Achievable Performance and Limiting Factors of Echo Cancellation in Wireless Communications

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***Abstract*— Self-interference cancellation for wireless communications, similar to its wireline counterpart echo cancellation, is considered in this paper. We analyze the factors that impacts echo/self-interference canceller performance. It is shown that the main factors limiting its performance are the excess MSE, echo-channel estimation error, and various non-linear effects. By establishing a linearized model of the channel estimator, we derive expressions of the errors at the echo canceller output as functions of the estimator's time constant, the bandwidth of channel variation, the receiver SNR and the echo level. The optimal adaptation time constant are derived. Non-linearities in the transmitter and receiver are shown to be the main limiting factor. Finally, implementations considerations of echo cancellers are discussed.**

I. Introduction

One of the main objectives for digital communications is to improve the system spectrum efficiency. In past a few years, the concept of transmitting and receiving data in the same band at the same time has attracted attention to the engineers and researchers in the wireless communication area. It has been called "single-channel full-duplex wireless communication", "In-band full-duplex (IBFD)" and etc.

The main idea of this technology is that the wireless communication units need to reduce, cancel and eliminate as much as possible its own transmitted (Tx) signal leaked to its receiver input, called *self-interference*, in order to achieve reliable communication. This idea was stimulated by the work in the Radar area [1][2]

While the idea of cancelling interference from the Tx signal of the same unit is relatively new in the wireless communication field, it is one of the key technologies that has been known, implemented and successfully commercialized in wireline communication area for over 35 years. The technology, called *echo cancellation*[1], was first proposed and published in mid 60's [3][4][5][6]. It was the main component of various full-duplex wireline modems. For example, the first echo cancellation modem standard CCITT-V.32 was established in early 1980's. The first of such modems was commercialized around 1985. Echo cancellation is also one of the main building blocks of ISDN, xDSL, 1G and 10G Ethernet modems. A lot of work had been done on the analysis and implementation of this technology, which should be useful for the extension to the wireless full-duplex communications.

To extend the work on full-duplex wireline echo cancellation modem to full-duplex wireless communication systems, researchers and engineers face new challenges. Mainly, there are three factors make the task more difficult. First of all, in most wireless systems, the link attenuation is quite severe. While wireline-link attenuation is usually less than 30 to 40 dB, wireless link could easily have more than 100 dB attenuation, e.g., in a cellular system. Secondly, the channels over wireline media could be viewed as static. It changes most likely due to temperature change and very slow. The wireless channels always experience time variance, or fading. Even in relatively stationary environment, it can also change due to the movement of the objects around it or even air movement. It has been observed that when inside a room, the fading frequency is similar to a vehicle will experience when moving 1 to few kilometers per hour. Moreover, such signal variation can also be due to phase noise, which is usually worse in wireless systems than for wireline modems. The final, and probably the most important, part of impairments that impact echo/self-interference cancellation is non-linear characteristics of the transmitter. Echo/self-interference cancellation can be perfect if the system is totally linear. However, any non-linearity presented in the systems cannot be cancelled by the digital base-band echo-cancellers used in both wireline and wireless full-duplex systems. RF amplifier, especially the higher power amplifiers usually presents higher non-linearities. Such non-linearity could be easily at about -30 to -40 dB and very difficult and costly to reduce.

On the other hand, wireless systems also provide some beneficial factors to achieve better cancellation/suppression of transmitted signals which interferes received signals.

Since wireless communications utilize radio waves as communication media, the spatial dimensions that can be used by system designers to reduce the interference from the transmitter signal to the receiver. For example, if separate Tx and Rx antennas are used, beam forming can steer the transmission wave away from the Rx antenna in the same unit to achieve better isolation. Thus, the burden of eliminating echo by the baseband echo canceller can be reduced.

As concerning channel fading, its performance impact to the echo canceller is a function of the ratio of the data symbol rate to the fading frequency. Considering the modern wireless

[1] Echo and self-interference shall be used for wireline and wireless communications respectively, even though they are essentially the same impairments in these communication systems.

communication systems usually has wider bandwidth then wireline systems, especially comparing to wireline modems, such impact may be not as severe as at the first appearance. However, as shall be shown below, it cannot be ignored when a high cancellation requirement is needed.

Investigations of the impacts of these impairments are the main topics of this paper and shall be presented in the subsequent sections. Since a lot of investigations have been made in the echo cancellation area, we shall describe what we learned there. We feel these lessons and conclusions should have reference values for people working the wireless areas.

This paper is organized as follows. In section II, we take a look of the main components of the two-wire wireline full-duplex echo cancellation modems and the wireless in-band full duplex systems. Our emphasis will be on the analysis of the adaptive base-band echo/self-interference cancellers used in both of such wireline and wireless systems in Section III. In particular, its performance under fading environment will be analyzed. The practical implementation considerations of the digital base-band cancellers that we learn from wireline echo canceller are discussed in Section IV. In Section V we summarize the factors that impact the self-interference cancellers and discuss its achievable performance under practical constraints. Section VI presents our conclusions and discusses possible directions for future investigation.

