Broadband Analog Network Coding With Robust Processing for Two-Way Relay Networks

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***Abstract*— In this letter, we study the robust processing for a single carrier two-way relay network with broadband analog network coding and imperfect channel state information. Based on a statistical model for channel estimation error, we derive the optimal coefficients of self interference cancellation and linear frequency domain equalization in the sense of minimizing the conditional mean square error given a channel estimate. Simulation results show that the proposed robust design can achieve better bit error rate performance than the conventional non-robust scheme.**

***Index Terms*— Broadband analog network coding, frequency domain equalization, robust processing, self interference cancellation, two-way relay network.**

I. Introduction

FUTURE wireless networks should provide high throughput to meet the explosive demand for high date rate services while guaranteeing the transmission reliability over detrimental wireless fading channels. Analog network coding (ANC) has been considered as a promising technique to improve the throughput of a two-way relay network (TWRN) because it needs only two time slots to exchange the information between two source nodes with the help of a relay node [1]. To enable the ANC in frequency selective fading channels, ANC should be combined with orthogonal frequency division multiplexing (OFDM) or single carrier (SC) frequency domain equalization (FDE) technology [2], which is referred to as broadband ANC (BANC). It has been shown that the BANC with SC-FDE has similar performance to that with OFDM, but has much lower peak-to-average power ratios (PAPRs) [2]. Comparing with other physical-layer network coding (NC) techniques in SC TWRN [3], [4], BANC with SC-FDE has lower complexity since the relay node only needs to amplify-and-forward the superposed signal.

Most of the existing works on the BANC design or related signal processing for TWRNs assume perfect channel state information (CSI). However, because of imperfect channel estimation in practical networks, the two source nodes in a TWRN cannot completely remove their self interferences from the superposed signal [5]. Furthermore, as the propagation channels in the two time slots with BANC are cascaded due to the amplify-and-forward strategy, the FDE performance in an SC TWRN is more sensitive to imperfect CSI than that in a point-to-point communication system. Although there have been some works focusing on the analysis of performance degradation due to imperfect CSI [5]–[12] and the techniques to improve the channel estimation quality in TWRNs [13], [14], no work has been made in the perspective of robust self interference cancellation (SeIC) and FDE for BANC by considering the effect of imperfect CSI to the best of our knowledge.

In this letter, we investigate the robust processing for an SC TWRN with BANC and imperfect CSI. Based on a statistical model for characterizing the channel estimation error, we first derive the mean square error (MSE) after imperfect SeIC and linear FDE with imperfect channel knowledge. From the derived conditional MSE, we find that different from the conventional point-to-point SC communication systems, where the imperfect CSI increases the power of the remaining inter-symbol interference after FDE [15], in the TWRN, the imperfect CSI further leads to a component of remaining self interference due to imperfect cancellation. Aiming at minimizing the conditional MSE, we then derive the optimal coefficients for robust SeIC and FDE. Simulation results show that the proposed robust design can achieve a gain of 1dB in E_b/N_0 over the conventional non-robust signal processing with BANC.

II. Network Model

We consider an SC TWRN consisting of two source nodes, A and B, and a relay node R. It is assumed that the two source nodes cannot communicate directly without the help of the relay node R, because either they are too far away to reach each other or there is some obstacle between them. It is also assumed that each node is equipped with only one antenna and operates in the half-duplex mode.

We consider the two-time-slot ANC for the information exchange between A and B with the help of R [2]. In the first time slot, assume that the information bits at nodes A and B are mapped to quadrature amplitude modulation (QAM) symbols and every N QAM symbols, denoted by $x_i[n]$ with normalized symbol energy $E\{|x_i[n]|^2\} = 1$, for $i \in \{A, B\}$ and $n = 0, \ldots, N-1$, are packed in one block for the SC transmission. It is assumed that all the channels in the network suffer from frequency-selective fading. To deal with such fading effect, it is assumed that a cyclic prefix (CP), which is repetition of the last part of the data block, is inserted in front of each block at the transmitters of A and B. With the help of CP and by assuming perfect time and frequency synchronization in the network, the superposed baseband signal received at node R can be expressed as

$$y_R[n] = \sqrt{P_A}\sum_{l=0}^{L-1} h_A[l]x_A[(n-l)_{\mathrm{mod}N}] + \sqrt{P_B}\sum_{l=0}^{L-1} h_B[l]x_B[(n-l)_{\mathrm{mod}N}] + u_R[n], \quad (1)$$

