An FD/FDD Transceiver with RX Band Thermal, Quantization, and Phase Noise Rejection and >64dB TX Signal Cancellation

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***Abstract*—A transceiver system with active cancellation of the TX signal for full duplex (FD) or frequency division duplex systems (FDD) is presented. A replica cancellation digital-to-analog converter and highly linear receiver with +25dBm OOB IIP3 enable FDD operation without a duplexer at TX power up to +17dBm, FD operation without a circulator up to +5dBm, and FD operation with a circulator up to +13dBm. In addition to providing over 64dB of RF cancellation for a 20MHz modulated TX signal, the front-end demonstrates techniques to cancel noise sources in the RX band, including 3dB reduction of the thermal noise from the self-interference cancellation circuits, >25dB cancellation of quantization noise from the digital TX, and >20dB cancellation of TX LO phase noise in the RX band.**

***Index Terms*—frequency division duplex, full duplex, self interference, active cancellation, thermal noise cancellation, phase noise cancellation.**

I. Introduction

Agile hardware which can transmit and receive at overlapped frequency in full duplex (FD) operation, or tune to arbitrarily spaced transmit/receive (TX/RX) channels for frequency division duplex (FDD) operation is a key enabler for efficient spectrum utilization in next generation mobile systems. The challenge for such systems is providing frequency tunable isolation of the high power TX signal at the RX input during simultaneous operation. Prior work [1][2] has demonstrated active cancellation techniques enabling fully integrated transceivers with simultaneous TX/RX operation without a duplexer. However, these systems are 1) limited in their ability to handle the noise of the cancellation circuits in the RX band, and 2) limited to operation at <+15dBm TX output power due to RX compression [1][2][6].

This work presents a transceiver system addressing these two limitations through 1) a feedback cancellation loop for thermal noise created in the RX band by the cancellation circuits, 2) digital cancellation of the TX quantization noise in the RX band, 3) feed-forward cancellation of the TX LO phase noise in the RX band, and 4) a highly linear RX design enabling operation in the presence of residual TX interference. Together, these techniques enable simultaneous TX/RX operation at a record +17dBm TX output power at the antenna without an off-chip duplexer, with >64dB analog isolation and >25dB digital cancellation for 20MHz modulated signals.

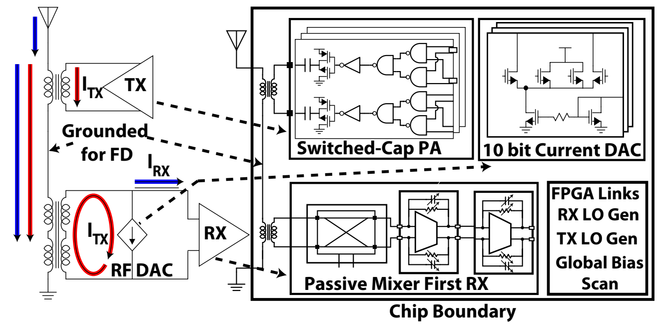

Fig. 1. Top level transceiver system diagram.

II. Transceiver Architecture

Fig. 1 shows the system architecture of the transceiver. As shown in [1], a replica cancellation digital-to-analog converter (DAC) is placed in shunt with the RX port, providing a TX signal dependent virtual ground across the RX. When placed in the series combination, the virtual ground shields the TX from RX loading, and the RX from the TX signal. To test FD operation, the node in between TX and RX in Fig. 1 can be shorted to ground, and the TX and RX connected through an external circulator.

The main RX band noise sources in this approach are the thermal noise of the replica current DAC, the TX residual after analog cancellation set by the DAC least significant bit (LSB) value, and TX phase noise. Further, the digital switching transmitter produces strong harmonic content that is not cancelled by the replica DAC, which can compress the RX at high TX output powers.

