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Full-Duplex Wireless Communication using Transmitter Output Based Echo Cancellation*

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***Abstract*—In this paper, we present a duplexing method for wireless communication that allows both transmission and reception to occur at the same frequency band and time slot simultaneously. The self-interference caused by coupling between the transmit and receive signals is cancelled to allow the detection of the desired receive signal. This scheme can increase the capacity of wireless communication systems since only one channel instead of two is needed for duplexing. Furthermore it also opens up the possibility of using new MAC layer wireless protocols that make use of simultaneous transmission and reception to boost the overall throughput of wireless network. In this work we use a baseband transmitter output based echo cancellation approach with a 2-stage iterative echo canceller at the receiver and the reference samples are obtained from the output of the transmit power amplifier. Implementation issues are considered, including system imperfections and RF impairments such as phase noise. The system performance is evaluated by computer simulation based on IEEE 802.11a/g system parameters. It is shown that the output SINR is sufficient to support full-duplex communication, and the overall channel capacity can be increased by a factor of up to 1.8.**

I. Introduction

Efficient use of the wireless spectrum is a key objective of modern wireless communication systems. One technique for increasing spectral efficiency that has not been well explored is the area of duplexing. Most communication systems require two separate channels in order to achieve full-duplex, and common techniques that can be employed are Time Division Duplexing (TDD) or Frequency Division Duplexing (FDD), which separate the uplink and downlink channels in the time domain and frequency domain, respectively. In this paper we would like to investigate a method to transmit and receive signals on a single channel simultaneously so that one channel instead of two is needed for duplexing. One key advantage of this technology is that the channel capacity can potentially be doubled in the ideal case. It may also be applied to Cognitive Radio and Relay Systems where simultaneous transmission and reception can offer significant advances [1]. In addition, new MAC layer wireless protocols can also be designed that will boost the overall throughput of wireless network [2].

While this topic is a little unusual, some existing recent research can be found in [3]–[5]. In [3], antenna cancellation is combined with RF interference cancellation to eliminate the self-interfering signal. In [4], the authors propose one RF analog echo-cancellation method which provides isolation of 72dB for the self-interference. In addition, [5] presents some experimental results on narrow-band self-interference cancellation performance and shows that a full-duplex system based on self-interference cancellation is feasible and can achieve higher data rates than half-duplex systems. In wired communications systems there are also examples where transmit and receive signals use the same channel, as described in [6]–[8] and the references therein. The techniques used in these systems are echo cancellation, which has also been applied to on-frequency repeaters in [9] [10] to isolate transmit and receive antennas. In this paper we apply baseband transmitter output based echo cancellation to wireless systems to achieve full-duplex communication. The self-interfering signal arising from the coupling between the transmit and receive antennas of the wireless transceiver is cancelled from the received signal so that the wanted signal can be retrieved. Since the suppression ratio of the self-interference needs to be much higher than that in wire-line full-duplex systems, the effect of RF impairments and system imperfections has to be considered. For example ADC resolution, PA nonlinearity, local oscillator (LO) Phase noise, I/Q imbalance, time Jitter of ADC/DAC and channel variation. In this paper, we propose a transceiver structure to specifically address some of these problems. A transmitter output based 2-stage iterative echo canceller is introduced. The reference signal for performing echo cancellation is obtained from the output of the transmit power amplifier so that the problem of PA nonlinearity and transmitter I/Q imbalance can be circumvented. We target an IEEE 802.11a/g system and provide a detailed description of our system configuration in section II. Simulation results are given in section III with the overall channel capacity gain demonstrated in section IV. Finally, a conclusion is provided in section V.

II. System Architecture

We consider a basic system configuration with one base station and one mobile station as shown in Fig. 1. Both the forward link and the reverse link transmit at the same frequency band and time slot simultaneously and different antennas are utilized for transmission and reception at each terminal. For either terminal, the received signal consists of three components: a strong self-interference due to coupling between the transmit and receive antennas (denoted here as near-end signal), a weak wanted signal (far-end signal), and

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noise. The near-end signal is removed from the received signal by echo cancellation in order to receive the far-end signal.

