NON-COOPERATIVE CROSS-CHANNEL GAIN ESTIMATION USING FULL-DUPLEX AMPLIFY-AND-FORWARD RELAYING IN COGNITIVE RADIO NETWORKS

Bijia Huang[†], *Guodong Zhao*[†], *Liying Li*[‡], *Xiangwei Zhou*[♭], *and Zhi Chen*[†]

[†]National Key Laboratory of Science and Technology on Communications, UESTC, Chengdu, China
[‡]Automation Engineering, UESTC, Chengdu, China
[♭]School of EECS, Louisiana State University, Baton Rouge, LA, USA

ABSTRACT

In this paper, we propose a new estimation method to obtain the cross-channel gain, which avoids the severe interference to the *primary receiver* (PR) in existing relay-assisted estimation methods. In our method, we let the cognitive transmitter add a time delay when it conducts the full-duplex amplify-and-forward relaying. This forces the *time-difference-of-arrival* (TDOA) between the direct and relay signals to be large enough rather than randomly large or small. Then we develop our estimation method only in the large TDOA case and precisely control the interference to the PR. Simulation results indicate that the proposed method can significantly reduce the interference to the PR.

1. INTRODUCTION

In cognitive radio networks, it is very challenging to estimate the cross-channel gain between the *cognitive transmitter* (CT) and the *primary receiver* (PR) [1–3] in *frequency-division duplexing* (FDD) systems. This is because the cross-channel gain can only be estimated at the PR and sent back to the CT via the backhaul link between cognitive and primary systems. However, such an assumption is not always valid.

Recently, a non-cooperative method, called *proactive estimation*, has been introduced into cognitive radio networks [4–8], which enables the CT to autonomously estimate the cross-channel gain in FDD systems. However, this method may cause severe interference to the PR since it requires the CT to send jamming signal to the PR. Very recently, a new kind of proactive estimation methods, called *relay-assisted proactive estimation*, has been proposed in [9–11] to deal with the interference issue. Instead of sending jamming signal to the PR, the CT conducts the full-duplex *amplify-and-forward* (AF) relaying to forward the primary signal to the PR. This can effectively reduce the interference to the PR. However, it requires the *time-difference-of-arrival* (TDOA) between the direct and relay signals to be small enough. In fact, this may not be true since the TDOA is actually a random variable and it is determined by the locations of the three nodes [12–14]. For example, if the CT is located in the segment between the PT and PR, the TDOA is small. However, if the CT is far away from the PT and PR, the TDOA becomes large. Therefore, the existing relay-assisted methods designed under the small TDOA case may still cause severe interference to the PR since the TDOA may be randomly small and large in practical systems.

This work was supported in part by the National Natural Science Foundation of China under Grant 61201130, by the U.S. National Science Foundation under Grants 1343380 and 1560437, and by the High-Tech Research and Development (863) Program of China under Grant 2015AA01A707. The corresponding author is Guodong Zhao (gdngzhao@gmail.com).

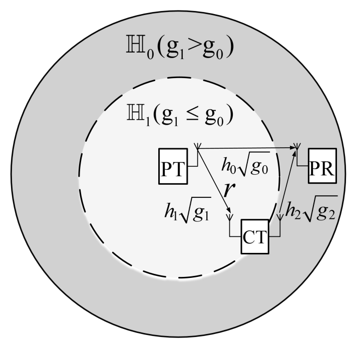

Fig. 1. System model.

In this paper, we investigate the relay-assisted cross-channel gain estimation in FDD systems and design a new estimation method to obtain the cross-channel gain, in which a preset interference probability to the PR can be satisfied. In our method, we let the CT artificially add a time delay during the full-duplex AF relay. This forces the small TDOA to be the large one no matter where the three nodes are located. Then we can design our estimation method only in the large TDOA case, which allows us to precisely control the interference to the PR.

2. SYSTEM MODEL

Fig. 1 provides the system model of this paper. In the figure, the PT serves the PR that is uniformly located inside the disk region with the center PT and the radius R. At the same time, a CT intends to estimate the cross-channel gain from the CT to the PR for spectrum sharing, and the distance between the CT and the PT is r. Here, we denote $h_k\sqrt{g_k}$ ($k = 0, 1,$ and 2) as the channel coefficients among the three nodes, where h_k and g_k represent the small-scale fading and large-scale path loss coefficients, respectively. In the small-scale fading, the coefficient h_k follows Rayleigh distribution with unit variance. In the large-scale[1] path loss [15], the coefficient g_k can be obtained by $g_k(\text{dB}) = -128.1 - 37.6\log_{10}(l)$ for $l \geq 0.035$ km, where l denotes the distance between two nodes. Furthermore, we consider the block fading channel, where the Rayleigh fading coefficients are constant within each block and they are independent for different blocks.