II. Architecture and Components of Wireline and Wireline Modems with Echo/Self-Interference Cancellation

Even though there are differences in the wireline and wireless communication systems utilizing echo/self-interference cancellation, their basic operations, especially for the base-band digital cancellers, have many features in common. Thus, we first take a detailed look at their operations and the constituting components.

A. *Basics of wireline modem echo canceller*

Figure 1 is a block diagram of a wireline echo cancellation modem.

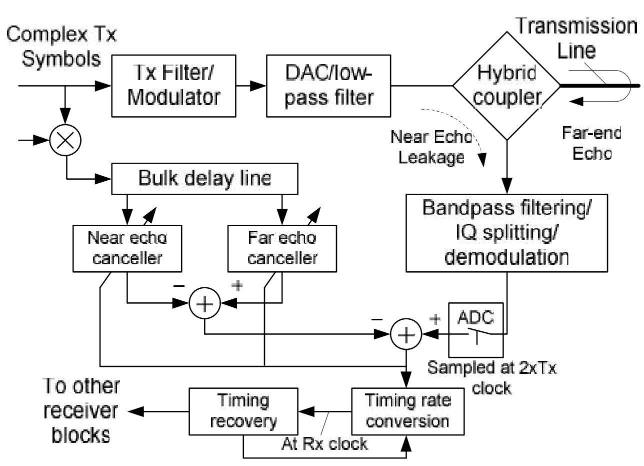

Figure 1 A typical wireline modem with echo canceller

The baseband complex data symbols are modulated at a frequency $f_c > 1/2T$, where T is the symbol rate, i.e., the DC component shifts to f_c. Frequencies that are negative or beyond $2xf_c$ are removed by digital filtering. The real part of the signal is converted to analog form by digital to analog converter (DAC). Any energy in the analog signal beyond $2xf_c$ are removed by analog filters. After proper amplification, the signal is sent to the Tx port of a hybrid coupler. It arrives to the transmission line to be sent to the remote modem. The modem receiver receives the signal from remote modem at the receiver port of the hybrid coupler.

The hybrid coupler's function is to isolate the transmitter port and the receiver port. However, since the isolation is not perfect, there are always residual Tx signals leaking back and corrupt the received remote signal. It is called the near echo.

For modem data transmission, the transmission line usually has a four wire portion for half duplex transmission. At the remote side it is converted to a two wire section, called local loop, to connect to the remove modem. The conversion is done by another hybrid coupler. At this point, part of the transmission signals is reflected back to form the so-called far-end echo. Thus, the received signal is corrupted by both near and far echo but nothing between. The modem echo canceller needs to remove both near and far echoes.

The received signals are converted to baseband complex signals after filtering, demodulation and IQ-splitting. To remove the echoes, the echo cancellers, which are usually implemented as FIR transversal adaptive filters, estimate the echo channel responses and synthesize the echoes using the known Tx symbols and the emulated echo channel. The echoes in the Rx signal can be eliminated by simply subtracting the synthesized echoes from the signal, as long as the echo channel is estimated and emulated accurately.

Let us assume γ dB of the signal to noise ratio (SNR) is required to ensure normal receiver operations. In order for the residual echo not to affect noticeably the receiver performance, we would like the residual echo to be at least 6dB below the allowable noise level. Using a wireline modem as an example, the Tx signal level is typically –9 dBm, and the range of the Rx signal level is typically from -43 dBm to -9 dBm. In the worst case, a residual echo level of $-43-\gamma-6$ dBm or lower is required. On the other hand, the minimum insertion loss of a hybrid coupler, i.e., the attenuation from the transmitter port to the receiver port, is equal to 6 dB, i.e., the highest echo level at the hybrid receiver port is -15 dBm. Hence, the echo canceller is required to provide a $43+\gamma+6-15$ dB echo rejection. The ratio of the echoes before and after echo cancellation is called the echo return loss enhancement (ERLE) of the echo canceller. For γ=24dB, we would like the echo canceller to provide an ERLE of 58 dB, i.e., the echo power need to be reduced by a factor of 640000. Thus, an echo canceller must estimate the echo channel very accurately. Other type of wireline echo canceller could impose even higher ERLE requirement but usually not much more than 70 dB

Assuming that the echoes are indeed a perfectly linear transformed transmitted data symbols, it is possible, at least in theory, to achieve an infinite ERLE. Practically, there are other factors that limit the achievable ERLE. First, there are always some non-linearities in the echo channel. Obviously, any nonlinear echo cannot be cancelled by a linear canceller. Secondly, when implemented using digital signal processing, the accuracy in channel modeling and echo canceller implementation are limited by finite precision effects. Thirdly, if there are channel variations, the accuracy of channel estimation are theoretically limited. We will discuss these factors in more detail below.

