Wideband Tx Leakage Cancellation using Adaptive Delay Filter at RF Frequencies

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Abstract — **Transmitter leakage is caused by limited isolation between Tx and Rx paths of an frequency-division duplexing (FDD) transceiver due to use of non-ideal duplexers. As the leakage can seriously impair the performance of the receiver, it is important to increase the isolation of the existing system. A portion of the leakage falling into the Rx band can't be removed using conventional band-pass techniques, which leads to exploration of active cancellation methods. This paper presents an active Tx leakage cancellation structure with adaptive delay filter, together with a related design methodology. Based on S-parameters characterisation of a SAW duplexer, simulation results show that for 20 MHz bandwidth more than 25 dB cancellation can be achieved, for LTE band 8 carrier. The complexity of proposed cancellation structure is proportional to the characteristics of a duplexer used in the transceiver. Using finite number of elements, the number of auxiliary paths needed depends on how complex the characteristics of the duplexer are within the band of interest.**

Keywords — **Leakage cancellation, full duplex radios, FDD, adaptive delay filter**

I Introduction

A duplexer allows a frequency-division duplexing (FDD) transceiver to transmit and receive signals simultaneously. Ideally, the isolation between the transmitter (Tx) and the receiver (Rx) should be infinitely large. In practice however, there is a portion of transmitted signal being injected to the receiver, known as Tx leakage, consisting of in-band and out-of-band components. The out-of-band leakage reduces the sensitivity of the receiver, especially in a zero-IF architecture. It also increases the linearity requirement of the Rx front-end due to its significantly high power levels in comparison to the received signals. Furthermore, in the case when Tx leakage is large, the LNA and mixer can then be forced into saturation resulting in desensitization of the Rx front-end [1]. Conventionally, the out-of-band Tx leakage can be suppressed by placing a sharp band pass filter at the receiver between the LNA and the down converter. Such solution limits the flexibility of the receiver, since high performance band-pass filters have a relatively poor tunability.

The in-band Tx leakage can be a more serious issue as the leakage falls within the same frequency band as Rx signal. In this case, a conventional filtering method cannot separate such interference from the signal of interest, resulting in Rx signal being swamped in noise. It is therefore not possible for the receiver to capture the signal of interest directly. For the situations where passive filtering approach is no longer feasible, active cancellation methods have to be employed.

There are many publications available on reduction of the Tx leakage using active cancellation [2]-[8]. The baseband cancellation presented in [2] has a relatively limited performance in a presence of power amplifier intermodulation products or when a Rx front-end saturates due to high leakage power. The methods presented in [3]-[7] use one or two auxiliary signal paths to cancel the leakage of a circulator type of duplexer. Each secondary path consists of one fixed delay line and tunable attenuator. When the circulator is replaced with a standard duplexer, the performance of the above methods decreases dramatically. This is because a circulator generally has a flatter magnitude and more linear phase response within its operational bandwidth in comparison with a standard duplexer. As a result, the circulator leakage can be improved using no more than two auxiliary signal paths over a relatively narrow frequency range. The analog cancellation from [8] provides an improved cancellation performance with 16 auxiliary paths. However, when the circulator is replaced by a standard duplexer, for instance, a SAW duplexer, the number of paths required is much larger to achieve the same performance. On the other hand, if the phase characteristics of the circulator can be characterised, less number of paths could be sufficient.

This paper proposes a Tx leakage cancellation method deployed directly at RF. It combines the adaptive delay filter proposed in [9] directly with highly tunable delay line/attenuation circuitry at RF. This method has a potential for providing leakage cancellation over a wide bandwidth when a standard duplexer is used. Additionally, the proposed structure is immune to high order distortions caused by the PA. The paper is organized as follows. Section II proposes the structure of the cancellation method. Section III explains the concept behind the proposed structure. Section IV presents the simulation results.

II Proposed Tx leakage cancellation structure

The proposed active Tx leakage canceller is shown in Fig. 1 and consists of RF and baseband calculation sections, respectively. In the RF section, the reference signal is taken from the output of the PA and is used to cancel the leakage in a feed-forward manner. There are M cancellation paths (taps) in the form of a tuneable delay module (d_i, where i is the tap index, $i = 1, 2, \ldots, M$) and a variable attenuator (w_i). The signals from each path are added together to form the reconstructed signal, which is then subtracted from the leakage signal (then system adaptation is completed). The delays are determined on tap-by-tap basis. That is, the delay of the later tap will not be determined until the ones of the previous taps have been estimated for a given error. This can be done by maintaining the attenuator and delay of the determined taps unchanged, then setting the attenuation of the tap to be processed to the lowest and keeping the attenuator of undetermined paths to the highest attenuations, respectively. After the delay of one tap has been estimated, the attenuation level (tap weight) of the current tap needs to be determined before next path can be processed.