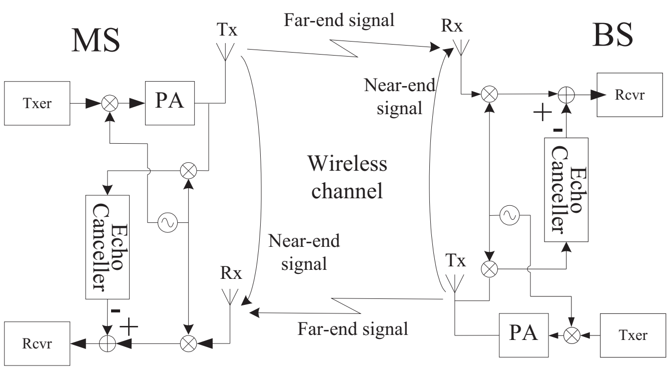

Fig. 1. Basic configuration: BS and MS signals on the same frequency band

In order to achieve an adequate suppression of the near-end signal, the RF and analog impairments within the transceiver need to be addressed as follows;

- Dynamic Range of Analog/Digital converter (ADC)
- Nonlinearity of Power Amplifier (PA)
- Phase Noise of Local Oscillator (LO)
- I/Q imbalance
- Frequency offset of far-end signal
- Time Jitter of ADC/DAC
- Variation of wireless channel
- Power Delay Profile of Wireless Channel

In this contribution, we apply baseband transmit output based echo cancellation to the received signal. The output of the PA is down-converted and fedback as a reference signal, as is shown in Fig. 1. The key advantage of this approach is that the I/Q imbalance introduced during up-conversion and the nonlinearity of the power amplifier are excluded from the signal path and thus will not affect the cancellation precision of the near-end signal. An alternative approach is to use digital pre-distortion in the transmitter as in [11] so that the I/Q imbalance and PA nonlinearity are compensated, and this will be discussed elsewhere.

Frequency offset of the far-end signal arises because independent LOs are used at the mobile station and base station and this can be compensated digitally. The variation of wireless channel can be neglected by properly reducing the frame length during processing. In addition, according to [12], the relative received signal power in a typical office decays exponentially with respect to the excess delay. Therefore most of the received near-end signal can be cancelled by FIR (finite impulse response) filters with a reasonable number of taps. However, the LO phase noise, time jitter of ADC/DAC and quantization noise introduced during AD conversion are difficult to compensate and will limit the maximum achievable suppression ratio of the near-end signal. Since it is desired that the near-end signal can be suppressed to the level of the noise floor, our technique is more suitable for systems working in relatively static environments with low transmit power, which is why we target an IEEE 802.11a/g system in this paper.

A. Transceiver Structure

We propose a transceiver structure as in Fig. 2. For the simplicity of illustration, it is denoted in baseband equivalent form, with procedures like up-conversion and down-conversion omitted. The output samples of the OFDM modulator, with a baud rate of $1/T$, are passed through a pulse shaping filter. The output samples are subsequently transformed into an analog signal, up-converted (not shown in Fig. 2) and passed through the Power Amplifier. I/Q imbalance will be introduced during up-conversion (shown as 'I/Q error' in Fig. 2).

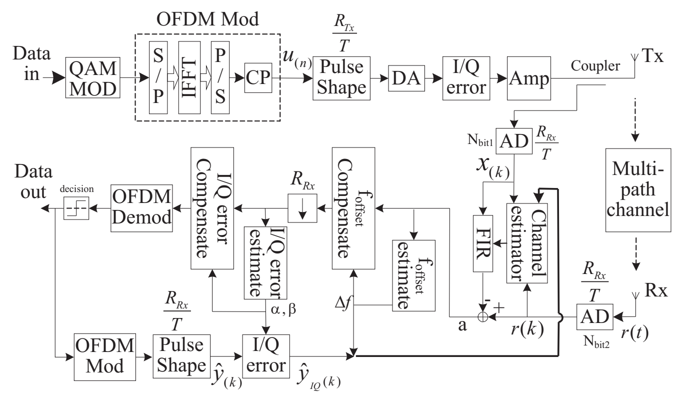

Fig. 2. Structure of proposed full duplexing transceiver

At the receiver side, the signal is sampled and passed to the echo canceller to remove the near-end signal before detecting the far-end data. The PA output is used as a reference signal for echo cancellation ($x(k)$ in Fig. 2). A coupler is attached to the transmit antenna for this purpose. It is assumed that a one-stage down-converter and digital demodulation is employed as in [11], and therefore the received signal and reference signal are free of demodulation error during down-conversion. Furthermore, we assume that the distance between transmit and receive antenna is 1 meter, which corresponds to a path loss of about 40dB for a carrier frequency of 2.4GHz. Since the maximum allowable transmit power is on the order of 20dBm for an IEEE 802.11 system, in order for the quantization error of the received near-end signal to be lower than the noise floor, a 14bit ADC is required, which is high but within current practical implementation limits.