Another interesting point of modem echo canceller is that there exist a near echo and a far echo, the total echo channel response may become very long, e.g., up to 600ms or longer. As the near and far echoes are separated by a long silence, an echo canceller can be implemented as two shorter parts separated by a long time delay and its complexity is greatly reduced.

B. Overview of self-interference cancellers for wireless communications

In-band full-duplex wireless communications have attracted attentions in the recent years [7-11]. In principle it is very similar to wireline full-duplex communications. Namely, the basic idea is to use the known Tx symbols to cancel the self-interference caused by the same Tx signal. Many aspects of the wireline echo cancellation modems are applicable to its wireless counterpart. Below, we shall mainly describe their similarities and, especially, differences.

Figure 2 is a conceptual diagram of such an in-band full-duplex wireless communication unit.

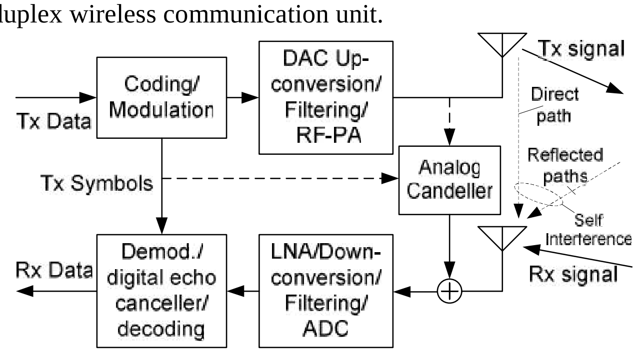

Figure 2 A typical wireless in-band full-duplex communication system with self-interference cancellation

As shown in the figure, the transmitter path is essentially the same as the echo-cancellation modem. Since the wireless unit is operated at RF frequency, the signal is amplified by an RF power amplifier. The resulted RF signal is radiated by Tx antennas to be received by remote units.

The transmitted RF signal will inevitably to some degree leaked back to the receiver antenna immediately. Moreover, such Tx radio signal will also be reflected by surrounding objects back to the Rx antenna. The Tx signals leaked/reflected back to the Rx antenna interfere the received signal if both transmission and reception utilize the same frequency band and at the same time. Effectively suppressing and removing such interferences, called self-interference, are the main tasks in order for the In-band full-duplex wireless communication units to operate reliably.

Without special design considerations, such self-interference can be 10s or even over 100 dB higher than the received signal. To reduce such self-interference, a few special design arrangements can be taken. As the first stage, isolation should be as high as possible between the Tx and Rx antennas, achieved by proper Tx and Rx antenna design and arrangement. For example, by using beam forming and proper antenna positions, it may be possible to have as little overlap between their patterns as possible. If it is desirable to use a single antenna instead of two, RF *circulators*, which is very similar to the hybrid coupler in wireline modems, can be used. However, the sizes of such circulators are usually too large for handheld devices. Moreover, the isolation between its Tx and Rx ports is typically around 15dB. To achieve higher isolation, the circulator size would be even larger.

Secondly, part of the self-interference at the Rx antenna output can be removed by an analog canceller. By incorporating analog cancellers, the interference level is reduced to facilitate further digital processing. In particular, the selection of analog to digital converters (ADCs) will be much easier as shorter word-lengths can be used to operate due to the reduction of the dynamic range of the self-interference in the Rx signal after analog cancellation.

The input to the analog canceller can either be generated from the Tx digital modulation symbols or directly taken from the Tx RF signal. If the Tx RF signal is used as the reference input, the analog canceller's performance is less impacted by the non-linearities in the Tx RF signal. However, since the processing will be in analog domain, the implementation will be less straight forward than the first case. The RF reference signal can either be tapped from the Tx antenna port or from a special antenna that is coupled to receive the Tx signal.

After part of the self-interference are removed, the Rx signal with the remaining self-inference are down converted in frequency, filtered and converted to digital samples by ADCs. To further reduce the self-interference, the receiver employs a digital self-interference canceller very similar to the echo canceller in wireline modems. It estimates the channel response of the remaining self-interference in the digital samples and synthesizes the interference using the Tx symbols and the emulated channel. The remaining interference in the Rx digital samples can be eliminated by simply subtracting its synthesized form from the samples, as long as its channel response is estimated and emulated accurately.

Since the digital canceller is linear, theoretically its performance is limited by the channel estimation accuracy and the non-linearities in the Tx signal and the Tx-Rx echo path. For wireless systems with high power RF amplifiers, it is very difficult and costly to reduce the amplifiers non-linearity. Such non-linearity may be estimated and incorporated in the digital cancellation unit. It will be interesting and important to investigate such non-linear compensation process.

