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Enabling Wireless In-band Full-duplex

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***Abstract*—This paper presents a baseband model and an enhanced implementation of the wireless full duplex analog method introduced by [1].Unlike usual methods based on hardware and software self- interference cancelation, the proposed design relies on FSK modulation. The principle is when the transmitter of a local end is sending data by modulating the carrier with the appropriate frequency deviation, its own receiver is checking if the remote transmitter is using the opposite deviation. Instead of architectures often used by both non-coherent and coherent receivers that require one filter (matched filter for coherent detection) for each frequency deviation, our design uses one mixer and one single integrator-decimator filter. We test our design using Universal Software Radio Peripheral as radio frequency front end and computer that implements the signal processing methods under free and open source software. We validate our solution experimentally and we show that in-band full duplex is feasible and synthesizable for wireless communications.**

***Keywords*—In-band full-duplex, Wireless, USRP, SDR, GNU radio**

I. Introduction

DUE to the rapid development of wireless communications in recent years, demand on wireless spectrum has been growing dramatically. The new devices call for wireless communication techniques with high spectral efficiency. Current wireless devices work on half duplex mode or use two channels to achieve bidirectional communication which results in poor spectral efficiency. It was believed that full duplex mode is not possible for wireless communication because of the self interference that results [2]. If its possible, wireless communications could cut their spectrum needs by half and the problems such as exposed terminal, hidden terminal and fairness issue will be solved [3]. In-band full-duplex radio has emerged as an attractive solution for spectral congestion in reality. Researchers in academia and industry have proposed designs based on software and hardware cancellation of self transmitted signal. The patent [1] has introduced a new method based on frequency modulation. Our method extends the idea behind this patent and enhances the proposed solution by using just one signal mixer and one integrator decimator filter. We implement our design using USRP-2901 National Instruments as RF front end and GNU radio software toolkit for baseband signal processing algorithms. The tests held on our laboratory show that the design works as expected.

The first full duplex narrow-band wireless communication experiments were held in 1998 [4]. Researchers were able to achieve 72dB duplex isolation between an AM and FM signal over 200 kHz radio channel by using a dual-antenna and an adaptive RF echo canceller. Since then, few researchers dedicate to implement full duplex for wide-band wireless communication. In 2007, researchers from MIT Lincoln Laboratory demonstrate a signal processing technique that uses antennas diversity for self interference cancelation [5]. Due to nonlinearities in the receiver system caused by the proximity of the receive antennas to the receiver hardware, the maximum isolation ratio was 60 dB over 100 kHz radio channel. In the same year, researchers have deposed a pattent, which has been published two year later in 2009, introduces a new technique based on frequency modulation instead of self interference cancellation [1]. Other team from Lund University has introduced in 2010 a cancelation method for full duplex relays using antenna with special directionality [6]. At same time, Rice University has published a study of three types of self interference cancellation mechanisms using off-the-shelf MIMO radios [7]. They have concluded that wireless full-duplex systems are feasible and can achieve rates larger than the rates achieved by half-duplex systems. Besides antenna cancellation mechanism, researchers from Stanford University have combined antenna cancellation with RF interference cancellation and digital cancellation to bring self-interference to within a few dB of the noise floor [8]. Same team has extended their work in 2011 [9]. They could reduce self-interference by up to 73dB for a 10MHz OFDM signal by replacing antenna cancelation mechanism with balun transformer. Two years later, in 2013, other team from Stanford University introduced a new approach. They designed a circulator in analog cancelation and cancelled linear and non linear components by digital cancelation [10]. Finally, they achieved to cancel 110 dB of self-interference in WIFI band and have implemented their system in 15 cm x 15 cm board. In 2014, active cancelation techniques have been introduced by [11] where an auxiliary transmit chain is employed to create the cancellation signal. In 2016, researchers from Columbia University implemented the first magnetic-free CMOS based circulator chip [12] which then used on a full duplex radio [13], [14], [15]. Until today, this team is still publishing more enhancements of their original work [16], [17], [18], [19], [20].

