The simulation reports the amount of digital cancellation achieved for different values of Rx LO path delay, $n_{\tau_{LO}}$.

B. Results

Fig. 2 shows the amount of digital cancellation achieved as a function of LO path delay for the different multipath channel and LO PSDs when ideal linear PA is assumed. In the case of the frequency flat channel, matching the LO delay and channel delay (i.e. $n_{\tau_{LO}} = 2$) a theoretically infinite amount of digital cancellation is achieved, as predicted by (9). However, without a matched delay the PN results in imperfect cancellation as expected. It is also clear from Fig. 2 and Fig. 3 that the shorter delay spread in the indoor channel reduces the impact of the phase noise, however, more significant fact is that for the realistic Qorvo oscillator PSD, the phase noise is not a limiting factor as the residual SI due to phase noise is below the receiver noise floor. As shown in [4], the flicker noise and thermal noise components can have a substantial impact on the level of cancellation, and as such the free-running model may under-predict the level of digital cancellation which can be achieved [4]. Since the two LO signals used in this simulation have the same 3-dB bandwidth the differences shown above can be attributed to the different spectral shape at higher frequency offsets. Whereas the free-running model predicts a reduction in digital cancellation due to phase noise, the results using practical LO spectral shape quantify performance when using a typical handset frequency synthesizer, and show that in practice the phase noise will not be a limiting factor.

Comparing Fig. 2 and 3, it can be seen that the interaction between phase noise and PA non-linearity has no substantial impact on the digital cancellation, as the only notable difference is that, for the frequency flat channel with perfect delay matching, the non-linearity prevents theoretically infinite cancellation; the cancellation is limited by residual non-linear SI, however this remains well below the noise floor. Finally, it may be noted that for the Qorvo LO, whilst delay matching does increase digital cancellation, the residual SI due to phase noise is significantly below the thermal noise floor for all simulated LO delays, and therefore in practice delay matching is not critical for digital cancellation.

IV. Experiment Results

A. Hardware Setup

To verify the mathematical analysis and simulation results, hardware measurements have been made using a National Instruments (NI) PXIe platform, with NI PXIe-5644R vector signal transceiver (VST) hardware to implement the transmitter and receiver subsystems. A Mini-Circuits ZX95-2500+ voltage controlled oscillator (VCO) is used as LO source. In open loop this device has a 3-dB bandwidth of <10 Hz but relatively high flicker noise compared to that of a closed loop synthesizer - it would therefore be expected that this device would have a larger impact on cancellation compared to the Qorvo device [4]. To emulate the Qorvo LO and free-running LO used in the simulations, LO output signals with the same characteristics are generated using another VST transmitter. This emulation is possible since the VST phase noise and thermal noise is below that of the emulated devices, thus allowing these LO signals to be accurately generated. Phase noise PSDs for all LO signals are given in Table I. The LO signal is shared between the Tx and Rx mixers using an RF splitter. To observe the effect of delay in the receiver LO path, measurements are taken with no relative delay between the Tx and Rx LO input, and with added delay in the Rx LO path using a length of cable. A Mini-Circuits ZX60P162 amplifier, operated close to its 1-dB compression point, is used in the PA, resulting in a non-linear SI channel. The PA's output power is 16 dBm. A 20 MHz LTE signal is used as the Tx signal and the VST receiver's thermal noise floor is about -165 dBm/Hz. Three SI channel configurations are used: a cabled connection

TABLE I
PHASE NOISE POWER SPECTRAL DENSITIES OF THE LO SIGNALS USED IN THE MEASUREMENTS AND SIMULATIONS. NF=NOISE FLOOR.

LO Source (3-dB BW (Hz))	PN (dBc/Hz) at offsets				
	1 kHz	10 kHz	100 kHz	1 MHz	NF
Qorvo	-85	-90	-93	-130	-150
Mini-Circuits	-62	-89	-111	-131	-161
Free-running (30)	-50	-70	-90	-110	-142
Free-running (50)	-48	-68	-88	-108	-140

TABLE II
MEASURED LEVEL OF DIGITAL CANCELLATION FOR DIFFERENT IBFD TRANSCEIVER ARCHITECTURES AND LO PSDS.

LO Source (3-dB BW (Hz))	SI Channel Configuration	RMS Delay Spread (ns)	Rx LO Delay	DC
Qorvo (30)	Dual antenna	12.4	0	55 dB
Qorvo (30)	Dual antenna	12.4	~6 ns	55 dB
Qorvo (30)	EBD	27.5	0	55 dB
Qorvo (30)	EBD	27.5	~6 ns	55 dB
Mini-Circuits (<10)	Dual antenna	12.4	0	55 dB
Mini-Circuits (<10)	Dual antenna	12.4	~6 ns	55 dB
Mini-Circuits (<10)	EBD	27.5	0	55 dB
Mini-Circuits (<10)	EBD	27.5	~6 ns	55 dB
Free-running (30)	Dual antenna	12.4	0	53 dB
Free-running (30)	Dual antenna	12.4	~6 ns	55 dB
Free-running (30)	EBD	27.5	0	50 dB
Free-running (30)	EBD	27.5	~6 ns	53 dB
Free-running (50)	EBD	27.5	0	48 dB
Free-running (50)	EBD	27.5	~8 ns	53 dB

with RF attenuators which provides a frequency flat channel with 53 dB loss, an electrical balance duplexer (EBD) of the configuration described in [13] (also with 53 dB attenuation), and another configuration with separate Tx and Rx antennas. In the dual antenna configuration the antennas are 20 cm apart and a metal sheet and absorbing materials are placed between the antennas to block the line of sight SI coupling, giving antenna isolation of 43 dB. To ensure all SI channels resulted in the same receive power (to ensure differences are due only to the phase noise, non-linearity, and delay spread and not the receiver signal-to-noise ratio) in the dual antenna setup a 10 dB attenuator is inserted at the receiver input.

B. Results

Table II gives the level of digital cancellation for the different hardware configurations. For all LO and delay matching configurations, the frequency flat SI channel resulted in no reduction in DC due to phase noise (not shown in Table II). When using the Qorvo and Mini-Circuits LOs, no reduction in DC was observed due to phase noise, with or without the LO path delay. Reductions in digital cancellation were only observed when using the LO signal generated according to the free-running model. In this case, as predicted by previous works [3], [4] a higher delay spread in the SI channel causes a larger reduction in cancellation. The effect of delay matching can also be observed; this provided several dB of improvement in cancellation for the free-running case. However, from Table I, it can be seen that the free-running model produces significantly higher phase noise than is observed in practice, and for the more realistic LO PSDs no impact on digital cancellation was observed, and there is no benefit in using delay matching. Thus in practice phase noise is not a limiting factor for digital cancellation for small form factor devices in indoor environments (provided sufficient RF cancellation has been achieved).

V. CONCLUSION

This paper has analysed the impact of phase noise in IBFD transceivers with linear and non-linear SI channels and shared Tx and Rx LOs for different modelled and practical LO PSD characteristics. In theory, even with shared LOs and perfect delay matching between the SI channel and Rx LO signal path, the interaction of phase noise and non-linearity results in phase noise dependent SI which cannot be completely cancelled. However, simulations and measurements have shown that for practical device parameters, the combination of phase noise and non-linearity does not limit digital cancellation. Simulations and measurements of hardware IBFD transceivers using practical LO signals in indoor environments have achieved digital cancellation of SI down to the receiver noise floor, demonstrating that in practice phase noise does not limit cancellation for this device type and environment.

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