Due to the effect of PA nonlinearity, the output signal from the power amplifier is no longer strictly band-limited, and baud rate sampling is not sufficient to represent it. Therefore in order to achieve a sufficient accuracy of near-end signal cancellation, the sampling frequency of the received signal and reference signal needs to be sufficiently high and is chosen to be an integral multiple of the baud rate in this paper, i.e. R_{Rx}/T as in Fig. 2.

The cancellation of the near-end signal is carried out on a frame basis. Basically, the impulse of the near-end channel is identified, with which replicas of the near-end signal are generated and subtracted out from the received samples. The

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cancellation procedure is made up of 2 stages as follows. In the first stage, the channel is estimated based on the whole frame of near-end reference samples $x(k)$ and received samples $r(k)$. Due to the existence of the far-end signal, which acts as an interference in channel estimation, the near-end channel identification is not precise enough and thus the suppression ratio of the near-end signal is not sufficiently high. However, the output far-end signal to residual near-end signal and noise ratio (SINR) is high enough to support the detection of far-end data with a mild error rate. Therefore the frequency offset and Tx I/Q imbalance of the far-end signal can be estimated and compensated, and a hard decision of the far-end message is determined.

In the second stage, the far-end message detected in the first stage is re-modulated. An estimate of the far-end transmit signal $\hat{y}_{IQ(k)}$ is then constructed by the pre-known pulse shaping filter and an estimation of the far-end transmitter I/Q imbalance parameters. This signal, although different from the actual transmit far-end signal due to detection/estimation error and the preclusion of far-end PA nonlinearity, is fed back together with the frequency offset for a more precise identification of the near-end channel, thus a finer near-end signal cancellation can be carried out. Then the far-end signal is detected again for a new round of near-end channel identification and echo cancellation.

This echo cancellation - detection - cancellation procedure is repeated until a satisfactory suppression ratio and output SINR is obtained. In this work we find that 2 iterations are adequate since the system performance is mainly limited by RF impairments and system imperfection instead of the performance of mathematical algorithm.

B. *Echo Cancellation Procedure*

We define $\mathbf{r} = [r_{(0)}, r_{(1)}, r_{(2)} \cdots \; r_{(N_f-1)}]^T$ as the received samples, where N_f is the number of samples contained in one frame. And then $\mathbf{r}$ can be approximately written as

$$\mathbf{r} \approx \mathbf{X}\mathbf{h} + e^{j\omega_0}\mathbf{F}\mathbf{Y}\mathbf{g} + \mathbf{z} \tag{1}$$

where $\mathbf{h} = [h_{(0)}, h_{(1)}, h_{(2)} ... h_{(L-1)}]^T$ and $\mathbf{g} = [g_{(0)}, \; g_{(1)}, \; g_{(2)} \; \cdots \; g_{(L-1)}]^T$ denote the Channel Impulse Response of the near-end and far-end channel respectively and L is the order of FIR filters. $\mathbf{X}$ and $\mathbf{Y}$ are defined by

$$\mathbf{X} = [\mathbf{x}_0, \mathbf{x}_1, ... \mathbf{x}_{L-1}], \mathbf{Y} = [\mathbf{y}_0, \mathbf{y}_1, ... \mathbf{y}_{L-1}] \tag{2}$$

where $\mathbf{x}_i = [x_{(-i)}, \; x_{(-i+1)}, \; \cdots \; x_{(-i+N_f-1)}]^T$ $\mathbf{y}_i = [y_{(-i)}, \; y_{(-i+1)}, \; \cdots \; y_{(-i+N_f-1)}]^T$, $i = 0, 1, ... L-1$. $x(k)$ and $y(k)$ are near-end and far-end transmit signal, respectively, and $x(k)$ is obtained from the transmit antenna as has been stated. To model the frequency offset of far-end signal,

$$\mathbf{F} = diag\left(\left[1, e^{j2\pi\omega}, e^{j2\pi\omega\times 2}, ...e^{j2\pi\omega\times(N_f-1)}\right]\right) \tag{3}$$

is incorporated in (1), where $\omega = (\Delta f T)/R_{Rx}$ and Δf is the carrier frequency difference between near-end and far-end user, while the term $e^{j\omega_0}$ in (1) denotes the initial phase error.