II. Method

Solution proposed in [1] is based on BFSK modulation where carrier is changed by a frequency deviation according to the input data. Digital message state for a binary 1 and for a binary 0 are represented by two different frequencies slightly offset F_d from the carriers frequency F_c with constant amplitude. By convention, symbol 1 is presented by the higher carrier frequency called mark frequency ($F_{mark} = F_c + F_d$) while symbol 0 is presented by the lower carrier frequency

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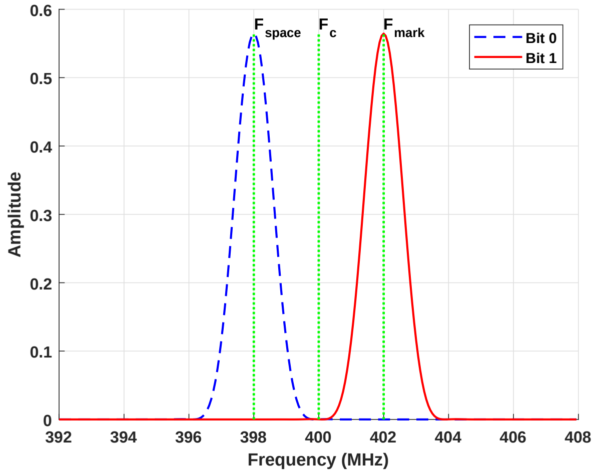

Fig. 1. Pattern of BFSK RF channel spectrum

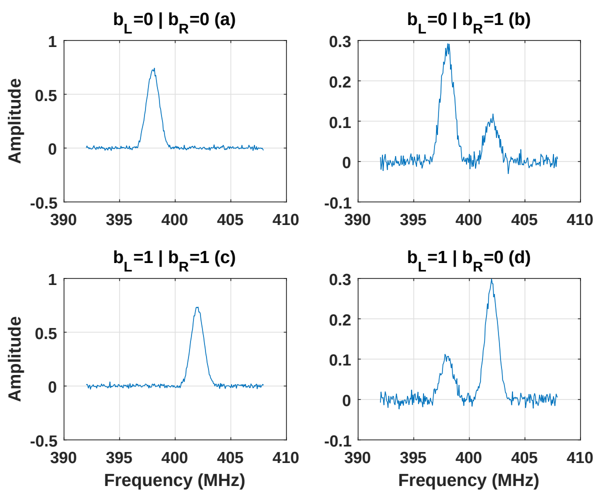

Fig. 2. Channel spectrum using two identical transceivers

called space frequency ($F_{space} = F_c - F_d$). Each frequency state lasts for a single bit period T_b (Fig.1). There is only one frequency on the channel while using a half duplex BFSK communication. If there are two transceivers called local-end and remot-end use the same frequency carrier and transmit data to each other at the same time, channel can contain both space and mark frequencies. Receiver can decode the signal using its own transmitted data. If the local transmitter is sending a symbol noted b_L (0 or 1), and there is just space or mark frequency on the local receiver signal (Fig.2 (a) and (c)), the remote end is sending the same symbol noted b_R ($b_R = b_L$). If there are both space and mark (Fig.2 (b) and (d)), the remote end is sending the inverse symbol ($b_R = \overline{b_L}$).

The proposed method on [1] is an analog solution for pass-band transmission. It uses two pass-band filters around the space and mark frequency (Fig.3). Assuming the local transmitter is sending a 1, receiver will check the signal power at the space filters output. If its above a given threshold; which means the presence of space frequency on the channel