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where $(.)_{\text{mod}N}$ denotes the module N operation, P_A and P_B denote the transmit power of nodes A and B, respectively, and u_R denotes the noise at node R, which is modeled as a circularly symmetric Gaussian variable with variance σ_u^2, i.e., $u_R \sim \mathcal{CN}(0, \sigma_u^2)$. Finally, $h_i[l]$ with $l = 0, \ldots, L-1$ denotes the L-tap channel impulse response (CIR) between node i and node R, for $i \in \{A, B\}$. In a typical wireless network, $h_i[l]$ is also modeled as a circularly symmetric Gaussian variable, i.e., $h_i[l] \sim \mathcal{CN}(0, \sigma_{i,l}^2)$, and is independent for different taps.

In the second time slot, node R normalizes its received signal in the first time slot by a factor $\beta = \sqrt{E\{|y_R[n]|^2\}}$, amplifies it with power P_R, and broadcasts to nodes A and B. Assuming that the channels between the two source nodes and the relay node are reciprocal in the two time slots, the received signals at nodes A and B can be represented as

$$y_i[n] = \frac{\sqrt{P_R}}{\beta} \sum_{l=0}^{L-1} h_i[l] y_R[(n-l)_{\text{mod}N}] + u_i[n], \quad i \in \{A, B\}, \tag{2}$$

where u_i denotes the noise at node i. It is assumed that all the three nodes have the same noise power. It is worth noting that in the ANC, the form in (2) implies that the CP is long enough to cover the delay spread of the cascaded channel in the two time slots. Then, nodes A and B cancel their self interferences embedded in the received signals, equalize the resulting signal, and finally retrieve the symbols from the other source node [2].

III. Robust Design for BANC

In this section, we first introduce a statistical model to characterize the channel estimation error and then derive the optimal robust coefficients for SeIC and FDE for BANC with imperfect CSI. In the two-time-slot ANC, both nodes A and B need the information of CIRs of h_A and h_B in order to perform SeIC and FDE. It is assumed that in the first time slot, before data transmission, both source nodes sequentially send a training sequence (TS) for the relay node to estimate the two channels separately. The estimated CIRs are assumed to be perfectly fed back to nodes A and B in the second time slot for their use in SeIC and FDE. It is also assumed that node R sends a TS before the data forwarding in the second time slot. In summary, node A (B) gets the CIR of its self channel, i.e., h_A (h_B), from its own estimation and the CIR of the other channel, i.e., h_B (h_A), from the feedback of the relay node R.

To investigate the robust design for BANC, we apply the classical statistical model for the channel estimation error [16], [17], where the relationship between the estimated CIR, $\widehat{h}_i[l]$, and the true one, $h_i[l]$, can be expressed as follows

$$\widehat{h}_i[l] = h_i[l] + \varepsilon_i[l], \quad \text{for } l = 0, 1 \ldots, L-1, \tag{3}$$

and $i \in \{A, B\}$, where $\varepsilon_i[l] \sim \mathcal{CN}(0, \zeta_{i,l}^2)$ models the estimation error component and is assumed to be independent for different taps.

The baseband signal processing at the receivers of nodes A and B consists of two steps: SeIC and FDE. The imperfect channel knowledge affects both of them. In the following, we first derive the MSE conditioned on a given channel estimate after the two-step processing, and then derive the optimal robust SeIC and FDE coefficients by minimizing the conditional MSE.