[1]To facilitate the development of the proposed method, we do not consider the shadowing in our model, which allows us to obtain some important closed-form expressions. However, in the simulation, we consider the shadowing to evaluate the performance of our method.

978-1-4799-9988-0/16/$31.00 ©2016 IEEE

First, we introduce the point-to-point model between the PT and PR when the CT keeps silent. Specifically, we consider the guaranteed primary service with *close-loop power control* (CLPC)[2], where the PT automatically adjusts its transmission power to maintain a certain target average *signal-to-noise ratio* (SNR) or *signal-to-interference-plus-noise ratio* (SINR), denoted as $\bar{\gamma}_T$. Then, we have the relationship between $\bar{\gamma}_T$ and p_0 as follows,

$$\bar{\gamma}_T = \frac{\mathbb{E}\left[\left|h_0(j)\sqrt{g_0 p_0}x(i,j)\right|^2\right]}{\sigma^2} = \frac{p_0 g_0 \mathbb{E}\left[\left|h_0\left(j\right)\right|^2\right]}{\sigma^2} = \frac{p_0 g_0}{\sigma^2}. \quad (1)$$

where p_0 is the transmission power of the PT, $x(i,j)$ is the transmitted signal of the PT with unit power, $\mathbb{E}\left[|x(i,j)|^2\right] = 1$, $\mathbb{E}\left[\cdot\right]$ is the expectation operator, and i and j denote the indices of N samples and M blocks. The corresponding average SNR at the CT can be obtained by $\bar{\gamma}_{c_0} = p_0 g_1/\sigma^2$.

Then, we develop the three-node relay model among the PT, CT, and PR. When the CT acts as a full-duplex AF relay with the amplitude gain G, the received signal of the PR has two components with a time delay τ, i.e.,

$$\begin{aligned} y_p(i,j) = &\underbrace{h_0(j)\sqrt{g_0 p_0}x(i,j)}_{S_d(\text{Direct signal})} + \underbrace{G h_2(j)\sqrt{g_2}h_1(j)\sqrt{g_1 p_0}x(i-\alpha, j-\beta)}_{S_r(\text{Relay signal})} \\ &+ \underbrace{G h_2(j)\sqrt{g_2}n_c(i,j) + n_p(i,j)}_{N(\text{Noise})}, \end{aligned} \quad (2)$$

where α and β are the numbers of the sample period T_s and the block period T_b, respectively, and $\tau = \alpha T_s + \beta T_b$.

Since this paper focuses on proposing the new idea and demonstrating its feasibility, we ignore the self-interference in full-duplex relay to facilitate the algorithm development. In Section 4, we will evaluate the performance loss caused by the imperfect self-interference suppression.

3. CROSS-CHANNEL GAIN ESTIMATION

3.1. Basic Principle

To control the interference to the PR, our idea is to let the CT artificially add a time delay when it conducts the full-duplex AF relaying. Then we can design our estimation method to meet a preset interference probability. This is because when the CT conducts the full-duplex AF relaying, the received signal at the PR has two components: one is the direct signal from the PT and the other is the relay signal from the CT. If the TDOA τ between the direct and relay signals is less than the *maximum allowable TDOA*[3] of the PR, i.e., $\tau < T_m$, it is the small TDOA case. Otherwise, if the TDOA τ is equal to or larger than the maximum allowable TDOA, i.e., $\tau \geq T_m$, it is the large TDOA case. In practice, since the small and large TDOA cases may randomly occur, the existing relay-assisted methods [9–11] designed under the small TDOA case may still cause severe interference to the PR. To deal with the issue, we can design a new method for the large TDOA. However, this is not enough because the CT needs to be capable of identifying its current TDOA and choosing the correct estimation method. To simplify the design, our idea is to let the CT add a time delay τ' ($\tau' \geq T_m$) during the full-duplex relaying. Then we can artificially convert the small TDOA to the large one. This allows us to design the estimation method only in the large TDOA case and precisely control the interference to the PR.