Note that the precision of (1) is limited by the existence of system impairments and more importantly, the sampling frequency of $x_{(k)}$ and $r_{(k)}$, i.e. the oversampling factor R_{Rx}, as has been stated. In addition, due to the propagation delay of the channel, the far-end signal arrives at the receiver later than the near-end signal. However, (1) still holds by re-defining $y_{(k)}$ as a delayed version of the actual far-end transmit signal. That is, the samples from the previous or next frame may be required for detecting the current frame of the far-end data, since it cannot be guaranteed that the frame is sliced in accordance with the boundary of the far-end OFDM symbols. This also needs to be considered when reconstructing far-end signals $\hat{y}_{(k)}$ (see Fig. 2) in the 2nd stage of echo cancellation.

In the 1st stage of echo cancellation, an estimate of the near-end channel impulse is obtained by LS algorithm, i.e. $\hat{\mathbf{h}} = \left(\mathbf{X}^H\mathbf{X}\right)^{-1}\mathbf{X}^H\mathbf{r}$ and an estimate of the near-end signal can be generated by $\mathbf{X}\hat{\mathbf{h}}$ and used for echo cancellation.

In the 2nd stage, we define $\widehat{\mathbf{Y}}$ and $\widehat{\mathbf{F}}$, which have the structure as (2) and (3) respectively, and are constructed based on the detection result of the far-end data. The elements of $\widehat{\mathbf{Y}}$ are $\hat{y}_{IQ(k)}$, and are obtained using $\hat{y}_{IQ(k)} = \alpha\hat{y}_{(k)} + \beta\hat{y}^*_{(k)}$, where α and β are the parameters of the transmitter I/Q imbalance of the far-end signal and are estimated using the method in [13]. $\hat{y}_{(k)}$ is an estimate of the far-end transmit signal and is constructed from the detected far-end data by a pulse shaping filter. The near-end channel impulse response can then be estimated as follows

$$\begin{bmatrix} \hat{\mathbf{h}} \\ e^{j\omega_0}\mathbf{g} \end{bmatrix} = \left(\mathbf{A}^H\mathbf{A}\right)^{-1}\mathbf{A}^H\mathbf{r} \tag{4}$$

where $\mathbf{A}$ is defined as $\mathbf{A} = [\mathbf{X} \quad \widehat{\mathbf{F}}\widehat{\mathbf{Y}}]$. Since the far-end signal is taken into account, $\hat{\mathbf{h}}$ would be more precise than that in the 1st stage. Thus a finer cancellation can be achieved.

In the following section we will provide some numerical results of the actual performance in terms of output SINR, and system capacity improvement will be investigated in sections IV.

III. Simulation Results

In this section we provide simulation results for our system. We choose the transmit power as 15dBm. The transmit messages are mapped to QPSK symbols and then passed to the OFDM modulator. The FFT size and CP (Cyclic Prefix) length are 64 and 16, respectively. Of all the 64 sub-carriers, 48 channels are used to carry data, 4 channels are allocated for pilots, and the others are null. No convolution coding or data scrambling is involved in the simulation however. The pulse shaping filter complies with the IEEE 802.11 standard, i.e. Hanning windowed sinc function, with oversampling factor R_{Tx}=16. The I/Q imbalance introduced in the up-conversion is assumed to exhibit an amplitude difference of 1dB and phase difference of 10 degree between I branch and Q branch, and the power amplifier model in [14] is used to model the effect of PA nonlinearity. The wireless channels are modeled as multipath fading, and the power delay profile (PDP) follows the Model D of standard IEEE 802.11 WLAN channel model in [15]. Each path has a bell Doppler spectrum shape with the maximum Doppler shift set to 6Hz, which correspond to a moving speed of 2.5Km/h for the surrounding objects when the carrier frequency is 2.4GHz. The carrier frequency offset between the near-end and far-end transmitter is set to 10 KHz. In order to incorporate the effect of LO phase noise, the received samples $r(k)$ in Fig. 2 are processed by $r(k) = e^{j\phi(k)}r(k)$ where $\phi(k)$ has a power spectrum density (PSD) similar to Fig. 4 of [16], except that n_1=-110dBc/Hz which is taken from the datasheet of one state of the art LO as in [17], and n_2=-154dBc/Hz.