Assuming the channel is stationary, at least in theory, its estimation accuracy can be infinite as long as it is linear. However, in wireless communication systems, the channel will never be perfectly stationary due to signal fading. In next section, the linear echo/interference cancellers will be discussed in detail. We will also investigate the theoretical achievable channel estimation accuracy under time-varying environments and its impact to the canceller performance.

III. Digital Nyquist echo/interference canceller

The digital echo/self-interference canceller is the last cancellation stage. It should be very accurate to ensure the best achievable performance. To achieve such a goal, it has to operate at Nyquist rate, i.e., there's no loss after analog to digital conversion. Thus, it is called Nyquist canceller and essentially the same for both wireline and wireless applications. In this section, we describe and investigate the basics of such cancellers in detail.

The Nyquist canceller is usually implemented as a transversal adaptive filter. To simplify implementation, LMS algorithm is usually used for adaptation

A. Description

The reference input to the canceller is Tx symbols at a rate of 1/T, where T is the Tx symbol rate, or baud rate. The Tx symbols are stored in tapped-delay-line. The Rx signal with interferences to be removed is sampled at a rate higher than 1/ T as to satisfy the Nyquist criterion. The Rx sampling rate is synchronized with Tx symbol rate. The most common sampling rate is 2/T, although a fraction number greater than one, such as 3/2T or 4/3T, can also be used. A block diagram of a 2/T canceller with span of LT is shown in Figure 3.

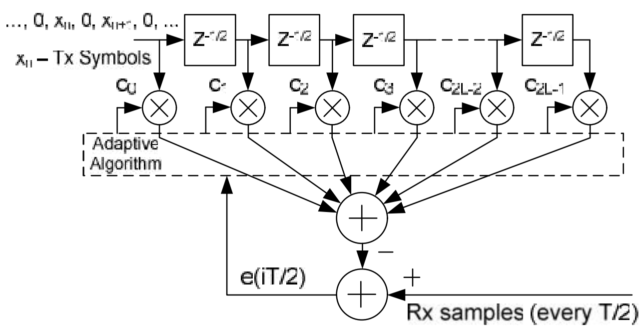

Figure 3 Block diagram of a Nyquist Echo/self-interference canceller

Let's assume the Rx sampling rate is twice of the Tx symbol rate. The canceller is operated at the rate of 2/T. One zero is inserted between two Tx symbols. The synthesized interferences are subtracted from Rx samples. The errors such generated are used for adaptively adjust the canceller coefficients. Alternatively, such a canceller can be viewed as two sub cancellers, each of which has L taps, operating independently. The synthesized interferences from the first filter are subtracted from the even Rx samples and the second outputs from the odd ones. It can be shown such a structure is not only convenient but also optimal (see e.g. [12]).

B. Basics of LMS adaptive canceller

The basic LMS adaptive algorithm is well known in the literature [13]. We will review the basic results as follows.

The i-th subcanceller's output, i.e., the estimated interference in the i-th sample stream, can be expressed as:

$$\hat{I}_i(n) = \sum_{k=0}^{L-1} x_{n-k} c_{k,i}(n) \tag{1}$$

The error between the estimate and the true interference is computed by:

$$e_i(n) = r_i(n) - \hat{I}_i(n) \tag{2}$$

The coefficients are updated at next iteration as:

$$c_{k,i}(n+1) = c_{k,i}(n) + \Delta x_{n-k} e_i^*(n) \tag{3}$$

where Δ is the adaptation step size and * represents complex conjugate operation.

The coefficients such computed will converge towards its optimal value when n goes to infinity. However, the noise in $r_i(n)$ will introduce errors in the coefficients. This results in an error term proportional to the irreducible error in $e_i(n)$, called *excess error*., denoted by e_{ex}. It can be shown that the mean square value of e_{ex} can be expressed as $£_{ex} = (« / 2)L£$, where $« = \Delta / LE[|x_n|^2]$ is the normalized step size and £ is the MSE of the irreducible error, i.e., the residual error even if all of the echo/self-interferences are removed.

Note that $r_i(n)$ contains both echo/self-interference and the received signal, which can be viewed as "noise" regarding to the echo/self-interference to be canceled. As a consequence, the canceller is operated at very low "SNR" and £ is equal to the total Rx signal energy. To achieve an Rx SNR of γ, the excess MSE must be ¥ plus a few, say 6, dB below £. In other words, $« \le 2 \times 10^{-(¥+6)/10} \big/ L$. For example, if ¥ = 27 dB and L = 100, we will need « to be less than 10^{-5}, or if we let if ¥ = 21 dB and L = 40, we will need « to be less than 10^{-4}.

It should be noted that the number, *L*, of the canceller coefficients should at least equal to the number of non-trivial taps of the echo/self-interference channel, which consists of the propagation/reflect channel combined with the Tx and Rx filters. Thus, its length is at least 20-30T or longer.