[2]This assumption has been used in the literatures on proactive estimation, e.g., [4–8] and [16], since the CLPC has been widely adopted in modern wireless systems.

[3]Generally, depending on the system bandwidth and also the signal processing ability, the PR may only collect the received signals within a certain time duration, which is called the maximum allowable TDOA T_m. Depending on the relationship between the maximum allowable and actual TDOAs, the PR may treat the two signals in different ways. If $\tau < T_m$, the PR treats both direct and relay signals as the desired signals. If $\tau \geq T_m$, the PR only treats one of them as the desired signal and the other as the interference.

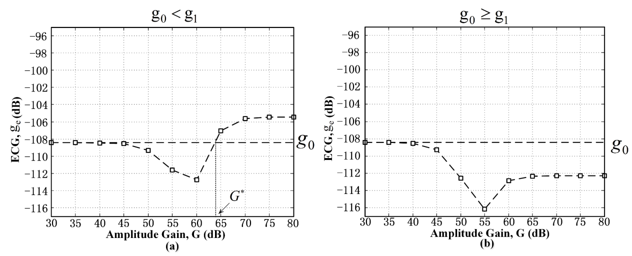

Fig. 2. EGG versus the amplitude gain for different relationships between g_0 and g_e.

3.2. Estimator Design

In this subsection, we first derive the *equivalent channel gain* (ECG) between the PT and PR, denoted as g_e, which indicates the impacts of the CT's full-duplex relaying on the primary link. Then, we discuss the relationship between the original PT-PR channel gain g_0 and the ECG g_e to obtain the proposed estimator.

When we add the time delay τ' to the full-duplex relay, the TDOA between the direct and relay signals becomes larger than the maximum allowable TDOA, i.e., $\tau' + \tau > T_m$. Then, the PR treats the strong one (between the direct and relay signals) as the desired signal and the other as the interference. Since either the direct signal from the PT or the relay signal from the CT can be the strong one, we discuss them respectively.

• *Strong Direct Signal*: The PR treats the direct signal as its desired signal as long as the direct signal can provide higher SINR than what the relay signal can provide. Then the ECG can be obtained by

$$g_e = \frac{g_0}{2}\left(\frac{1}{\frac{G^2 p_0 g_1 g_2}{\sigma^2} + G^2 g_2 + 1} + 1\right). \quad (3)$$

• *Strong Relay Signal*: The PR treats the relay signal as its desired signal as long as the relay signal can provide equal or higher SINR than what the direct signal can provide. Then the ECG can be obtained by

$$g_e = \frac{G^2 g_1 g_2}{\frac{p_0 g_0}{\sigma^2} + G^2 g_2 + 1}. \quad (4)$$

Theorem 1. *In the large TDOA, the relay enhances the primary link if the first hop of the relay channel is stronger than the primary channel and the amplitude gain is greater than the value G^*. Otherwise, the relay degrades the primary link, i.e.,*

$$\begin{cases} g_0 < g_e, & \text{if } g_1 > g_0 \text{ and } G > G^*, \quad (5a) \\ g_0 \geq g_e, & \text{if } g_1 \leq g_0 \text{ or } G \leq G^*, \quad (5b) \end{cases}$$

where

$$G^* = \sqrt{\left(\bar{\gamma}_T^2 + \bar{\gamma}_T\right)/\left(\left(\bar{\gamma}_{c_0} - \bar{\gamma}_T\right) g_2\right)}. \quad (6)$$

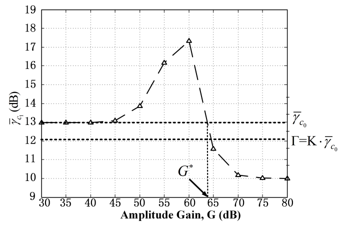

Fig. 3. The measured SNR of the CT $\bar{\gamma}_{c_1}$ versus the amplitude gain G, where the target SNR of the PR is $\bar{\gamma}_T = 10$ dB.