As has been stated in section II, the distance between transmit and receive antennas is assumed to be 1 meter for each station. The effect of ADCs and DACs is modeled by uniform quantization. Unless otherwise stated, the oversampling factor R_{Rx} at the receiver side is selected as 4, and the echo cancellers use FIR filters with 40 taps; all the ADCs and DACs are chosen to be 14bits in the simulation. 2 iterations are performed in the 2nd stage echo cancellation. The frame length is chosen to be N_f=3840 and corresponds to the duration of 12 OFDM symbols, noting that a long frame will degrade system performance due to channel variation and LO phase noise, while small N_f will reduce the efficiency of processing. The far-end channel state information and carrier frequency offset are acquired from transmitted training symbols in the preamble [18].

We define ESR, SNR and $SINR_{out}$ as the received near-end signal to far-end signal ratio, received far-end signal to noise ratio and the output far-end signal to interference and noise ratio after echo cancellation (point "a" in Fig. 2), respectively. Using the path loss model as in [15], and denoting distance between two communication terminals as l, we plot ESR,SNR and $SINR_{out}$ in Fig. 3 for different l. The results are averaged over 100 independently generated channels. As can be anticipated, ESR increases while SNR decreases with l, since the received far-end signal power decays as l increases. The output SINR is very close to received SNR, with a difference of around 5dB. And a suppression ratio of 65dB in near-end signal can be observed, which is independent of l and is mainly limited by LO phase noise and channel variation. The results are also shown when lower resolution ADCs are used for the reference signal (N_{bits1} in Fig. 2, N_{bits2} remains 14). It can be noticed that the system performance is degraded for low Nbits1 because of inaccurate reference signal. When N_{bits1} is lower than 12, the near-end signal suppression ratio is determined by quantization error, as the precision of ADCs increases, RF and analog impairments emerge as the dominating factor that limit the system performance.

The relationship between output SINR and the sampling rate of the reference signal and received signal is shown in Fig. 4. The output SINR is plotted for different oversampling rate R_{Rx}. Note that the frame duration in terms of time is kept unchanged. When R_{Rx} changes, the total number of samples in a single frame (N_f) will change accordingly (the same for FIR filter order L). When R_{Rx} is selected as 1 or 2, the dependence of current samples on the further ones (non-causal effect) needs to be considered, therefore the received samples are delayed with respect to the reference samples in order to account for the negative taps of the FIR filter. It is revealed that the suppression ratio of near-end signal is constrained when the sampling frequency is low, since the reference signal or received signal cannot be represented by their samples accurately because of the non-linear nature of the PA. An oversampling factor higher than 4 is sufficient, at which point the system performance is limited by RF and analog impairments.