All the calculations and estimations are processed at baseband. The receiver needs to be set to off-line mode for estimating the delay for each cancellation path. When the estimation of all delays is finished, the receiver can then be turned back to the on-line mode. At this stage, attenuations of all paths can be jointly adapted to compensate for minor changes of the characteristics of the duplexer due to varying of the environment, such as change in the temperature, etc.

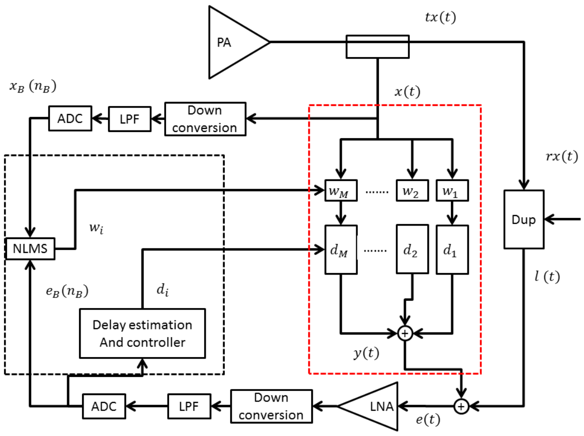

Fig. 1: Proposed structure of Tx leakage canceller with adaptive delay filter

III Estimation of delays and adaptation of tap weights

Let t_{sB} be the baseband sampling time. By assuming the RF signal is sampled with sampling time of t_{sRF}, the ratio between t_{sB} and t_{sRF} is

$$r = \frac{t_{sB}}{t_{sRF}} \tag{1}$$

Let D_i be the sample delay at RF for each tap i. The actual RF delay is then:

$$d_i = D_i \cdot t_{sRF} \tag{2}$$

a) Estimation of delays

The delay of each tap at RF is estimated at baseband in two steps. Firstly, a coarse estimation is performed by finding the cross-correlation between the baseband error signal before being cancelled by the current tap ($e_{pB}(n_B)$), and the baseband reference signal ($x_B(n_B)$):

$$corr_{x_B e_{pB}}(d) = x_B(n_B) \star e_{pB}(n_B) \tag{3}$$

where n_B is the baseband discrete time index. The estimated sample delay for tap i at baseband, D_{iB}, is the one which maximizes the absolute value of the correlation. Then the true sample delay at RF,

D_i, is in the range of

$$\begin{cases} 0 \le D_i \le \left\lfloor \frac{r}{2} \right\rfloor, & D_{iB} = 0 \\ -\left\lfloor \frac{r}{2} \right\rfloor < D_i - r \cdot D_{iB} \le \left\lfloor \frac{r}{2} \right\rfloor, & D_{iB} \neq 0 \end{cases} \tag{4}$$

where $\lfloor * \rfloor$ is the floor operator.

To determine what exact value D_i is, correlation in (3) must be re-investigated.

$$\begin{aligned} corr_{x_B e_{pB}}(d) &= x_B(n_B) \star e_{pB}(n_B) \\ &= x_B(n_B) * e_{pB}^H(-n_B) \end{aligned} \tag{5}$$

where $*$ is the convolution operator. Taking the Discrete Fourier Tranform of both sides in (5), we have:

$$\begin{aligned} CORR_{x_B e_{pB}} &= \mathcal{F}[x_B(n_B) * e_{pB}^H(-n_B)] \\ &= X_B \cdot E_{pB}^H. \end{aligned} \tag{6}$$

Because E_{pB} is the output of X_B through a linear system, we have

$$E_{pB} = A \cdot e^{j(2\pi f D_{iB} t_{sB} + \varphi)} X_B \tag{7}$$

where φ is the baseband phase misalignment casued by the coarse delay estimation of D_{iB}. Substiting (7) into (6), we have:

$$\begin{aligned} CORR_{x_B e_{pB}} &= A \cdot e^{-j(2\pi f D_{iB} t_{sB} + \varphi)} X_B \cdot X_B^H \\ &= A \cdot e^{-j\varphi} e^{-j2\pi f D_{iB} t_{sB}} X_B \cdot X_B^H. \end{aligned} \tag{8}$$