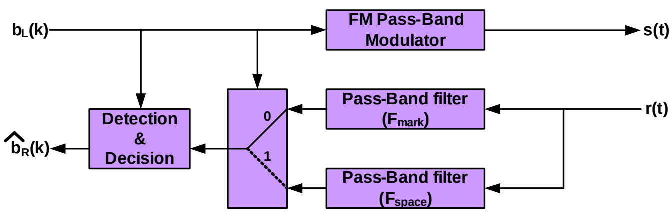

Fig. 3. Full duplex FM top level

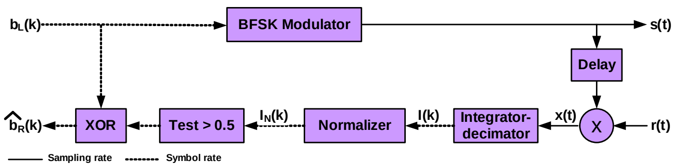

Fig. 4. Full duplex BFSK top level

while the local transmitter is producing a mark; receiver will deduct that the remote end is sending a 0. Otherwise, there is only a mark frequency on the channel and receiver will deduct that the remote end is sending the same data which is 1. The new method is designed for base-band processing. It enhances the design by replacing the two pass-band filters by a mixer and one integrator-decimator filter which reduce the implementation cost (Fig.4). Received signal is multiplied by the transmitted signal and integrated over one symbol period. For each symbol period, integrator-decimator block outputs one value (decimation) which then normalized using normalizer block. After that, the normalized value is compared with 0.5. if its below the threshold, both transceivers is sending the same bit. Otherwise, the normalized value is above the threshold and transceivers are sending opposite bits. We present in the next section the mathematical equations and signal processing methods for the proposed design.

III. BASEBAND SIGNAL PROCESSING

The frequency of BFSK signal is controlled by transmitted bit $b(k)$. Using base-band presentation, the instantaneous frequency bounces between $-F_d$ and $+F_d$ according to the transmitted bits. By convention, the frequency is set to F_d when $b(k)$ is 0 and to $+F_d$ when $b(k)$ is 1. BFSK signals frequency can be expressed as (3) where $a(k)$ is symbol state and $h(t)$ is unit pulse function defined respectively by (1) and (2). By definition, the instantaneous frequency F(t) is derivation function of the phase $\theta(t)$. Phase can be calculated by integrating the frequency function (4) and the complex baseband signal waveform can be as expressed like (9); where m is modulation index; using (5), (6), (7) and (8).

$$a(k) = \begin{cases} +1 & \text{if} \quad b(k) = 1 \\ -1 & \text{if} \quad b(k) = 0 \end{cases} \tag{1}$$

$$h(t) = \begin{cases} 1 & \text{if} \quad 0 \leq t \leq T_b \\ 0 & \text{if not} \end{cases} \tag{2}$$

$$F(t) = F_d . \sum_{i=0}^{\infty} a(i) h(t - iT_b) \tag{3}$$

$$\theta(t) = 2\pi \int_0^t F(u)du \quad (4)$$

$$m = 2.F_d.T_b \quad (5)$$

$$\theta(t) = 2\pi.a(k).F_d.t + m.\pi.\sum_{i=0}^{k-1} a(i) \quad (6)$$

$$\Phi(k) = m.\pi.\sum_{i=0}^{k-1} a(i) \quad (7)$$

$$\theta(t) = 2\pi.a(k).F_d.t + \Phi(k) \quad (8)$$

$$s(t) = e^{2j\pi.a(k).F_d.t}.e^{j.\Phi(k)} \quad (9)$$

Assuming there are two transceivers sending BFSK signal at the same time over the same frequency channel. The received signal on each transceiver will contain both transmitted signals with the appropriate attenuations (α_L and α_R) and delays (τ_L and τ_R). In order to simplify nomination, signals coming from local transceiver are indexed by L (Local end) while signals coming from the remote transceiver will be indexed by R (Remote end). Over a perfect channel without any noise and just one single path, received signal $r(t)$ is modeled in (10). The fist term presents the self interference part and the second is the remote end transmitted signal which presents the useful part. Its clear that traditional receivers cannot demodulate this signal due to the interference term. To resolve this ambiguity, our transceiver uses its own transmitted data to demodulate the received data. If both transceivers send the same data, there is just one frequency, otherwise there are two frequencies. In fact, each transceiver knows the interference terms frequency. The receiver can recover the remote end transmitted data by checking the presence of the second frequency.

$$r(t) = \alpha_L s_L(t - \tau_L) + \alpha_R s_R(t - \tau_R) \quad (10)$$

In order to avoid the usage of two pass-band filters, transceiver uses its own transmitted signal to produce a continuous component in frequency spectrum. The received signal is mixed with the local transmitted signal delayed by τ_L which is a known deterministic delay depends on transceiver hardware (11).The mixer output noted $x(t)$ contains two terms (12), the first one is the square root of self interference signal noted $x_L(t)$ (13) and the second one is the product of local end and remote end transmitted signals noted $x_R(t)$ (14). If we look at frequencies, $x_L(t)$ frequency is $\pm 2.F_d$ while $x_R(t)$ has two possibilities 0 or $\pm 2.F_d$.