III. Replica DAC Thermal Noise Cancellation

A key desensitization mechanism in the system is the current DAC's thermal noise that falls in the RX band. A schematic of the cancellation DAC is shown in Fig. 2. The DAC is class-A, with a DC current drawn from the center-tap of the RX transformer. A digital input code-word connects a fraction of unit cells to the RX input, while other cells shunt their current to the center-tap through Mb.

The DAC mixer switches are hard driven and only one switch (Ma-c) is enabled in a unit cell at a given time. If the tail transistor (M1) drain node bandwidth is higher than the RF center frequency, the high drain impedance will circulate the switch noise. The dominant source of thermal noise in the RX band is tail noise at

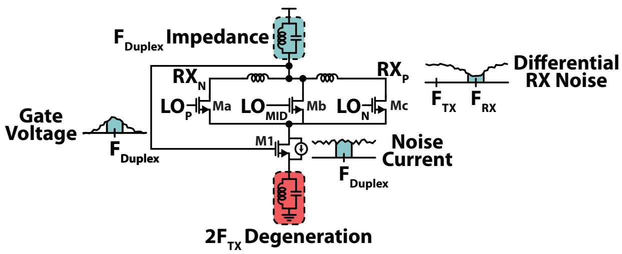

Fig. 2. High (red) and low (blue) frequency DAC noise reduction.

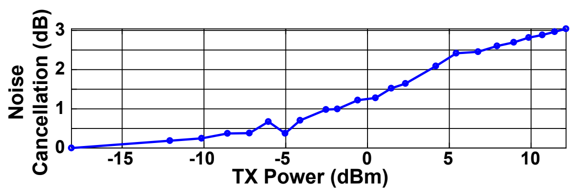

Fig. 3. RX band noise cancellation during FD operation.

F_{Duplex} and $2F_{TX} + F_{Duplex}$ up- and down-converted respectively by the fundamental of the TX LO. The DAC's thermal noise as a function of TX output power (P_{TX}), antenna resistance (R_{ANT}), RX transformer turns ratio (N_{Turns}), and tail overdrive voltage (V_{od}) is expressed as:

$$\frac{i^2_{\mathrm{n}}}{\Delta f} = 8kT\gamma\sqrt{\frac{2P_{\mathrm{TX}}}{R_{\mathrm{ANT}}}}\frac{1}{\pi N_{\mathrm{Turns}}}\frac{1}{\mathrm{V_{od}}}. \qquad (1)$$

Without thermal noise cancellation, at P_{TX}=+15dBm, the NF from the DAC alone is 8.2dB (N_{Turns}=2, R_{Ant}=50, V_{od}=400mV).

Two resonant feedback networks are used to reduce M1's thermal noise at F_{Duplex} and $2F_{TX} + F_{Duplex}$, shown in Fig. 2. Resonant networks are used to provide a high impedance for frequencies of interest, while not injecting significant noise from the passive components. As the resonance around $2F_{TX}$ is more sensitive to parasitics, the $2F_{TX}$ impedance was placed on the tail source, and the baseband resonance fed-back to the tail gate. The $2F_{TX}$ resonance is synthesized with a PCB via network, while the baseband resonance uses discrete passives.

At low TX powers, the majority of the noise which flows through the center tap impedance and is fed back is due to disabled cells, offsetting the feedback cancellation. However at low TX powers, the NF is dominated by the RX chain, not DAC thermal noise. The DAC thermal noise is most dominant at >+10dBm FD operation with external isolation. In this regime, peak thermal noise cancellation of 3dB is measured at +12dBm, shown in Fig. 3. The 1dB BW is 2.25MHz, corresponding to 15MHz BW at F_{Duplex} = 40MHz with the same resonant Q.