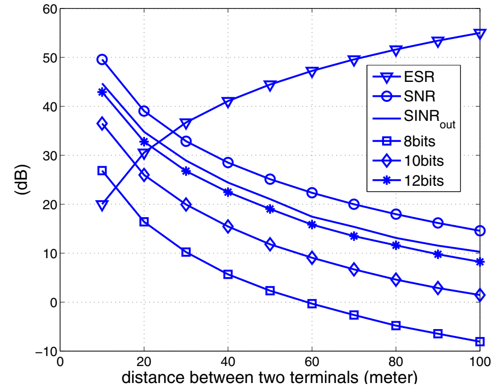

Fig. 3. ESR,SNR and Output SINR v.s. l for different N_{bit1}

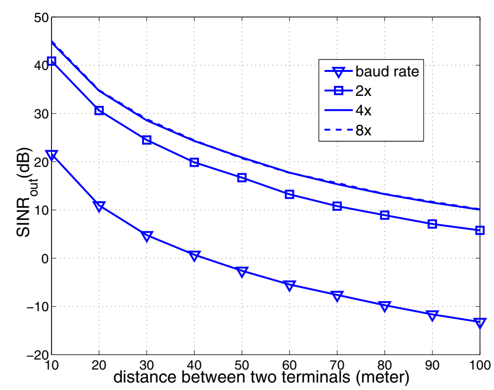

Fig. 4. Output SINR for different sampling rate

Fig. 5 compares the output SINR of the 1st stage and different iterations of the 2nd stage echo cancellation. Note that the 1st stage is equivalent to traditional echo cancellation, since only the near-end signal is utilized to identify near-end channel response. It can be seen that the introduction of the 2nd stage echo canceller offers dramatic improvement in output SINR. However, this improvement becomes almost invisible after 2 iterations. The reason is that the near-end channel identification and far-end data detection have reached their highest achievable accuracy in 2 iterations, and the system performance is mainly limited by RF impairments and system imperfections.

IV. Overall Channel Capacity Gain

With the knowledge of output SINR, we can evaluate the overall capacity gain that our system may provide compared

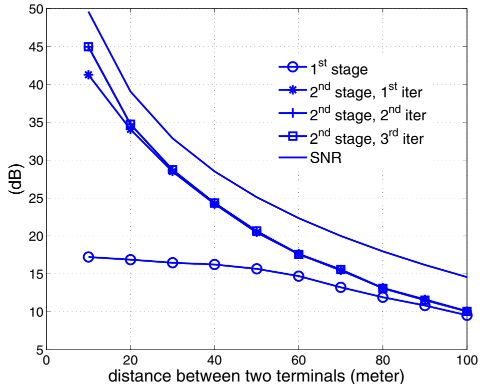

Fig. 5. SINR of 1st stage and different iterations of the 2nd Stage EC

to traditional FDD/TDD systems. Following [19], we denote the channel capacity of FDD/TDD and our full-duplex system as follows

$$
\begin{aligned}
C_1 &\approx \log_2(1+SNR) \quad bit/s\ /Hz \\
C_2 &\approx 2\log_2(1+SINR_{out}) \quad bit/s\ /Hz
\end{aligned} \tag{5}
$$

where it should be noted that our system has an extra factor of 2 due to channel reuse. We then define the overall capacity gain achieved by our system as the ratio between C_2 and C_1, and Fig. 6 shows the relationship between this gain and l. It is revealed that the total channel capacity can be improved by about 1.8 times when two terminals are close together. This gain decrease as l increases, and drops to 1.4 when l=100m, which however is still satisfactory. The results are also shown for different levels of phase noise. We can see that the channel capacity improvement is very close to 2 when no phase noise exists. However, this gain decreases significantly as the phase noise become stronger. When the phase noise level (n_1 in [16], Fig. 4) increases to -90dBc/Hz, the channel capacity can only be improved when l is very small, which reveals that RF and analog impairments are the key factor that limit the performance of our system.

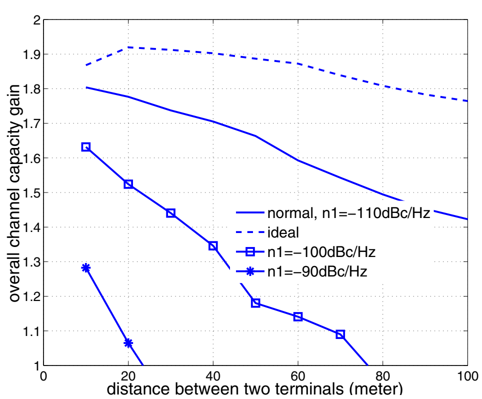

Fig. 6. Overall channel capacity improvement

V. Conclusion

In this paper we have proposed a new duplexing technique for wireless systems that allows two communicating users to transmit and receive signals simultaneously in the same channel. A transmitter output based 2-stage iterative echo canceller with frame-based signal processing is proposed and implementation issues including ADC resolution, Power Amplifier nonlinearity, I/Q imbalance, local oscillator (LO) phase noise are considered. It is shown by simulation that the overall channel capacity can be increased by a factor between 1.4 and 1.8 using our method, depending on the distance between the two communication terminals.

The proposed architecture could also allow the development of new MAC layer wireless protocols and thus can further increase system throughput. Moreover, it would be helpful in applications such as Cognitive radio and relay systems where simultaneous transmission is also a useful characteristic.

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