$$x(t) = s_L(t - \tau_L).r(t) \quad (11)$$

$$x(t) = x_L(t) + x_R(t) \quad (12)$$

$$x_L(t) = \alpha_L s_L^2(t - \tau_L) \quad (13)$$

$$x_R(t) = \alpha_R s_L(t - \tau_L)s_R(t - \tau_R) \quad (14)$$

In order to eliminate $x_L(t)$, receiver integrates $x(t)$. we define $I_L(k)$, $I_R(k)$ and $I(k)$ as integration of $x_L(t)$, $x_R(t)$ and $x(t)$ between $k.T_b + \tau_L$ and $(k + 1).T_b + \tau_L$ (15, 16 and 17). By developing the calculation of $I_L(k)$ using (18), we find out $I_L(k)$ is always null if m (modulation index) is an integer (19).

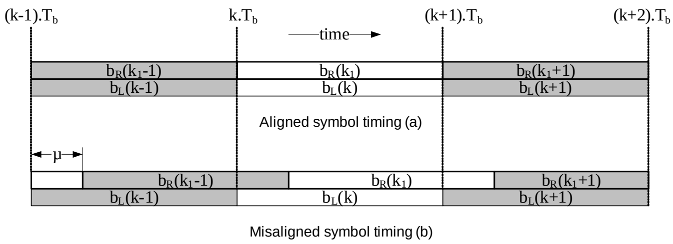

Fig. 5. Symbol timing alignment

$$I_L(k) = \int_{k.T_b+\tau_L}^{(k+1).T_b+\tau_L} x_L(t)dt \quad (15)$$

$$I_R(k) = \int_{k.T_b+\tau_L}^{(k+1).T_b+\tau_L} x_R(t)dt \quad (16)$$

$$I(k) = I_L(k) + I_R(k) \quad (17)$$

$$\int_{T_1}^{T_2} e^{2j\pi.a.f.t}dt = (T_2 - T_1) sinc(\pi a.f.(T_2 - T_1)).e^{j\pi a.f.(T_2+T_1)} \quad (18)$$

$$|I_L(k)| = \alpha_L T_b |sinc(m\pi)| \quad (19)$$

By setting m to an integer, receiver can recover the data by decoding the value of $I(k)$ which is equal to $I_R(k)$. $I_R(k)$ calculation depends on τ which is difference between τ_R and τ_L. Basically, τ is the resultant delay between local and remote end. It depends on physical distance between transceivers and the misalignment between symbol period timing of the two transceivers (Fig.5). Generally, it can be written as sum of multiple of T_b and fractional delay μ smaller than T_b (20).

In this work, we assume that symbol period timing of the two ends is aligned (Fig.5.a). In this case both a_L and a_R remain constant over one symbol period between $k.T_b + \tau_L$ and $(k + 1).T_b + \tau_L$. $I_R(k)$ calculations result (22) shows that the module of $I(k)$; since m is an integer; takes two values according to the transmitted bits. If the both transceivers send the same bits ($a_L(k) + a_R(k1) = \pm 2$), the value will be $I_{MIN} = 0$. Otherwise, its equal to $I_{MAX} = \alpha_R.T_b$.

$$\tau = \tau_R - \tau_L = p.T_b + \mu \quad (20)$$

$$k_1 = k - p \quad (21)$$

$$|I_R(k)| = \alpha_R T_b |sinc(m\frac{\pi}{2}[a_L(k) + a_L(k_1)])| \quad (22)$$

At this step, receiver has all the required information to decode remote ends transmitted bits. If an appropriate bit scrambling is adopted, there is no chance of both transceivers send a consecutive equal bits. The receiver can buffer a specific number of integration result and calculates the maximum and minimum. The integrator output will then normalized (23) and the result will be 0 or 1. In order to tolerate noise and impairments, the normalized value $I_N(k)$ is compared to 0.5. If its below the threshold, the receiver decides that remote end sends the same data as its own transmitter. Otherwise, it sends the opposite data (Table I). The received data in this case is

TABLE I
COMPARATOR OUTPUT AND RECOVERED BITS IN FUNCTION OF $b_R(k)$ AND $b_L(k)$

$b_R(k_1)$	$b_L(k)$	$Test(I_N(k) \geq 0.5)$	$\hat{b}_R(k)$
0	0	0	0
0	1	1	0
1	0	1	1
1	1	0	1