From the above theorem[4], we find that the ECG g_e is equal to the original PT-PR channel gain g_0 once the amplitude gain G is equal to the value G^*. In particular, the expression (6) indicates the relationship among the four values, i.e., the CT amplitude gain G^*, the PR target SNR $\bar{\gamma}_T$, the CT measured SNR $\bar{\gamma}_{c_0}$, and the cross-channel gain g_2. Since the variable $\bar{\gamma}_{c_0}$ is known to the CT, we need to obtain the first two valuables, i.e., $\bar{\gamma}_T$ and G^*. Then based on (6), the CT can estimate the cross-channel gain, i.e.,

$$\hat{g}_2 = \frac{(\bar{\gamma}_T^2 + \bar{\gamma}_T)}{G^{*2}(\bar{\gamma}_{c_0} - \bar{\gamma}_T)}. \tag{7}$$

3.3. Interference Control

In the previous subsection, we have developed our cross-channel gain estimator in (7), which requires the CT to conduct the full-duplex AF relaying and autonomously obtain the two variables. In this subsection, we discuss how to conduct the relaying and obtain the two variables to meet a preset interference probability.

3.3.1. Conduct the Full-Duplex AF Relaying

As indicated in (5a), the CT needs to satisfy the two interference-free conditions to conduct the relaying: 1) $g_1 > g_0$; 2) $G > G^*$. In practice, since it is easy to satisfy the second condition by using a large amplitude gain, the difficulty is to find the case that meets the first condition.

Here, we treat the above difficulty as a detection problem by defining $\mathbb{H}_0$ and $\mathbb{H}_1$ as the two hypotheses $g_1 > g_0$ and $g_1 \le g_0$, respectively. We choose the measured SNR $\bar{\gamma}_{c_0}$ at the CT as the test statistic. Then, the CT can distinguish $\mathbb{H}_1$ and $\mathbb{H}_0$ according to the following rule:

$$\text{Decision result} = \begin{cases} \mathbb{H}_0(g_1 > g_0), & \text{if } \bar{\gamma}_{c_0} > \eta, \\ \mathbb{H}_1(g_1 \le g_0), & \text{if } \bar{\gamma}_{c_0} \le \eta, \end{cases} \tag{8}$$

where η is the threshold. In this part, the interference probability to the PR can be obtained by

$$P_{I_1} = \Pr\{\mathbb{H}_0|\mathbb{H}_1\}\Pr\{\mathbb{H}_1\} = \int_\eta^{+\infty} f(z)dz \cdot \frac{r}{R}, \tag{9}$$

where the close-form of the *probability density function* (PDF) $f(z)$ will be presented in our journal version.

3.3.2. Obtain the Target SNR $\bar{\gamma}_T$ and the Amplitude Gain G^*

We provide Fig. 3 to depict the CT measured SNR $\bar{\gamma}_{c_1}$ during the relaying versus the amplitude gain G. Here, we also mark the CT measured SNR $\bar{\gamma}_{c_0}$ before conducting the relaying for comparison. From the figure, we observe that the SNR $\bar{\gamma}_{c_1}$ curve is the up-side-down version of the ECG curve in Fig. 2-(a). In particular, if $\bar{\gamma}_{c_1} < \bar{\gamma}_{c_0}$, it indicates that the relay does not cause interference to the PR. Meanwhile, when the amplitude gain is large enough, the measured SNR $\bar{\gamma}_{c_1}$ is equal to the target SNR $\bar{\gamma}_T$, i.e.,

$$\lim_{G\to\infty} \bar{\gamma}_{c_1} = \lim_{G\to\infty} \frac{g_1 p_1}{\sigma^2} = \bar{\gamma}_T g_1 \lim_{G\to\infty} \left(\frac{1}{g_1} + \frac{\bar{\gamma}_T + 1}{G^2 g_1 g_2}\right) = \bar{\gamma}_T, \tag{10}$$

where $p_1 = \sigma^2 \bar{\gamma}_T / g_e$ is the transmission power of the PT during the relaying. Therefore, when conducting the relaying with a large amplitude gain $G_{\max}$, where $G_{\max} > G^*$, the CT is able to obtain the target SNR of the PR by $\widehat{\bar{\gamma}_T} = \bar{\gamma}_{c_1}(G_{\max})$.