The processing at nodes A and B is exactly the same. Without loss of generality, we take node A as an example. By applying the normalized discrete fourier transform (DFT) to both sides of (2), we have

$$Y_A[k] = \frac{\sqrt{P_R}}{\beta} H_A[k] Y_R[k] + U_R[k], \quad \text{for } k = 0, \ldots, N-1, \tag{4}$$

where $Y_A[k] = \frac{1}{\sqrt{N}} \sum_{n=0}^{N-1} y_A[n] e^{-j2\pi nk/N}$, $U_R[k] = \frac{1}{\sqrt{N}} \sum_{n=0}^{N-1} u_R[n] e^{-j2\pi nk/N}$, and $H_A[k] = \sum_{l=0}^{L-1} h_A[l] e^{-j2\pi kl/N}$ denotes the channel frequency response.[1] With its own estimated CSI $\widehat{H}_A[k]$, node A performs SeIC as follows

$$\overline{Y}_A[k] = Y_A[k] - \alpha_A[k] X_A[k], \tag{5}$$

where $\alpha_A[k]$ is the SeIC coefficient and will be optimized later. The signal after SeIC is then linearly equalized in the frequency domain. That is,

$$\widetilde{x}_B[n] = \frac{1}{\sqrt{N}} \sum_{k=0}^{N-1} W_A[k] \overline{Y}_A[k] e^{-j2\pi nk/N}, \tag{6}$$

where $W_A[k]$ is the coefficient in the k-th frequency tone. Define the MSE conditioned on a given channel estimate as

$$J_B \triangleq E\{|\widetilde{x}_B[n] - x_B[n]|^2 \big| \widehat{\mathbf{h}}_A, \widehat{\mathbf{h}}_B\}, \tag{7}$$

our target is to find the optimal SeIC coefficients $\alpha_A[k]$ and FDE coefficients $W_A[k]$ to minimize J_B. By substituting (6) into (7) and using the property that the transmit symbols from nodes A and B and the noise component are independent to each other, we have

$$J_B = \frac{1}{N} E\{\Big| \sum_{k=0}^{N-1} (\overline{Y}_A[k] W_A[k] - X_B[k]) e^{j2\pi nk/N} \Big|^2 \big| \widehat{\mathbf{h}}_A, \widehat{\mathbf{h}}_B\}. \tag{8}$$

By defining the expectation of each term in the summation in (8) as $J_B[k]$ and by substituting (4) and (5) into it, we have

$$\begin{aligned} J_B[k] = E\{&\Big| \frac{\sqrt{P_B P_R}}{\beta} H_B[k] H_A[k] W_A[k] - 1 \Big|^2 \\ &+ |X_A[k] W_A[k]|^2 \Big| \frac{\sqrt{P_A P_R}}{\beta} H_A^2[k] - \alpha_A[k] \Big|^2 \\ &+ \sigma_u^2 |W_A[k]|^2 (|\frac{\sqrt{P_R}}{\beta} H_A[k]|^2 + 1) \big| \widehat{\mathbf{h}}_A, \widehat{\mathbf{h}}_B\}. \end{aligned} \tag{9}$$

The key issue is to derive the conditional means, auto-correlations, cross-correlations of H_A and H_B, and fourth moment of H_A, given the channel estimate $\widehat{\mathbf{h}}_A$ and $\widehat{\mathbf{h}}_B$.

It is assumed that the two channels between the two source nodes and the relay node are independent to each other. Thus, the above conditional statistics of a channel are only determined by the channel estimate of itself (not related to

[1] Throughout this letter, we use the normalized DFT/IDFT to represent the time/frequency domain relationship for the transmitted, received, and noise signals, and express the channel frequency response as the un-normalized DFT of CIR.