The canceller must be implemented properly to accommodate such small step-sizes. The time constant due to small step-sizes will also affect the ability of tracking changes in the interference channel characteristics. Below we investigate the tracking characteristics.

C. LMS adaptive canceller tracking characteristics

The input data to the echo/self-interference cancellers, i.e., the Tx symbols, are uncorrelated with each other. Thus, the input's autocorrelation matrix is an identity matrix. This is advantages for the tracking ability of the canceller. As have been shown previously, the LMS algorithm exhibits an exponential convergence property. We can view such an adaptive filter as a linear system with an impulse response of $h(n)=U(n)«(1-«)^n$, which has a time constant of $(1–«)^{-1}$T. To evaluate the performance of such an adaptive canceller in a fading channel environment, we first analyze its frequency domain characteristics. By taking DFT of $h(n)$, we have

$$H(e^{j\mu}) = \sum_{n=0}^{\infty} «e^{-j\mu n}(1-«)^n = \frac{«}{1-e^{-j\mu}(1-«)} \tag{4}$$

For an ideal estimator the frequency response should be a constant 1. The difference between our LMS filter's frequency response and the ideal response is

$$1-H(e^{j\mu})=1-\frac{«}{1-e^{-j\mu}(1-«)}=\frac{(1-e^{-j\mu})(1-«)}{1-e^{-j\mu}(1-«)} \quad (5)$$

This equation characterizes the tracking error of the LMS adaptive filter relative to the ideal channel estimate. It should be noted that as we have shown previously, the exponential windowed LS algorithm has the same characteristics in tracking mode [14][15]. Thus, this result is also applicable to such LS adaptive filters. Before we discuss the general cases, let's first examine the adaptive filters' behavior for some simple yet fundamental cases.

Let us first consider the stationary case. Since the channel impulse response does not change with time, its frequency domain characteristic is a unit impulse at DC. From Eq. 4, we have $H(e^{j\mu})|_{\mu=0}=1$. Thus, this estimator is ideal for stationary channel estimation. It is interested to point out that this conclusion is independent of the selection of step-size «. However, the smaller the step-size is, the narrower is the bandwidth of the adaptive system. Thus, it can better reject additive noise to have less channel estimation error caused by the noise. On the other hand, as we shall see below, it yields larger tracking error in fading environment.

Secondly, let us consider the case of a channel varying as a single complex sinusoidal $e^{j\mu_0}$. After some simplification, Eq. (5) can be written in the following form:

$$1-H(e^{j\mu_0})=\frac{2(1-«)(1-\cos\mu_0)}{(2-«)(1-\cos\mu_0)-j«\sin\mu_0} \quad (6)$$

Then, the power of the estimation error relative to the channel coefficient energy due to the sinusoidal can be computed as

$$\left|1-H(e^{j\mu_0})\right|^2=\frac{4(1-«)^2(1-\cos\mu_0)^2}{(2-«)^2(1-\cos\mu_0)^2+«^2\sin\mu_0^{\,2}} \quad (7)$$

By using $\cos\omega_0 \approx 1-(\omega_0)^2/2$ and $\sin\omega_0 \approx \omega_0$, we have,

$$\begin{aligned}\left|1-H(e^{j\mu_0})\right|^2&=\frac{4(1-«)^2(1-\cos\mu_0)^2}{(2-«)^2(1-\cos\mu_0)^2+«^2\sin\mu_0^{\,2}}\\ &\approx\frac{4(1-«)^2\mu_0^4/4}{(2-«)^2\mu_0^4/4+«^2\mu_0^2}=\frac{(1-«)^2\mu_0^2}{(2-«)^2\mu_0^2/4+«^2}\end{aligned} \quad (8)$$

In (8), ω_0 can be expressed as $2\pi f_D T$, where f_D is the fading frequency and T is the Tx symbol interval. In Table 1, we show the ratio of the total self-interference power to the tracking error with respect to $f_D T$. The results, i.e., the limits of achievable cancellation for a given single tone fading for $\mu = 10^{-4}$ and 10^{-5} are given in Table 1. It is interesting to note that for small μ and very low f_D relative to the Tx symbol rate, as usually the case of echo/self-interference cancellation, the residual interference after cancellation is proportional to μ^2 and inversely proportional to the square of f_D.

Limit of Cancellation	f_DT (ratio of fading frequency to Tx symbol rate)			
	10^{-7}	10^{-8}	10^{-9}	10^{-10}
$\mu = 10^{-4}$	44 dB	64 dB	84 dB	104 dB
$\mu = 10^{-5}$	24 dB	44 dB	64 dB	84 dB

Table 1. Limit of achievable cancellation with fading

The above results show that the limit of achievable cancellation increases by 20 dB for every 10 times reduction of the fading frequency. Thus, channel fading impose a non-neglectable limit on achievable cancellation. Assume the Tx symbol rate is 5 meg symbols per second, even at a fading frequency of 0.05 Hz ($f_D T=10^{-8}$), the cancellation is limited by 64 dB for $\mu = 10^{-4}$ and 44 dB for $\mu = 10^{-5}$.