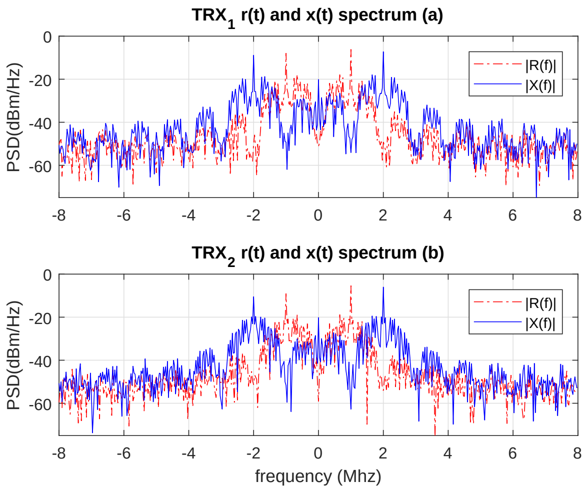

Fig. 6. Received and mixed signal spectrum

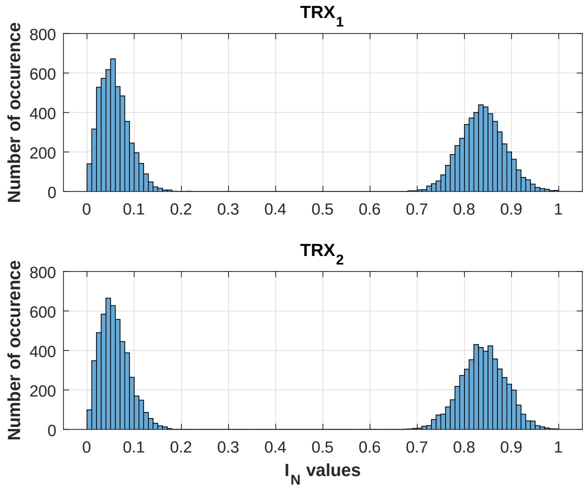

Fig. 7. I_N distribution (10000 samples)

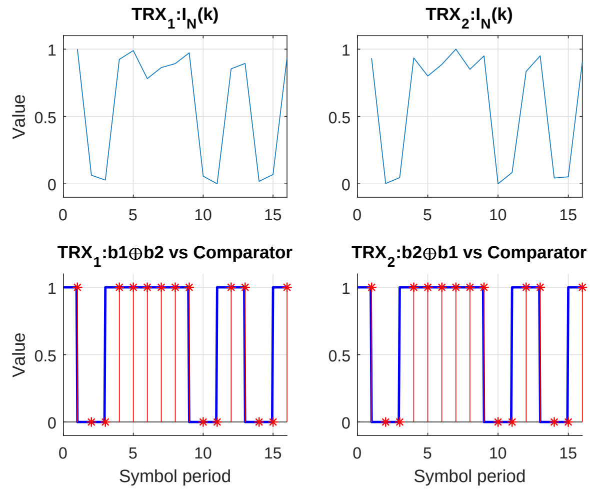

Fig. 8. Integrator and comparator outputs

simply the logical XOR between the local transmitted data and the comparator output (24).

$$I_N(k) = \frac{|I(k)| - I_{MIN}}{I_{MAX} - I_{MIN}} \tag{23}$$

$$\hat{b}_R(k) = b_L(k) \oplus Test(I_N(k) \geq 0.5) \tag{24}$$

IV. SIMULATION

We have modeled and simulated the proposed design using MATLAB. We have used discrete time model with a sampling rate of 16 Mhz to evaluate the simulation results of a 1 Mb/s RF link. The model simulates a single channel full duplex transmission of two bits sequences between two synchronized transceivers TRX_1 and TRX_2 ($\mu = 0$) over AWGN (Additive White Gaussian Noise) channel. The SNR (Signal Noise Ratio) of self interference part is 40 dB while the SNR of remote signal is 10 dB. Fig.6 shows the frequency spectrum of received signal $R(f)$ and mixer output $X(f)$ in both transceivers corresponding to a full duplex transmission of 128 bits using modulation index $m = 2$ ($F_d = 1$ MHz). As expected, there are two lobes at F_d and $+F_d$ in received signal spectrum before mixing while the mixer output contains three lobes at $-2F_d$, 0 and $+2F_d$.