the other channel). Furthermore, since it is assumed that the channel gains are independent for different channel taps, i.e., $h_i[l]$ and $h_i[\tau]$, where $i \in \{A, B\}$, are independent for $l \neq \tau$, we have $\widetilde{h}_i[l] \triangleq E\{h_i[l]|\widehat{\mathbf{h}}_i\} = E\{h_i[l]|\widehat{h}_i[l]\}$ and $E\left\{h_i[l]h_i^*[\tau]|\widehat{\mathbf{h}}_i\right\}$ is equal to $E\{|h_i[l]|^2|\widehat{h}_i[l]\}$ if $l = \tau$ and $\widetilde{h}_i[l]\widetilde{h}_i^*[\tau]$ otherwise. The key issue is to find the conditional probability density function (PDF) of $h_i[l]$ given $\widehat{h}_i[l]$, which is equal to their joint PDF divided by the marginal PDF of $\widehat{h}_i[l]$. According to [16] and [17], we get the conditional mean as

$$\widetilde{h}_i[l] \triangleq E\{h_i[l]|\widehat{\mathbf{h}}_i\} = \frac{\sigma_{i,l}^2}{\sigma_{i,l}^2 + \zeta_{i,l}^2}\widehat{h}_i[l], \tag{10}$$

the conditional variance as

$$c_{i,l}^2 \triangleq E\{|h_i[l] - \widetilde{h}_i[l]|^2|\widehat{\mathbf{h}}_i\} = \frac{\sigma_{i,l}^2\zeta_{i,l}^2}{\sigma_{i,l}^2 + \zeta_{i,l}^2}, \tag{11}$$

and the conditional auto-correlation as

$$E\left\{h_i[l]h_i^*[\tau]|\widehat{\mathbf{h}}_i\right\} = \widetilde{h}_i[l]\widetilde{h}_i^*[\tau] + c_{i,l}^2\delta[l-\tau], \tag{12}$$

where $\delta[l - \tau]$ is equal to 1 if $l = \tau$, and 0 otherwise. To derive the conditional mean of the channel frequency response, which is defined as $\widetilde{H}_i[k] \triangleq E\{H_i[k]|\widehat{\mathbf{h}}_i\}$, since $H_i[k] = \sum_{l=0}^{L-1} h_i[l]e^{-j2\pi kl/N}$, we have

$$\widetilde{H}_i[k] = \sum_{l=0}^{L-1} E\{h_i[l]|\widehat{\mathbf{h}}_i\}e^{-j\frac{2\pi}{N}kl} = \sum_{l=0}^{L-1}\widetilde{h}_i[l]e^{-j\frac{2\pi}{N}kl}. \tag{13}$$

Similarly, we have the conditional autocorrelation of $H_i[k]$ as

$$E\{|H_i[k]|^2|\widehat{\mathbf{h}}_i\} = \sum_{l=0}^{L-1}\sum_{\tau=0}^{L-1} E\{h_i[l]h_i^*[\tau]|\widehat{\mathbf{h}}_i\}e^{-j\frac{2\pi}{N}k(l-\tau)}. \tag{14}$$

By substituting (12) and (13) into (14), we have

$$E\{|H_i[k]|^2|\widehat{\mathbf{h}}_i\} = |\widetilde{H}_i[k]|^2 + \xi_i, \tag{15}$$

where $\xi_i \triangleq \sum_{l=0}^{L-1} c_{i,l}^2$.

Another difficulty in (9) is to derive $E\{|H_i[k]|^4|\widehat{\mathbf{h}}_i\}$. According to its definition, we have

$$\begin{aligned} &E\left\{|H_i[k]|^4|\widehat{\mathbf{h}}_i\right\} \\ &= E\left\{H_i[k]H_i^*[k]H_i[k]H_i^*[k]|\widehat{\mathbf{h}}_i\right\} \\ &= \sum_{l=0}^{L-1}\sum_{\tau=0}^{L-1}\sum_{s=0}^{L-1}\sum_{t=0}^{L-1} E\left\{h_i[l]h_i^*[\tau]h_i[s]h_i^*[t]|\widehat{\mathbf{h}}_i\right\}e^{-j\frac{2\pi}{N}(l-\tau+s-t)k}. \end{aligned} \tag{16}$$