IV. Digital Cancellation of TX Quantization Noise

Active cancellation front-ends [1][2][5] provide limited front-end isolation, just sufficient to linearly operate the RX. The remaining TX residual must be subtracted in the digital domain to restore full RX sensitivity. In this architecture, the DAC is oversampled relative to the TX and RX bandwidth. Accordingly, the residual TX power in the RX band after analog cancellation, due to DAC quantization error, is reduced by 3dB per octave of oversampling. This has been verified in measurement; increasing the oversampling ratio from 3.125x to 6.25x to 12.5x results in a 3dB per step decrease in the RX band leakage. Utilizing 9 bits of the DAC dynamic range for a 20MHz modulated signal at 12.5x oversampling results in 64dB of measured front-end cancellation, as predicted by

$$6\frac{dB}{bit}\times 9bits + 10\log_{10}(12.5) = 65dB. \qquad (2)$$

This residual TX signal is further subtracted in the digital domain. Note that the quantization error is a data correlated sequence - for a fixed TX/DAC code pairing, the static nonlinearity is a slowly varying but fixed error.

Digital cancellation is performed via an off-chip digital lookup table (LUT) implemented in software, which for each TX/DAC code-word pairing stores a sampled version of the resulting $TX - DAC$ difference symbol. This LUT is then used to predict the quantization error for subsequent random data sequences, composed of the same TX/DAC symbol pairings, as shown in Fig. 4.

The residual signal at the output of the RX is predicted and cancelled for random TX data inputs by 25-30dB on top of the analog cancellation. Accordingly, this digital cancellation technique is capable of cancelling the TX and DAC quantization noise, extending the total system cancellation to 4-5 bits below the LSB level of the replica DAC. When paired with the 64dB analog cancellation, the system provides over 90dB of total (analog plus digital) cancellation of the TX signal.

V. Phase Noise Cancellation

TX source phase noise is a dominant source of RX desensitization – note that a -150dBc/Hz TX LO phase noise scaled by a +10dBm TX signal can result in -140dBm/Hz RX band noise. As the TX and cancellation DAC shown in Fig. 1 can share the same LOs, close-in source phase noise correlated between the TX and DAC can be feed-forward cancelled in the same manner as the TX signal. While this approach has been demonstrated in [1][3], providing cancellation up to 20dB at 40MHz offset from the TX carrier, phase noise cancellation falls to

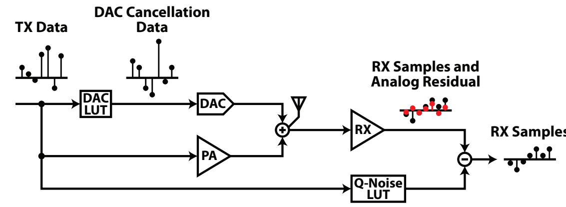

Fig. 4. Back-end digital cancellation of DAC quantization noise.

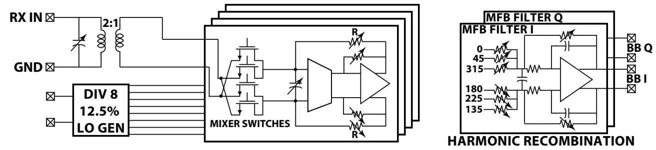

Fig. 5. RX top level schematic.

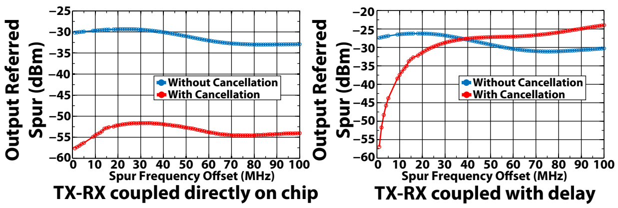

Fig. 6. Phase noise cancellation with wideband (left) and transmission line (right) TX/RX coupling.

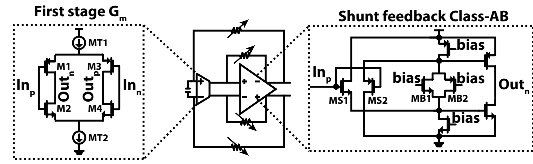

Fig. 7. RX first stage amplifier design.

<10dB at a far-out offset of 100MHz.