Take the inverse of Discrete Fourier Transform of both side, we then get:

$$corr_{x_B e_{pB}} = A \cdot e^{-j\varphi} corr_{x_B x_B}(n - D_{iB}). \tag{9}$$

A maximum of the absolute value of $corr_{x_B e_{pB}}$ occurs when x_B is delayed by D_{iB}. The phase of such peak point is $-\varphi$. We want the phase misalignment φ equal to $k\pi$,where k is any integer number, i.e., $k = -\infty, \cdots, -1, 0, 1, \cdots, +\infty$. This cannot be achieved by delaying the reference signal x_B by integer number of samples at baseband. Therefore, we need to compensate such phase misalignment by multiplying the the reference signal by the phase compensator of $e^{-j2\pi f_c D_i t_{sRF}}$ where f_c is the centre frequency of the band of interest, so that the reference signal becomes:

$$\hat{x}_B(n_B) = x_B(n_B) \cdot e^{-j2\pi f_c D_i t_{sRF}}. \tag{10}$$

The compensator is accurate only when $f = f_c$ and it is getting less accurate when the frequency is away from the centre. This is the reason why a number of taps are needed to achieve a better cancellation performance. The more complex the unknown system is, the larger number of taps is required.

(9) now becomes

$$corr_{\hat{x}_B e_{pB}} = A \cdot e^{j\xi} corr_{x_B x_B}(n - D_{iB}) \tag{11}$$

where $\xi = 2\pi f_c D_i t_{sRF} + \varphi$. We want

$$\xi = k\pi. \tag{12}$$

Ideally, the solution for D_i in (12) is the desired delay at RF, i.e.,

$$D_i = \frac{k\pi - \varphi}{2\pi f_c t_{sRF}}, \tag{13}$$

and it must also satisfy the condition shown in (4). In fact, it may not be possible to find a integer solution for D_i. Therefore, the actual sample delay should be the one that makes ξ closest to $k\pi$.

b) Adaptation of tap weights

Normalised least mean square (NLMS) algorithm is used to adapt the weights for both tap-by-tap offline and online modes. However, the standard NLMS adaptation equations for complex signals [12] cannot be used here. This is because the weights in the standard equations are inherently complex-valued. In the proposed cancellation structure, real-valued weights are needed. Furthermore, the described algorithm is calculated at baseband but executed at RF. As a result, any phase mismatches caused by the down-conversion have to be compensated before being used for weights adaptation.

The band-limited error signal for tap i is given by:

$$e_c(n) = e_p(n) - w_i \cdot x(n - D_i). \tag{14}$$

where $e_c(n)$ and $e_p(n)$ are error signals generated by current and previous taps, respectively. When $i = 1$, $e_p(n) = l(n)$. Since the adaptation of tap weights occurs after the esimation of delays, at this stage, D_i is known. Then such error signal at baseband is then:

$$e_{cB}(n_B) = e_{pB}(n_B) - w_i \cdot \Omega_x(n_B) \tag{15}$$

where

$$\Omega_x(n_B) = x_B(n_B - D_{iB}) \cdot e^{-j2\pi f_c D_i t_{sRF}}. \tag{16}$$

The cost function therefore becomes:

$$\begin{aligned} J =& E[e_{cB}(n_B) \cdot e_{cB}^H(n_B)] \\ =& E[e_{pB}(n_B) \cdot e_{pB}^H(n_B)] \\ & - E[e_{pB}(n_B) \cdot w_i \cdot \Omega_x^H(n_B)] \\ & - E[e_{pB}^H(n_B) \cdot w_i \cdot \Omega_x(n_B)] \\ & + E[w_i^2 \cdot \Omega_x(n_B) \cdot \Omega_x^H(n_B)] \end{aligned} \tag{17}$$

where $E[*]$ represents the expected value. Taking the gradient of the cost function:

$$\begin{aligned} g =& \nabla J(w_i) \\ =& \frac{\partial J(w_i)}{\partial w_i} \\ =& -E[e_{pB}(n_B)\cdot\Omega_x^H(n_B)] \\ & -E[e_{pB}^H(n_B)\cdot\Omega_x(n_B)] \\ & +2E[w_i\cdot\Omega_x(n_B)\cdot\Omega_x^H(n_B)]. \end{aligned} \tag{18}$$