To compute the conditional fourth moment of $h_i[l]$ in (16), according to [18], we have

$$\begin{aligned} E\left\{h_i[l]h_i^*[\tau]h_i[s]h_i^*[t]|\widehat{\mathbf{h}}_i\right\} &= \widetilde{h}_i[l]\widetilde{h}_i^*[\tau]\widetilde{h}_i[s]\widetilde{h}_i^*[t] \\ &+ \widetilde{h}_i[l]\widetilde{h}_i^*[\tau]c_{i,s}^2\delta[s-t] + \widetilde{h}_i[l]\widetilde{h}_i^*[t]c_{i,s}^2\delta[s-\tau] \\ &+ \widetilde{h}_i[s]\widetilde{h}_i^*[\tau]c_{i,l}^2\delta[l-t] + \widetilde{h}_i[s]\widetilde{h}_i^*[t]c_{i,l}^2\delta[l-\tau] \\ &+ c_{i,l}^2c_{i,s}^2\delta[l-\tau]\delta[s-t] + c_{i,l}^2c_{i,s}^2\delta[l-t]\delta[s-\tau]. \end{aligned} \tag{17}$$

By substituting (17) into (16), we have

$$E\left\{|H_i[k]|^4|\widehat{\mathbf{h}}_i\right\} = |\widetilde{H}_i[k]|^4 + 4|\widetilde{H}_i[k]|^2\xi_i^2 + 2\xi_i^4. \tag{18}$$

Furthermore, since it is assumed that $H_A[k]$ and $H_B[k]$ are independent, we have $E\{H_A[k]H_B^*[k]|\widehat{\mathbf{h}}_A, \widehat{\mathbf{h}}_B\} = \widetilde{H}_A[k]\widetilde{H}_B^*[k]$. By substituting these results into (9), and further into (8), we have the conditional MSE in (19), which is shown at the bottom of this page. It can be seen from the second line in (19) that the optimal SeIC coefficients in the sense of minimizing J_B are

$$\alpha_A[k] = \frac{\sqrt{P_AP_R}}{\beta}\widetilde{H}_A^2[k], \quad \text{for } k = 0, \ldots, N-1. \tag{20}$$

Then, by substituting (20) into (19) and setting the partial derivative of J_B with respect to $W_A[k]$ to zero, we have the optimal coefficients of the robust linear FDE as follows

$$W_A[k] = \frac{\sqrt{P_BP_R}}{\beta}\widetilde{H}_B^*[k]\widetilde{H}_A^*[k]Q_A[k], \tag{21}$$

where $Q_A[k]$ is given by

$$\begin{aligned} Q_A[k] = \Big(&\frac{P_BP_R}{\beta^2}(|\widetilde{H}_B[k]|^2 + \xi_B)(|\widetilde{H}_A[k]|^2 + \xi_A) \\ &+ |X_A[k]|^2\frac{P_AP_R}{\beta^2}\big(4|\widetilde{H}_A[k]|^2\xi_A + 2\xi_A^2\big) \\ &+ \sigma_u^2\Big(\frac{P_R}{\beta^2}(|\widetilde{H}_A[k]|^2 + \xi_A) + 1\Big)\Big)^{-1}. \end{aligned}$$

IV. Simulation Results

In the simulation, we assume that the large scale fading of both channels is the same. For the small scale fading, we consider the long term evolution (LTE) extended pedestrian A (EPA) channel [19]. The system bandwidth is set to 20MHz with a carrier frequency of 2GHz. The maximum Doppler frequency is set to 5Hz. A square-root raised cosine filter with a roll-off factor of 0.33, whose impulse response is truncated to a length of 8 symbols, is applied at both transmit and receive sides. The transmit power of three nodes is set to be the same. We assume that each block contains $N = 1024$ uncoded 4QAM symbols. The least square (LS) method is applied to channel estimation, where the time division multiplexing