Practically, channel fading is usually not a single tone but a random process. If the power spectrum of the fading process is known, we can compute the limit of the residual interference due to channel fading by integrating (7) or (8) over its power spectrum.

Comparing to channel estimation in receivers [15][16], the requirements for echo/interference cancellers' channel estimation are usually tougher. In receivers, the error introduced due to channel estimation is proportional to the signal itself. As a result, the channel estimation accuracy may not need to be much more than required SNR. Moreover, these channel estimation errors could be compensated in part by demodulator or equalizers. For echo/self-interference cancellers, the interference level could be much higher than the received signal. Thus, the estimation accuracy also needs to handle this additional gap. Moreover, any channel estimation error will be translated to residual interference and there's no clear solution how to further reduce it.

IV. IMPLEMENTATION CONSIDERATIONS

In this section we consider three areas specific to implementation of echo/self-interference cancellers.

A. ADC word-length requirements

If we want to achieve a cancellation ratio of Γ dB, the quantization error should be Γ+6 dB below the echo RMS value. Furthermore we can assume the peak to RMS ratio of the interference signal is 12 dB (2bits). The required word-length of ADC would be at least Γ/6+3 bits. For example, if the cancellation requirement is 60 dB, the ADC used should have a word-length of 13 bits. Considering the system may operate at multi-meg samples per second the cost of a high precision ADC would not be insignificant.

B. Finite word-length effect of echo/interference estimation

Adaptive echo/self-interference cancellation is realized in two steps: computing the synthesized echo/interference and updating the canceller coefficients. The estimated echo/interference is generated by convolving Tx data symbols with canceller coefficients. The convolution output should have comparable precision as the signal at the receiving side generated by ADC. Thus, the required word-lengths of the canceller coefficients and Tx symbols should be comparable to the ADC word-length. However, to reduce the overall loss, they may be chosen to have somewhat higher precision than ADC because the cost of using longer coefficient and Tx symbol word-lengths is less than longer ADC word-length.

Canceller coefficients updating step deserves more careful examination. To achieve satisfactory residual error in the estimated echo/interference, i.e., the excess error, it is necessary to use a small step size. Especially for the self-interference cancellation in wireless applications, there is usually not training sequence. Thus, the canceller training and updating are operated when the received signal exists. As mentioned above, the received signal power is usually much higher than the excess mean square error of the adaptive canceller. Hence, a very small step size should be used. The correction term, $\Delta x_{n-k} e_i^*(n)$, is the product of the Tx symbol, the step-size and the error term. In order for the adaptation to continue when the error has converged, the correction term at least should not underflow. Since the step size Δ is small when high SNR is needed, the correction terms become much smaller than the canceller coefficients. In order to track the channel change, no matter how slow it is, the word-length of canceller coefficients should be quite long even though the required MSBs for convolution may not be that many. For example, in most cases, 16 or fewer bits are sufficient for convolution. However, the coefficients need to be probably 24-32 bits during adaptation, although only less than 16 MSBs are used for echo/interference estimate. It should be noted that only a long accumulator is needed when performing the adaptation. The multiplier can be much smaller.

Another popular method for echo canceller coefficient updating is the *gradient averaging algorithm*. Namely, the correction terms are computed and accumulated with a large scaling factor, as long as no overflow occurs. The accumulated value is scaled down to update the canceller coefficients when appropriate and then the accumulation process starts again.

These two methods, i.e., using "double precision" coefficients for updating and the gradient averaging are essentially equivalent in performance. Their complexities are also comparable.

Above, we only discussed qualitative considerations of echo/interference canceller design and implementation. More details and design examples can be found in [12].

C. Sampling rate conversion

To remove the echo/self-interference from the received signal, a canceller uses Tx symbols as the reference input. Thus, its operations and the received signal sampling must be synchronous to the Tx clock. After cancellation, to recover the remotely transmitted data from the interference-removed Rx signal, the receiver should operate synchronously to the remote transmitter clock, called Rx clock below. A sampling rate converter is used to perform the Tx to Rx clock conversion of the interference canceled samples.

In early days of echo cancellation modems, the conversion is done by using analog means. Since mid-80s, digital rate converters are used to perform such conversion to reduce the cost of product among other advantages [17]. A modem block diagram with digital echo/interference cancellation and digital resampling conversion is shown in Figure 4.