On the other hand, when the amplitude gain G reduces from $G_{\max}$ to G^*, the measured SNR $\bar{\gamma}_{c_1}$ increases to $\bar{\gamma}_{c_0}$. Thus, it is reasonable for the CT to conduct the full-duplex relaying and gradually reduce the amplitude gain. Once a stop threshold $\bar{\gamma}_{c_1} = \bar{\gamma}_{c_0}$ is satisfied, the corresponding value G^* can be obtained. Theoretically, if the amplitude gain G is reduced continuously, the value G^* can be obtained without interfering with the PR, i.e., $\bar{\gamma}_{c_1} < \bar{\gamma}_{c_0}$ always holds. However, in practice, since the amplitude gain G is adjusted by a reducing step $\Delta G > 0$ dB, the value G^* is actually obtained at $\bar{\gamma}_{c_1} > \bar{\gamma}_{c_0}$, which introduces interference to the PR. To limit the interference, we adjust the stop threshold by a coefficient $0 < K \le 1$, i.e., change the original threshold $\bar{\gamma}_{c_0}$ to a new one $K\bar{\gamma}_{c_0}$, which are shown in Fig. 3. As a result, when the CT conducts the full-duplex relaying and reduces the amplitude gain from $G_{\max}$ with a reducing step ΔG, it can obtain the value G^* as follows

$$\begin{cases} G^* < G, & \text{if } \bar{\gamma}_{c_1} \le K\bar{\gamma}_{c_0}, \\ G^* = G, & \text{if } \bar{\gamma}_{c_1} > K\bar{\gamma}_{c_0}. \end{cases} \tag{11}$$

In this part, the interference probability to the PR is

$$P_{I_2} = \Pr\{\bar{\gamma}_{c_1} > \bar{\gamma}_{c_0}\}\Pr\{\mathbb{H}_0|\mathbb{H}_0\}\Pr\{\mathbb{H}_0\}. \tag{12}$$

From the above analysis, we find that the CT may cause interference to the PR in two situations. The first one may occur when the CT detects the case $g_1 > g_0$ and the corresponding interference probability is P_{I_1} in (9); The second one may occur when the CT obtains the value G^* and the corresponding interference probability is P_{I_2} in (12). Therefore, the overall interference probability becomes

$$\begin{aligned} P_I = P_{I_1} + P_{I_2} = \Pr\{\mathbb{H}_0|\mathbb{H}_1\}\Pr\{\mathbb{H}_1\} \\ + \Pr\{\bar{\gamma}_{c_1} > \bar{\gamma}_{c_0}\}\Pr\{\mathbb{H}_0|\mathbb{H}_0\}\Pr\{\mathbb{H}_0\}. \end{aligned} \tag{13}$$

Since the goal of this paper is to propose our idea and demonstrate its feasibility, we use a simplified method to choose the three parameters: First, we equally divide the overall interference probability, i.e., $P_{I_1} = P_{I_2} = 0.5P_I$. Then, we obtain the threshold η to meet the first interference probability P_{I_1}. Based on the obtained value η, we find the value of K to meet the second interference probability P_{I_2}, where the value of ΔG is set as a constant.

4. SIMULATION RESULTS

In this section, we adopt the same model as shown in Fig. 1, where the radius of the coverage is $R = 0.5$ km and the PT-CT distance is r km. In the simulation, the target SNR $\bar{\gamma}_T$ is uniformly distributed

[4]Due to the page limit, we omit the detail proof of this theorem.

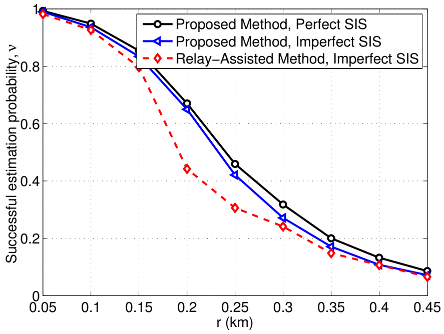

Fig. 4. The successful estimation probabilities of different methods.

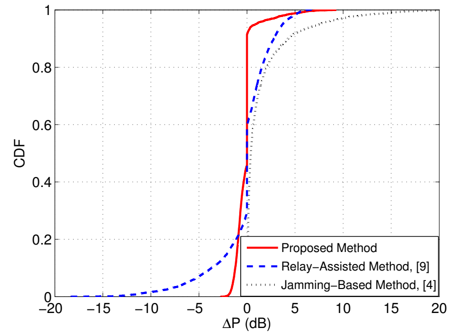

Fig. 6. Interference comparison of different methods.