Modulation index is set then to 1 which is the smallest value supported by the proposed method. After self interference term elimination by integrating the received signal, the normalized integrator output in both transceivers is around two values 0 or 1. The distribution of I_N value is shown in Fig.7 . Its clear that at $SNR = 10$ dB for remote signal, there is no ambiguity detected along 10000 symbols. All I_N values are far from decision threshold 0.5. In this case, comparator output ($Test(I_N \geq 0.5)$) is equal to $b_R \oplus b_L$ at each symbol period (Fig.8). Fig.9 presents 16 recovered bits by comparing the comparator output with b_L using a logical XOR and shows the matching between the recovered and the transmitted bits sequences.

By decreasing the SNR, the lobes around 0 and 1 become wider and there edges are closer to 0.5. Fig.10 shows the distribution of I_N values in one transceiver for different three values of SNR (8 dB, 5 dB and 3 dB). The receiver can recover the data even with an SNR of 5 dB. But for SNR equal to 3 dB. The two lobes overlap and the comparator output cannot be trusted.

V. SDR IMPLEMENTATION AND RESULT

We have implemented our design using SDR (Software Defined Radio) technology where components that have been

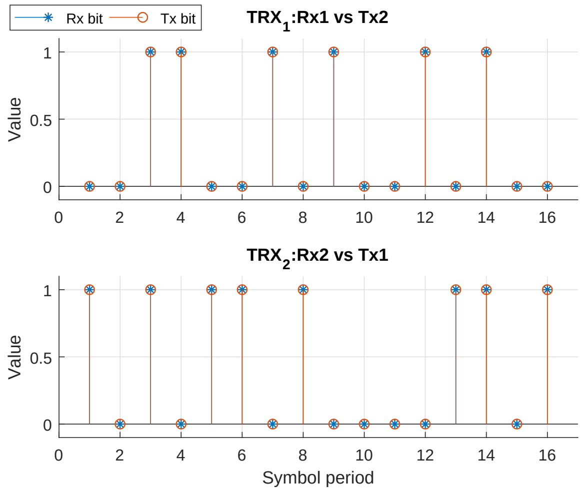

Fig. 9. Rx bits vs Tx bits

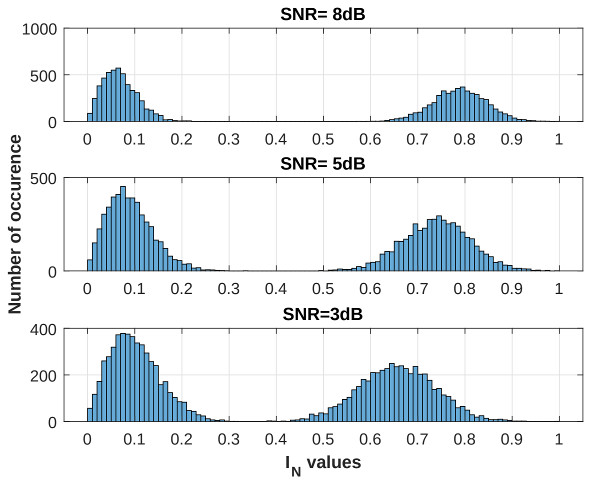

Fig. 10. I_N distribution vs SNR (10000 samples)

traditionally implemented in hardware are instead implemented by means of software on a computer. In our case, SDR hardware must provide both transmitter and receiver. Discrete signal generated on computer side is sent to RF front-end hardware which converts the discrete signal to analog by means of DAC (Digital to Analog Converter). Analog signal is then converted to RF signal using an analog mixer and broadcasted into the air trough Tx antenna. RF signal received by Rx antenna is down converted from a predefined RF frequency to the baseband and converted to digital signal by means of ADC (Analog to Digital Converter) before is sent to computer for signal processing. We have used National Instruments USRP-2901 as RF front-end which provides two bidirectional separated channels that allow us to implement and test two transceivers using one USRP and one computer [21]. USRP-2901 is made by Ettus Company. Based on the mapping between National Instruments and Ettus products [22], the corresponding Ettus USRP in our case is B210 [23]. This feature allows us to use GNU radio which is an open and free software toolkit designed for SDR applications.