The instantaneous estimation of gradient of the cost function is:

$$\begin{aligned} \hat{g} &= -v - v^H + 2q \\ &= -\Omega_x^H(n_B).e_{cB}(n_B) - \Omega_x(n_B).e_{cB}^H(n_B) \end{aligned} \tag{19}$$

where $v = e_{pB}(n_B)\cdot\Omega_x^H(n_B)$ and $q = w_i\cdot\Omega_x(n_B)\cdot\Omega_x^H(n_B)$. As the weight w_i is a real number, the real part of $\hat{g}$ is used. Therefore, (19) becomes:

$$\begin{aligned} \Psi &= Re(\hat{g}) \\ &= Re(-\Omega_x^H(n_B).e_{cB}(n_B) - \Omega_x(n_B).e_{cB}^H(n_B)) \\ &= -2Re(\Omega_x(n_B).e_{cB}^H(n_B)). \end{aligned} \tag{20}$$

The LMS weights adaptation equation for the RF Tx leakage canceller is:

$$w_i = w_i - \frac{1}{2}\mu\Psi \tag{21}$$

and the NLMS equation for weight adaptation is:

$$w_i = w_i - \frac{1}{2}\frac{\hat{\mu}}{\|\Omega_x(n_B)\|^2}\Psi \tag{22}$$

where $\hat{\mu}$ is the step size for NLMS.

Finally, for each baseband sample time n_B, calculate:

$$\Omega_x(n_B) = x_B(n_B - D_{iB})\cdot e^{-j2\pi f_c D_i t_{sRF}} \tag{23}$$

$$e_{cB}(n_B) = e_{pB}(n_B) - w_i\cdot\Omega_x(n_B) \tag{24}$$

$$\begin{aligned} w_i &= w_i - \frac{1}{2}\frac{\hat{\mu}}{\|\Omega_x(n_B)\|^2}\Psi \\ &= w_i + \frac{\hat{\mu}}{\|\Omega_x(n_B)\|^2}Re(\Omega_x(n_B)\cdot e_{cB}^H(n_B)) \end{aligned} \tag{25}$$

where Ω_x, e_{cB}, x_B and e_{pB} are all $1\times N_B$ vectors and N_B is the total number of samples taken at baseband.

When the system is used online, i.e., with received signal presented, Ω_x is changed into a matrix which consists of phase compensated baseband reference signals for all taps. The NLMS equations become:

$$\Omega_x(n_B)\mid_{i=1:M} = x_B(n_B - D_{iB})\cdot e^{-j2\pi f_c D_i t_{sRF}} \tag{26}$$

$$\Upsilon(n_B) = \begin{bmatrix} \Omega_x(n_B)\mid_{i=1} \\ \Omega_x(n_B)\mid_{i=2} \\ \vdots \\ \Omega_x(n_B)\mid_{i=M} \end{bmatrix} \tag{27}$$

$$e_{cB}(n_B) = e_{pB}(n_B) - w^T\cdot\Upsilon(n_B) \tag{28}$$

$$w = w + \frac{\hat{\mu}}{\|\Upsilon(n_B)\|^2}Re(\Upsilon(n_B).e_{cB}^H(n_B)) \tag{29}$$

where Υ is a $M\times N_B$ matrix and $\Upsilon(n_B)$ is the column vector for each sample time n_B. The weighting vector w is in the size of $M\times 1$.

IV Simulation results

Simulation results are obtained using Matlab. Sampling frequency of 4.092 GHz is used and the number of samples is equal to 131072. Baseband sampling frequency is chosen to be 512 MHz. The step size of NLMS is set to be 0.2. White Gaussian noise is used as the Tx signal and it is band-limited by the low pass filter at baseband. The duplexer model is based on the S-parameters of a WCDMA band VIII SAW duplexer [10] with a balun [11] for converting the balanced ports of Rx into a single ended un-balanced port. S-parameters are obtained by measuring the duplexer with a Vector Network Analyzer. The measured magnitude and phase response of the duplexer is shown in Fig. 2. For the same 20 MHz bandwidth, the magnitude and phase response at 925MHz are more complex and less linear, respectively than the ones at 890 and 905 MHz. There is a deep null at around 927 MHz and a sharp rise between 920 and 925 MHz in the magnitude response. These rapid changes in frequency domain are manifestation of a relatively complex impluse response of the system. Therefore, to achieve the same cancellation performance for the same bandwidth at 925 MHz, more taps are needed than at 890 and 905 MHz.