Digital rate conversion, or resampling, is essentially done by digital interpolation technique [18]. Both linear and non-linear interpolation methods can be used for this purpose. For the digital resampling, the conversion rate should be an arbitrary number. To achieve this, we can either use a poly-phase filter bank with many subfilters. A sub-filter with the output timing phase, or delay, that is most close to the desirable one is used to generate the output. We call such a method the *zero-th order interpolation*. Using this method, only one output needs to be computed for each sample. However, to ensure sampling phase accuracy, the number of subfilters would be quite large.

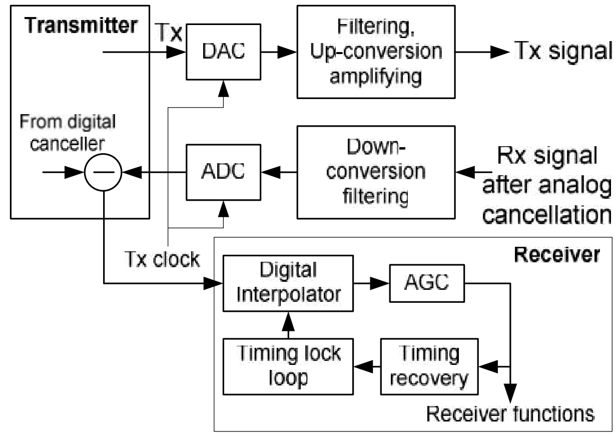

Figure 4 Modem with digital resampling

To reduce the number of subfilters, another approach is to compute two outputs that have the timing phases closest to the desire sampling phase. Then a simple linear interpolation is used to compute the final output sample. This approach can greatly reduce the required number of subfilters but it will more than double the computation. This method is called *first order interpolation*.

It can be shown the ratios of the signal to the error introduced by the zero-th and first order interpolators are bounded by $¥_{sd}^{(0)} = P_s / P_e^{(0)} \geq 12U^2 / \mu_y^2$ for the zero-th order interpolator and $¥_{sd}^{(1)} = P_s / P_e^{(1)} \leq 320U^4 / \mu_y^4 \geq$ for the first order interpolator, where U is the number of the sub-filters needed and μ_y is the Rx signal bandwidth of the input samples. More details of digital interpolation and timing recovery can be found in [19] and the online slide-deck [20].

Since the digital interpolations are operated on the digital samples for receiver functions after cancellation, its accuracy is determined by the required signal to noise ratio and independent of the echo cancellation requirement.

V. PERFORMANCE LIMITING FACTORS OF ECHO/SELF-INTERFERENCE CANCELLERS

If the transmitter and the overall channel are linear and static, the cancellation can be perfect with ideal implementation. However, practically, there are always impairments in the received interferences and variations in propagation channels. Thus there are limits to the achievable performances of cancellation even with optimal implementation. Below we list and discuss these limiting factors and their impacts to the achievable performance.

A. Nonlinear distortion of Tx signal

There are mainly two types of distortions in the Tx signal: memoryless nonlinearity and multiplicative interference due to phase noise.

For wireless communications, signals are amplified by RF power amplifiers before radiated by antenna. For higher power RF transmission, it is difficult and costly to employ highly linear RF power amplifiers. Normally, the objective of the design is to make sure the non-linear effects will not impact the receiver performance, i.e., the distortion due to non-linearity should not noticeably affect the received signal SNR. As a consequence, the non-linear requirements of the amplifier are usually no more than 30 to 40 dB.

However, since a linear canceller cannot cancel the distortion introduced by non-linearity, the non-linearity constitutes a limit on its performance. Moreover, the echo/self-interference could be much higher than the received signal level. Thus, there is additional requirement margin equal to the worst case interference to received signal ratio. Such a demand would be very difficult to meet.

One way to rectify this problem is use an analog canceller with the distorted RF signal as its reference as we discussed above. By using such analog cancellers, the interference entering the digital canceller may be reduced to more manageable level. However, the design and implementation of the analog canceller are more difficult and could be challenging.

Another method to reduce the non-linear effect is to pre-distort the reference input to the digital canceller based on the estimated non-linearity in the received interference. In theory this approach could work well. However, further investigation will be needed to find out how accurate such estimation can practically achieve.

B. *Noise floor in Tx signal due to phase noise*

It is well known that the imperfections of phase locked loops (PLLs) used for frequency synthesis for the transmitters introduce phase noise that would corrupt Tx signal. The low frequency components of such phase noise behave as time variation of the signal similar to channel fading. Such time variation will introduce adaptive cancellers' tracking error and limit its performance as show above and further discussed below. The high frequency components of phase noise will behave as multiplicative noises to constitute a noise floor in the Tx signal. Such noises in the Tx signal cannot be cancelled by linear cancellers. However, the analog canceller discussed above may cancel part of low frequency components but its effectiveness is also need to be seen.