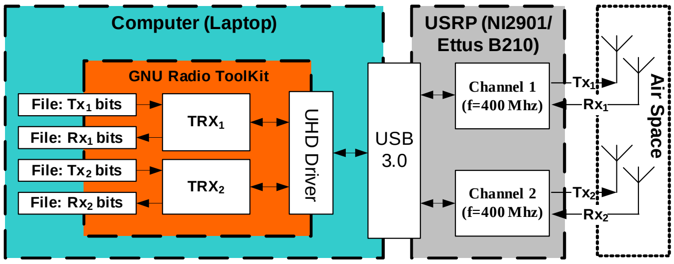

Fig. 11. Test implementation work flow

GNU radio library contains UHD driver to support Ettus SDR platforms [24]. As already mentioned, transmitter and receiver of the same transceiver should have the same time reference and the delay τ_L between them must be known. In order to satisfy those two conditions, we have made modification in GNU radio library to force TX and RX start at the same time which has allowed us to estimate τ_L based on some calibration tests.

In order to ensure the synchronization between transmitters, we have taken advantage of the two channels provided by USRP-2901 to implement two transceivers on the same PC and the same USRP. Fig.11 shows the work flow of this implementation. There are two bits sequences stored in two separated files Tx1-bits and Tx2-bits. Each bits sequence is modulated and transmitted to USRP through USB3.0 using respectively TRX_1 and TRX_2 blocks which are the discrete implementation of the proposed algorithms under GNU radio (modulation index $m = 1$). Both channels of USRP-2901 are set to the same radio frequency 400 Mhz to produce the interference. Signals Received by RX_1 and RX_2 antennas are transformed to the baseband and sent respectively to TRX_1 and TRX_2. Each TRX_i $(i = 1; 2)$ module decodes the data and stores them to Rxi-bits file. The bits sequence stored in Rxi-bits file is always equal to the bits sequence sorted in Txj-bits. This test proves that TRX module is able to recover the data without any error under the predefined conditions.

VI. Discussions

Instead of usual radio transceivers that use two separated channels for downlink and uplink direction, our solution uses one single channel for both directions. The most important advantage of full duplex radio is bandwidth optimization by cutting the spectrum needs by half. Compared with FSK transceiver that uses the same modulation index of our transceiver, the proposed design requires the half of normal bandwidth. Unfortunately, the minimum modulation index that our design can achieve is 1. However, normal FSK transceiver can achieve modulation index of 0.5 by using MSK technique which is the smaller possible modulation index. This fact leads us to compare our solution with MSK technique to see if our design is always the optimal choice for FSK wireless communication. Bandwidth requirements of single channel FSK communication B_{single} can be estimated by (25) using Carson rule [25]. Using the defining of modulation index m, B_{single} can be written as (26) in function of m and T_b. Based on this rule, bandwidth requirement note B

($B = B_{single}$) of our design is $3/T_b$ which always better than using double channels transceiver. In fact the smallest bandwidth requirement ($B = 2.B_{single}$) that double channel FSK transceiver can achieve is $5/T_b$ when using MSK. Based on spectral efficiency (noted ξ) parameter, which represents the bit rate can be transmitted over 1 Hz bandwidth (27), the maximum spectral efficiency that double channel FSK transceiver can achieve is 0.2 bit/s/Hz when $m = 0.5$. Our design spectral efficiency is 0.33 bit/s/Hz even using m=1. To sum up, the required bandwith for full-duplex wireless transmission using usual MSK transceiver can be acheived by our proposed design only using a modulation index $m = 3$.

$$B_{single} = 2.(F_d + \frac{1}{T_b}) \tag{25}$$

$$B_{single} = \frac{m+2}{T_b} \tag{26}$$

$$\xi = \frac{1}{B.T_b} \tag{27}$$

VII. Conclusion

This paper proves that the single channel full duplex is possible and synthesizable for FSK wireless systems. Instead of usual radio transceivers that use two separated channels for downlink and uplink direction, our solution uses one single channel for both directions. Compared with FSK transceiver that uses the same modulation index of our transceiver, the most important advantage of full duplex radio is bandwidth optimization by cutting the spectrum needs by half. Compared with the original method, the enhanced method reduces the implementation cost by using just one mixer, one decimator-integrator filter and simple combinatorial logical blocks. However, the method assumes that both transmitters are synchronized with the same clock. Timing synchronization is still under studying and was not discussed on this paper. Also the performance and reliability of the system where not evaluated.

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