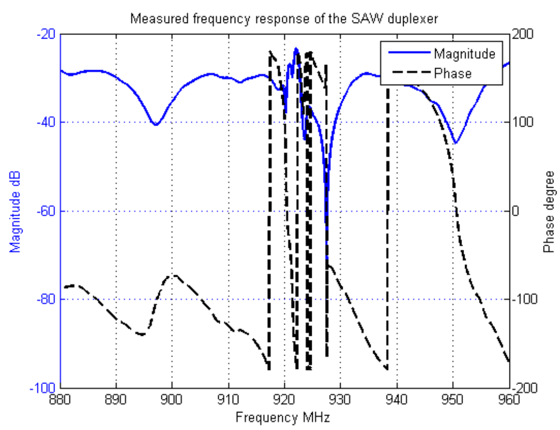

Fig. 2: Magnitude and phase response of the duplexer

The simulation results in Fig. 3-5 depict cancellation perfromance for three different numbers of taps (8, 16 and 64, respectively) for a 20 MHz bandwidth around centre frequency of 925 MHz. Although the larger number of taps is not feasible in practice, this is only to show that in the case where the frequency characteristics of the duplexer are relatively complex, a larger number of taps is needed to achieve the desired performance. To confirm this requirement, the comparison for the same bandwidth of 20 MHz for three different centre frequencies (890, 905 and 925 MHz, respectively) and increasing number of taps is shown in Fig. 6. Over 25 dB of cancellation has been achieved by using more than 8 taps for centre frequency of 890 and 905 MHz. The cancellation performance decreases dramatically for the centre frequency of 925 MHz to between 10 and 15 dB.

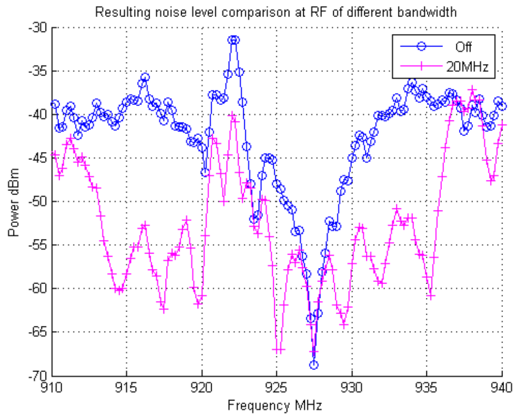

Fig. 3: Leakage level comparison with cancellation On/Off for 8 taps, centre frequency of 925MHz

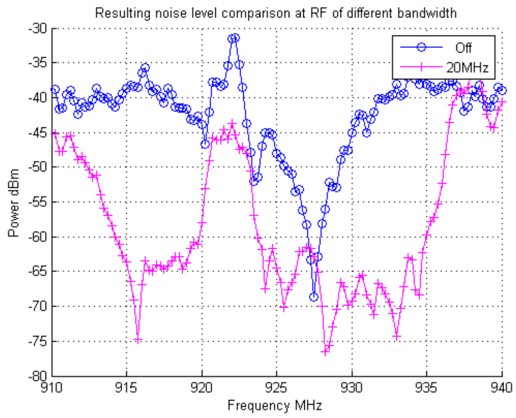

Fig. 4: Leakage level comparison with cancellation On/Off for 16 taps, centre frequency of 925MHz

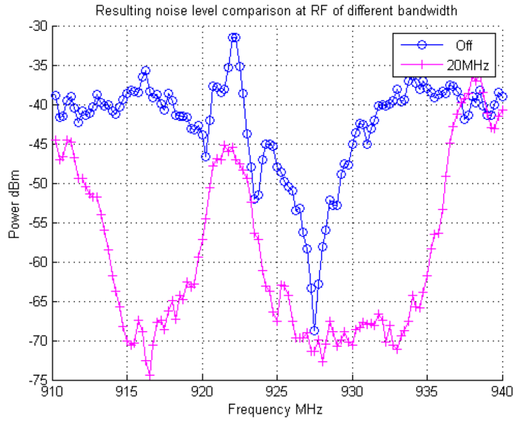

Fig. 5: Leakage level comparison with cancellation On/Off for 64 taps, centre frequency of 925MHz