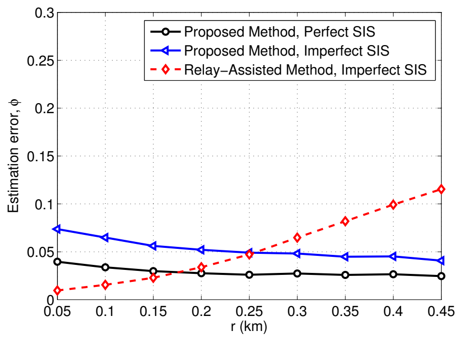

Fig. 5. The estimation errors of different methods.

between 5 dB and 30 dB, the maximum amplitude gain is $G_{\max} = 80$ dB, the reducing step of the amplitude gain is $\Delta G = 0.5$ dB, the interference probability is $P_I = 0.1$, the noise power is -114 dBm, the numbers of blocks and samples are $M = 200$ and $N = 100$, and the number of Monte Carlo trails is 10^3. For the wireless channels among the three nodes, we consider the path loss, shadowing, and small-scale fading. In particular, the shadowing coefficient follows log-normal distribution with standard deviation of 4. For the full-duplex AF relay, we also consider the impacts of the imperfect *self-interference suppression* (SIS). Here we raise the noise floor at the CT by 2 dB according to [17] and [18], in which the self-interference can be reduced to the noise level.

To evaluate the performance, we use both the successful estimation probability and the estimation error because the uncertainties of the wireless channel and noise may lead to the failure of the estimation, i.e., the estimator may output negative values. Therefore, we define ν as the successful estimation probability, i.e., $\nu = N_c/N_s$, where N_c is the number of the successful estimations and N_s is the number of Monte Carlo trails. For the number of successful estimations, we further define ϕ as the estimation error, i.e., $\phi = \mathbb{E}[|10\lg(\hat{g}_2) - 10\lg(g_2)|/10\lg(g_2)]$.

Fig. 4 compares the successful estimation probabilities of different methods, where the imperfect SIS is considered. From the figure, we observe that the successful estimation probabilities in all curves decrease as the PT-CT distance r increases. This is reasonable since both methods can only estimate the cross-channel gain under the $\mathbb{H}_0$ case, but the prior probability $\Pr\{\mathbb{H}_0\}$ decreases as the PT-CT distance r grows. Specifically, when we consider the proposed method, the imperfect SIS slightly degrades the performance by about 3% in average. When we compare the proposed method with the one in [9] (both consider the imperfect SIS), the former outperforms the latter in particular in the region when the PT-CT distance is 0.15 km $< r < 0.3$ km.

Fig. 5 compares the estimation error of different methods, where the imperfect SIS is also considered. From the figure, we observe that the estimation errors of both methods are between 0.01 and 0.11. Specifically, when we consider the proposed method, the estimation errors with perfect and imperfect SIS are about 0.05 and 0.06, and they slightly reduce as the PT-CT distance r increases. When we consider the relay-assisted method in [9], the estimation error grows from about 0.01 to about 0.11 as the PT-CT distance increases from 0.05 km to about 0.45 km. In average, both methods obtain the similar estimation errors.

Fig. 6 compares the interference to the PR caused by different methods, including the proposed method, the relay-assisted method in [9], and jamming-based method in [4], where all methods have the same estimation error and $r = 0.25$ km. To demonstrate the interference to the PR, we provide the *cumulative distribution function* (CDF) of the PT's power adjustment, i.e., $\Delta P = p_1/p_0$, where p_0 and p_1 represent the transmission powers of the PT before and during the relaying. If ΔP in dB unit is positive, it means that the CT causes interference to the PR and PT has to raise the power to compensate the SNR loss. If ΔP in dB unit is negative, it means that the CT does not cause interference to the PR. From the figure, the relay-assisted method and the conventional jamming-based method have about 35% and 100% probabilities to interfere with the PR. For the proposed method, it has only about 10% interference probability. This is because that the two existing methods do not consider the interference issue and our method is designed under the interference probability constraint.

5. CONCLUSIONS

In this paper, we proposed a non-cooperative relay-assisted estimation method to obtain the cross-channel gain. Since we converted the small TDOA to the large one, we can precisely control the interference to the PR. We found that (1) by measuring the primary signal before conducting the relaying, the cognitive user is able to detect the potential interference to the PR that is caused by the full-duplex AF relay; (2) by conducting the full-duplex AF relaying with a large amplitude gain, the CT can autonomously estimate the target SNR at the primary receiver, which is required by the proposed estimator; (3) by comparing the estimation performance, the proposed method has much lower interference probability than the existing ones.