In this work, it is shown that the bandwidth of the leakage path from TX to RX sets the bandwidth of the phase noise cancellation. As the DAC provides a phase shift to match the TX signal, the DAC's phase noise close-in also matches the TX phase noise, and is feed-forward cancelled. However, at far-out offsets where the leakage network has a different phase shift, the DAC and the TX phase noise sum out-of-phase, and cannot be cancelled. This result is verified in Fig. 6. When the TX and RX are isolated, and coupled off-chip via a long transmission line, the TX signal at the RX input experiences a significant frequency dependent phase shift. Accordingly, the phase noise cancellation becomes narrowband. In fact, when the TX-RX network phase shift reaches 180°, the phase noise between TX and DAC add constructively, and the output noise floor is increased by 6dB. When the DAC and TX are coupled on-chip with a wideband passive network, the phase noise cancellation is shown to be wideband, and >20dB of source phase noise cancellation is provided past 100MHz offset from the TX carrier.

VI. High Linearity FD/FDD Receiver

A switched-capacitor digital power amplifier (DPA) is used to provide a low series impedance for the RX's antenna interface. Due to the switching nature of the DPA, strong emissions at TX harmonics, not cancelled by the DAC, can compress the RX. A receiver with high linearity and harmonic rejection (HR) is needed to operate in the presence of strong out-of-band (OOB) TX interference. The passive-mixer-first (PMF) receiver, shown in Fig. 5, is a good option for such FD/FDD systems, due to its superior linearity. Further, feed-forward paths in the first baseband (BB) stage can create a complex input impedance [4], serving to absorb the capacitance of the cancellation circuits into the RX input match.

An 8-phase design is chosen to provide rejection for strong emissions around the 3rd and 5th harmonics produced by the TX. The passive mixer switches of the RX can be sized trading off R_{ON} and C_{ON} to minimize the small signal noise figure. The first baseband amplifier is then the key block setting the RX noise figure (NF).

If an op-amp is used as the baseband amplifier, the baseband impedance is set by the baseband feedback resistance R and amplifier gain A as $R/(1+A)$. The baseband noise is due to the matching resistors, and the amplifier itself. By increasing the amplifier gain, the feedback resistor can be increased while maintaining a fixed input match, lowering the input referred voltage noise due to the resistors [5]. The input referred voltage noise from the amplifier can be lowered at the expense of power consumption. The thermal noise of the PMF baseband thus has no *fundamental* lower bound. Note that this increased gain trades off noise and linearity - given a fixed input impedance match and fixed power supply, the RX compresses under smaller input power for an increased gain. Given this trade off, the amplifier is designed for a gain of 22dB, and an input referred voltage noise of $1nV/\sqrt{\text{Hz}}$.

Fig. 7 shows the first, two-stage amplifier that achieves the target noise with high linearity. The input is a complementary transconductor. The tail current sources (MT1-2) allow for mid-rail input and output voltages, while biasing the input pair in weak inversion for >30mS G_m at <2mA current. The second stage is a class-AB stage placed in shunt feedback. The class-AB output is biased with a CMOS floating battery (MB1-2) and driven by the first stage through complementary source followers. The low impedance of the source followers at the class-AB gate node lowers the noise contribution from the floating battery bias. The second stage provides rail-to-rail swing from the 1.5V power supply, while the local shunt-feedback suppresses the voltage swing at the first stage output, allowing for high linearity. The input-referred thermal noise of the second stage amplifier is suppressed by the input pair's G_m, allowing for the overall amplifier noise to be lowered at the cost of power consumption.