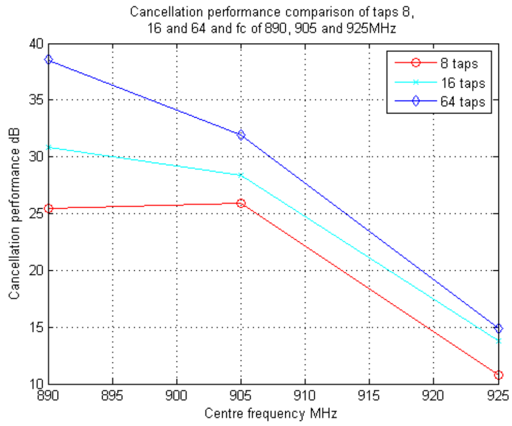

Fig. 6: Cancellation performace comparison for different numbers of taps and centre frequencies

V Conclusions

The RF Tx leakage cancellation structure is proposed in this paper, together with adaptation algorithm required for the canceller system. Simulation results show that for more complex characteristics of the duplexer, a larger number of taps is necessary. In the future, the tunable delay lines and variable attenuators will be realised by off-the-shelf components (where possible). In the presented paper, a resolution of the delay lines was approximately 240 ps, a value that is easily obtainable in practice. Both of the requirements are under investigation and they are outside the scope of this paper. Additional work has to be carried out to improve the isolation of the SAW duplexer even further, for the same bandwidth and higher frequency of operation. However, up to date, the adaptive delay filter approach proves to be the most promising method of cancellation for wide in-band interference, where filtering is not an option.

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References

[1] S. Ahmed and M. Faulkner, “Optimized Interference Canceling for Colocated Base Station Transceivers,” Vehicular Technology, IEEE Transactions on, vol. 60, no. 9, pp. 4175-4183, 2011.

[2] A. Frotzscher and G. Fettweis, “Least Squares Estimation for the Digital Compensation of Tx Leakage in zero-IF Receivers,” in Global Telecommunications Conference, 2009. GLOBECOM 2009. IEEE, 2009.

[3] V. Aparin, G. Ballantyne, C. Persico and A. Cicalini, “An integrated LMS adaptive filter of TX leakage for CDMA receiver front ends,” Solid-State Circuits, IEEE Journal of, vol. 41, no. 5, pp. 1171-1182, 2006.

[4] S. S. Hong, J. Mehlman and S. Katti, “Picasso: Flexible RF and Spectrum Slicing,” in Proceedings of the ACM SIGCOMM 2012 Conference on Applications, Technologies, Architectures, and Protocols for Computer Communication, New York, NY, USA, 2012.

[5] S. Kannangara and M. Faulkner, “Analysis of an Adaptive Wideband Duplexer With Double-Loop Cancellation,” Vehicular Technology, IEEE Transactions on, vol. 56, no. 4, pp. 1971-1982, july 2007.

[6] H. Kim, S. Woo, S. Jung and K.-H. Lee, “A CMOS Transmitter Leakage Canceller for WCDMA Applications,” Microwave Theory and Techniques, IEEE Transactions on, vol. 61, no. 9, pp. 3373-3380, 2013.

[7] N. Phungamngern, P. Uthansakul and M. Uthansakul, “Digital and RF interference cancellation for single-channel full-duplex transceiver using a single antenna,” in Electrical Engineering/Electronics, Computer, Telecommunications and Information Technology (ECTI-CON), 2013 10th International Conference on, 2013.

[8] D. Bharadia, E. McMilin and S. Katti, “Full Duplex Radios,” in Proceedings of the ACM SIGCOMM 2013 Conference on SIGCOMM, New York, NY, USA, 2013.

[9] D. Etter, “Identification of sparse impulse response systems using an adaptive delay filter,” in Acoustics, Speech, and Signal Processing, IEEE International Conference on ICASSP ’85., 1985.

[10] EPCOS, “SAW duplexer WCDMA band VIII,” datasheet, Jul. 2013.

[11] J. Technology, “900 MHz Balun P/N,” 0900BL18B100 datasheet, Aug. 2004.

[12] S. Haykin, “Normalized Least-Mean-Square Adaptive Filters,” in Adaptive Filter Theory, 4th ed., New Jersey: Prentice Hall, 2001.

[13] Jean Jiang; Etter, D.M., “Complex system modeling using a complex delay filter,” Signals, Systems and Computers, 1992. 1992 Conference Record of The Twenty-Sixth Asilomar Conference on , vol., no., pp.6,10 vol.1, 26-28 Oct 1992