6. REFERENCES

[1] S. Jayaweera, *Signal Processing for Cognitive Radios*, New York, NY, USA: Wiley, 2014.

[2] Z. Wang and W. Zhang, *Opportunistic Spectrum Sharing in Cognitive Radio Networks*, Berlin, Germany: Springer, 2015.

[3] B. Gopalakrishnan and N. Sidiropoulos, "Cognitive transmit beamforming from binary CSIT," *IEEE Trans. Wireless Commun.*, vol. 14, no. 2, pp. 895-906, Feb. 2015.

[4] R. Zhang, "On active learning and supervised transmission of spectrum sharing based cognitive radios by exploiting hidden primary radio feedback," *IEEE Trans. Commun.*, vol. 58, no. 10, pp. 2960-2970, Oct. 2010.

[5] I. Bajaj and Y. Gong, "Cross-channel estimation using supervised probing and sensing in cognitive radio networks," *in Proc. IEEE Int. Commun. Conf.* (ICC), Kyoto, Japan, Jun. 2011, pp. 1-5.

[6] Y. Noam and A. J. Goldsmith, "Blind null-space learning for MIMO underlay cognitive radio with primary user interference adaptation," *IEEE Trans. Wireless Commun.*, vol. 12, no. 4, pp. 1722-1734, Apr. 2013.

[7] Y. Noam , A. Manolakos and A. J. Goldsmith, "Null space learning with interference feedback for spatial division multiple access," *IEEE Trans. Wireless Commun.*, vol. 13, no. 10, pp. 5699-5715, Oct. 2014.

[8] F. Yuan, G. Villardi, F. Kojima, and H. Yano, "Channel direction information probing for multi-antenna cognitive radio system," in *EURASIP Journal on Advances in Signal Process.*, pp. 1-13, Aug. 2015. (Early Access: http://asp.eurasipjournals.com/content/2015/1/79)

[9] L. Zhang, G. Zhao, G. Wu, and Z. Chen, "Proactive channel gain estimation for coexistence between cognitive and primary users," in *Proc. IEEE GLOBECOM*, Anaheim, CA, USA, Dec. 2012, pp. 1410-1415.

[10] L. Zhang, G. Zhao, G. Wu, and Z. Chen, "Relay-assisted proactive channel gain estimation in cognitive radio," in *Proc. IEEE VTC-Fall*, Las Vegas, USA, Sept. 2013, pp. 1-5.

[11] L. Zhang, G. Zhao, G. Wu, and Z. Chen, "Cross-channel gain estimation with amplify-and-forward relaying in cognitive radio," in *Proc. IEEE GLOBECOM*, Atlanta, GA, USA, Dec. 2013, pp. 1003-1008.

[12] T. Riihonen and R. Wichman, "Power allocation for a single-frequency fixed-gain relay networks," in *Proc. IEEE Int. Symposium on Personal, Indoor and Mobile Radio Commun. (PIMRC)*, Athens, Greece, Sep. 2007, pp. 1-5.

[13] Y. Chen and P. Rapajic, "Decentralized wireless relay network channel modeling: an analogous approach to mobile radio channel characterization," *IEEE Trans. Commun.*, vol. 58, no. 2, pp. 467-473, Feb. 2010.

[14] H. Kim, S. Lim, H. Wang, and D. Hong, "Optimal power allocation and outage analysis for cognitive full duplex relay systems," *IEEE Trans. Wireless Commun.*, vol. 11, no. 10, pp. 3754-3765, Oct. 2012.

[15] 3GPP TR 25.814, "Physical layer aspects for evolved univeral terrestrial radio access (UTRA)," 2006.

[16] L. Li, G. Zhao and X. Zhou, "Enhancing small cell transmission opportunity through passive receiver detection in two-tier heterogeneous networks," *IEEE Trans. Signal Process.*, vol. 63, no. 13, pp. 3461-3473, Jul. 2015.

[17] T. Riihonen, S. Werner, and R. Wichman, "Mitigation of loopback self-interference in full-duplex MIMO relays," *IEEE Trans. Signal Process.*, vol. 59, no. 12, pp. 5983-5993, Dec. 2011.

[18] M. Jain, et al., "Practical, real-time, full duplex wireless," in *Proc. the 17th Ann. Int. Conf. on Mobile Radio Commun. (MobiCom)*, Las Vegas, Nevada, USA, Sep. 2011, pp. 1-12.