The baseband second stage provides both harmonic recombination and second-order filtering via a multi-feedback (MFB) biquad, with the feed-forward resistor

TABLE I
RX PERFORMANCE SUMMARY

RF Freq (GHz)	1-2	NF (Lossless Balun) (dB)	3.6
BW(MHz)	20	OOB P1dB (dBm)	+5
S11 (dB)	<-20	IB P1dB (dBm)	-20
Gain (dB)	35	OOB IIP3 (dBm)	+25
Supply (V)	1.5	IIP2 (dBm)	+66
Power (mW)	64	HR ($3^{rd}/5^{th}$) (dB)	48/53

TABLE II
COMPARISON TO PRIOR ACTIVE CANCELLATION FDD SYSTEMS

Specification	[2]	[6]	This Work
RF Freq (GHz)	.3-1.6	.8-1.5	1-2
Frontend Cancellation (dB)	25	40	64
Backend Cancellation (dB)	-	-	25-30
TX Power at Ant, 1dB RX Compression (FDD) (dBm)	+14[1]	+15[1]	+17[1]
System Cascade NF (dB)	11-16[1]	4.8-6.3[1]	7-15.4[2]
External Isolation (dB)	0	35	0

[1] F_{Duplex}=115MHz, [2] F_{Duplex}=40MHz

segmented to provide weighted current summation of the 8 input phases, as shown in Fig. 5.

VII. MEASUREMENT RESULTS

The 2.5mm x 2.5mm chip is implemented in TSMC 65nm GP technology. The RX's measured performance is summarized in Table I. The input match can be tuned to <-20dB S11 from 1GHz to 2GHz by tuning an input capacitor DAC, the first stage feedback resistor, and the first stage amplifier gain. Additionally, the RX's tunable complex input impedance can absorb up to 2pF of parasitic capacitance from the self-interference cancellation circuits into the input match. At a gain setting of 35dB, the RX maintains an OOB IIP3 of +25dBm, and a NF of 3.6dB for an RX BW of 20MHz, while providing over 35dB filtering of blockers at 100MHz offset, 48dB of third HR, and 53dB of fifth HR.

In Fig. 8, the system NF contributions when the TX is enabled are plotted against TX power at the antenna for FDD mode. The RX linearity enables FDD operation up to +17dBm TX power at 115MHz offset and +15dBm at 40MHz offset without a duplexer, FD operation up to +5dBm without a circulator, and FD operation up to +13dBm with a circulator before 1dB RX gain compression, as shown in Fig. 9. The transceiver's FDD performance is summarized against prior art in Table II.

VIII. CONCLUSION

This work addresses two key limitations in active cancellation techniques for FD/FDD systems, namely cancellation of RX band noise, and simultaneous TX/RX operation at >+15dBm TX power with a fully integrated antenna interface. The first limitation, RX band noise, is addressed through thermal noise feedback loops around

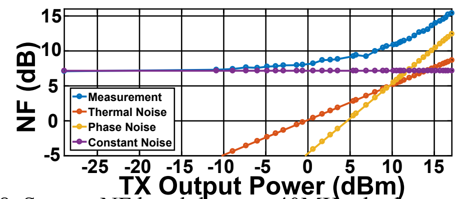

Fig. 8. System NF breakdown at 40MHz duplex spacing.

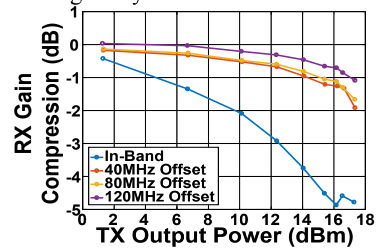

Fig. 9. RX gain compression vs. TX power and TX/RX spacing.

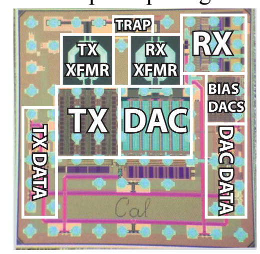

Fig. 10. Die photo.

the analog cancellation network, digital prediction of the TX's RX band quantization noise, and feed-forward phase noise cancellation. The second limitation is addressed through a passive-mixer-first receiver design which provides >+25dBm OOB IIP3 from a single 1.5V supply.

Extension of these techniques, such as additional resolution on the replica DAC, higher oversampling ratio, and DAC noise shaping provide a path for fully-integrated, high-power, high-sensitivity FD/FDD transceiver systems.

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