C. *Tracking error of the interference canceller*

In Section III.C, we analyzed the excess adaptation error and the tracking error introduced by the LMS adaptive canceller. The main conclusions are as follows. On one hand, to achieve low excess error, the step-size of the LMS canceller needs to be chosen rather small and its time constant could be quite long. On the other hand, as shown above, such long time constant makes the canceller's tracking error sensitive to even very low frequency channel variations. Our analysis shows that the tracking error reduces by 20 dB when the variation frequency reduces by a factor of 10. This reduction rate is not very steep. Considering there are always some channel variation exist, this may become a key limiting factor in the cases when high cancellation ratio is needed. It should be pointed out that this conclusion affects the both analog and digital cancellers.

D. *Tx and Rx signal isolation*

In the case that the channel has large path loss, such as in cellular systems, adaptive cancellers may not be able to provide enough cancellation to ensure satisfactory communications. In a wireless system with separate Tx and Rx antennas, employing better designed antennas can achieve additional signal isolation to reduce the adaptive canceller's burden. Such an approach is more suitable for point to point system with symmetric links. For asymmetric, such as point to multipoint communications system, e.g., cellular systems, it would be easier to implement on the base-stations side. For hand-held devices, there may not be enough space to implement a highly isolated antenna system between the Tx and Rx sides. Moreover, consider the usage environment of hand-held devices, the radio signal reflections would be more complex and may more likely to have path variation. As a result, it would be difficult to achieve high Tx to Rx signal isolation.

To summarize, we may summarize the factors discussed in this section in the following empirical formula of the total achievable total self-interference rejection, $R_{c,total}$:

$$R_{c,total} \le \min[(R_{NL} + R_{NL-compensation}), R_{ch-\mathrm{var}}] + R_{Isolation} \quad (9)$$

where R_{NL} is the achievable rejection under non-linear effects in the interference including the phase noise and signal non-linearity, $R_{NL-comp}$ is the additional gain achieved by non-linear compensation, R_{ch-var} is the rejection limit imposed by channel variations and $R_{Isolation}$ is the self-interference reduction by the Tx/Rx isolation. How much total rejection can achieve depends on many different factors. Based on the past experiences for echo cancellation and analysis give above, the first term is likely to be in the range of 30 to 60 dB, depending on how bad the non-linearity/multiplicative interference are. Any additional required rejection would need to be provided from the isolation between the Tx and Rx ports. Of course, these numbers given are just educated guesses. Only practical design and implementation can show how much rejection is indeed achievable in practically designed systems.

VI. Summary and Conclusions

In this paper, we looked into the self-interference cancellation technology for wireless communications based on our understanding of echo-cancelation for wireline modems. Their similarity and differences are discussed. Fundamentally they both reject and remove their own Tx signal components that corrupt the received signals to achieve full duplex communication over the same media at the same time. In order to do so, such Tx signals should be isolated from the receiver inputs as much as possible. The leaked and reflected interferences are removed by adaptive cancellers utilized in both cases.

However, there are also differences between them. If employing separate Tx and Rx antennas is possible, it is possible to create better isolation between the Tx and Rx ports of a wireless full-duplex system than that a hybrid coupler can do in a wireline modem. On the other hand, a wireless transmitter often produces higher non-linearity and phase noises than its wireline counterpart. Moreover, while the wireline echo channel is essentially static, leaked and reflected wireless Tx signals inevitably has time variations in some degree, which limits the achievable cancellation.

Specifically, we analyzed the adaptive self-interferences canceller in the wireless full-duplex systems. It is shown, due to the required SNR of the receiver, it is necessary to use a long adaptation time constant as was well known for the wireline echo canceller. Such a long time constant limits the ability of tracking channel variation to yield irreducible tracking error. By introducing a linear system model of the adaptive canceller, we derived an expression of the channel estimation error of an LMS estimator as a function of its time constant and the frequency of a single tone process. This result can be extended for a wider band fading process and applicable to LS estimator with exponential windowing. It is interesting to note that the relationship between estimation error and fading frequency follows a square-law. Thus, even low frequency channel fading may cause problem if the system has a high cancellation requirement. The author is not aware of that this result has been reported previously. Thus, it would be desirable to have independent verification through analyses and simulations.

In the last section, we summarized the limiting factors to the self-interference canceller. We also provide an overall empirical expression of achievable cancellation based on the ability of the individual components.

Based on our analysis, it appears that the self-interference cancellation based in-band full duplex is most suitable for applications when the channel attenuation is not too high, especially for point to point systems and/or on the base station side. It will be a challenge for its applications in systems with high propagation loss and in complex environments.

In this author's opinion, the following areas may be worth looking into for further understanding the behaviors and for performance improvement of this technology.

- The achievable isolation between Tx and Rx signals, in various, especially complex, environments.
- Practical cancellation methods and performance of signal with non-linearities
- Fading characteristics of self-interference channels for practical applications environments
- Analog cancellation techniques, algorithms, and their behavior for Tx signal with non-linearities and multiplicative noises.

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