$$\begin{aligned} J_B = &\frac{1}{N}\sum_{k=0}^{N-1}\left|\frac{\sqrt{P_BP_R}}{\beta}\widetilde{H}_B[k]\widetilde{H}_A[k]W_A[k] - 1\right|^2 + |W_A[k]|^2\frac{P_BP_R}{\beta^2}\left(|\widetilde{H}_B[k]|^2\xi_A + |\widetilde{H}_A[k]|^2\xi_B + \xi_A\xi_B\right) \\ &+ |X_A[k]W_A[k]|^2\left(\left|\frac{\sqrt{P_AP_R}}{\beta}\widetilde{H}_A^2[k] - \alpha_A[k]\right|^2 + \frac{P_AP_R}{\beta^2}(4|\widetilde{H}_A[k]|^2\xi_A + 2\xi_A^2)\right) + \sigma_u^2|W_A[k]|^2\Big(\frac{P_R}{\beta^2}(|H_A[k]|^2 + \xi_A) + 1\Big). \end{aligned} \tag{19}$$

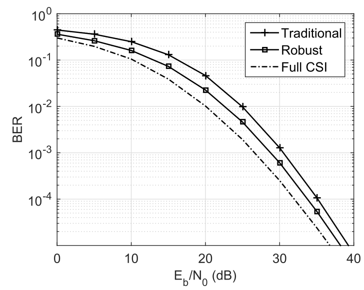

Fig. 1. BER v.s. E_b/N_0 for different schemes.

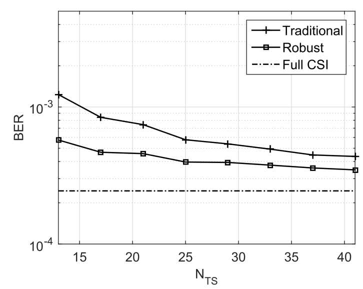

Fig. 2. BER v.s. Channel estimation quality for different schemes.

Zadoff-Chu sequence is considered for training. It is assumed that the estimated CIR of channel h_A (h_B) is perfectly fed back to node B (node A) by node R.

Fig. 1 shows the BER performance as a function of E_b/N_0 for different ANC schemes. The curve labeled as 'Traditional' is the performance of the traditional ANC without considering the effect of imperfect CSI, where the estimated CIR is directly used for SeIC and FDE at nodes A and B regardless of the channel estimation error. The curve labeled as 'Robust' is the performance of the proposed ANC with robust processing. The curve labeled as 'Full CSI' is the performance when all nodes have perfect channel knowledge and perfect SeIC. Here the length of TS (N_{TS}) is set to 13. It is shown in the figure that the 'Robust' scheme attains an approximate 1dB gain over the traditional scheme, for example, at a target BER of 10^{-5}.

Fig. 2 shows the BER performance as a function of N_{TS} when E_b/N_0 is fixed at 30dB. Since all nodes use the same TS and estimation method, the estimation error variances are the same for different channels and different taps. It can be seen that compared with the traditional scheme, the training overhead in terms of N_{TS} can be greatly reduced with the proposed robust processing.

V. Conclusion

We have investigated the robust design to reduce the impact of imperfect CSI on an SC TWRN with BANC. Based on the statistical model for the imperfect channel estimation, we derived the optimal SeIC and linear FDE coefficients under the criterion of minimizing the conditional MSE given a channel estimate. Simulation results showed that the proposed robust scheme can achieve a gain of about 1dB in E_b/N_0 than the traditional non-robust counterpart.

As for future work, it was assumed in this study that the channel coefficients of different channel taps are uncorrelated. In practical systems, they may be correlated due to the transmission pulse. It is of interest to extend the proposed robust design by considering the correlation among channel taps. Another assumption we made in this work is that the estimated CIRs can be perfectly fed back to the source nodes. It will be interesting to study the BANC with limited feedback and the robust processing with the consideration of both channel estimation errors and quantization errors. Furthermore, another way of channel estimation to avoid the CIR feedback at the relay node in BANC is to let the two source nodes estimate the CIR of the concatenated channel. The robust design in this scenario is also of interest to investigate.

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