Circuit-Aware Cognitive Radios for Energy-Efficient Communications

Yunfei Chen, *Senior Member, IEEE,* Josef A. Nossek, *Fellow, IEEE,* Amine Mezghani, *Student Member, IEEE*

***Abstract*—In this letter, the tradeoff between sensing and throughput is studied for cognitive radios by considering the energy consumed in the circuit. The total energy is minimized with respect to the power fractions of the power amplifier and the low noise amplifier, the bit resolution of the analog-to-digital converter, the input backoff of the power amplifier and the sensing time in spectrum sensing, for a fixed achievable number of bits per packet. Numerical results show that the optimal sensing time increases when the circuit energy is considered. They also show that the analog-to-digital converter consumes more energy and that it is advantageous to operate the power amplifier at high input backoffs in cognitive radios.**

***Index Terms*—Achievable rate, circuit, cognitive radio, energy efficiency, spectrum sensing.**

I. Introduction

COGNITIVE radios (CR) have attracted great research interest in recent years. It is important for cognitive radios to determine how much time or energy should be allocated for spectrum sensing and data transmission given a fixed total time or energy. Several researchers have considered the tradeoff between sensing and throughput in the literature. For example, in [1], the tradeoff between spectrum sensing and data throughput was first studied. This tradeoff was reconsidered for wideband spectrum sensing where several frequency bands were sensed at the same time in [2] and [3]. In [4], by taking the Markovian traffic of the cognitive radio into account, more realistic tradeoff between sensing and throughput was derived. Finally, in [5], the sensing-throughput tradeoff was obtained by considering the primary user traffic. In all these works, the energy considered is the transmission energy from the transmitter to the receiver, while the energy consumed by the circuit is completely ignored. On the other hand, it was reported in several works that the energy consumed by the circuit is as significant as the transmission energy, if not more significant, especially in short-range applications [6], [7]. Based on this observation, in [8] and [9], different system parameters, including the power consumptions at the analog components, the bit resolution of the analog-to-digital converter (ADC) and the input backoff of the power amplifier (PA), were optimized. However, this optimization only works for traditional radios and it may not be optimal for cognitive radios that perform spectrum sensing before data transmission.

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Y. Chen is with the School of Engineering, University of Warwick, Coventry, U.K., CV4 7AL (e-mail: Yunfei.Chen@warwick.ac.uk).

J. A. Nossek and A. Mezghani are with the Institute for Circuit Theory and Signal Processing, Technische Universitaet Muenchen, Germany (e-mail: {josef.a.nossek, amine.mezghani}@tum.de).

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In this letter, energy-efficient cognitive radios are proposed to optimize the total energy in both spectrum sensing and data transmission with respect to the power consumptions at the analog components, the bit resolution of the ADC, the input backoff of the PA and the sensing time, considering the circuit energy. Both static additive white Gaussian noise (AWGN) channel and Rayleigh fading channel are considered. Numerical results are presented to show that the optimal sensing time increases when the circuit energy is considered.

II. System Model

We use a similar system model to that in [8], where only the most energy-consuming analog components of the circuit are considered as shown in Fig. 1. In this model, the power consumption at the PA can be written as [8, eq. (12)]

$$P_{PA} = \frac{1-e^{-z^2}}{\bar{\eta}}\sigma_x^2 \tag{1}$$

where z is the input backoff to be optimized later, $\bar{\eta} \approx \frac{\pi^{3/2}}{8z}\text{erf}(z) - \frac{\pi-4}{4}z^2[e^{-z^2} + \text{Ei}(-z^2)]$ is the average PA efficiency, σ_x^2 is the variance of the input to the PA, $\text{erf}(x) = \frac{2}{\sqrt{\pi}}\int_0^x e^{-t^2}dt$ is the Gaussian error function, and $\text{Ei}(x) = \int_{-\infty}^x e^t/tdt$ is the exponential integral. The power consumption at the low noise amplifier (LNA) is [8, eq. (13)]

$$P_{LNA} = \frac{G_{LNA}\cdot B\cdot N_0}{(N_F-1)FOM_{LNA}} \tag{2}$$

where G_{LNA} is the gain of the LNA, B is the operating bandwidth, N_0 is the single-sided power spectral density of the noise, N_F is the noise figure of the LNA, and FOM_{LNA} is a constant that depends on process technology and certain resistors. The power consumption at the ADC is [8, eq. (14)]

$$P_{ADC} = 2c_{ADC}\cdot B\cdot N_0\cdot 2^{2b} \tag{3}$$

where c_{ADC} is a constant depending on the ADC architecture, b is the bit resolution of the ADC to be optimized and other symbols are defined as before.

It was also reported in [8] that the achievable rate of such a system can be approximated as [8, eq. (17)]

$$R(SNR) \approx B\log_2\left(\frac{1+SNR}{1+SNR\cdot 2^{-2b}}\right) \tag{4}$$

where SNR is the signal-to-noise ratio given by

$$SNR = \frac{G_c\alpha^2\sigma_x^2}{G_c\sigma_{nPA}^2 + N_F\cdot B\cdot N_0} \tag{5}$$

with G_c being the channel gain, $\alpha = 1-e^{-z^2}+\frac{\sqrt{\pi}z}{2}\text{erfc}(z)$ being the amplitude of the signal component in the PA output, $\sigma_{nPA}^2 = (1-e^{-z^2}-\alpha^2)\sigma_x^2$ being the variance of the distortion component in the PA output, $\text{erfc}(x) = 1-\text{erf}(x)$ being the complementary error function, and all the other symbols being defined as before.

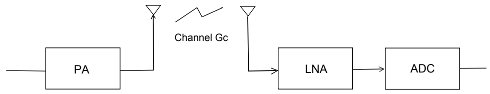

Fig. 1. Diagram of the signal path considered.

III. Derivation

In a cognitive radio system, the whole packet is divided into two parts, where the first part with a time duration of T_s is dedicated to spectrum sensing and the second part with a time duration of $T_t = T - T_s$ is dedicated to data transmission.

During T_s, cognitive radio only receives signal from the interested frequency band for spectrum sensing. Thus, the energy use in the first part can be written as

$$E_{ss} = (P_{LNA} + P_{ADC})T_s. \tag{6}$$

During T_t, cognitive radio sends information to another cognitive radio for data communications and thus, the energy use in the second part can be written as

$$E_{dt} = (P_{PA} + P_{LNA} + P_{ADC})T_t \tag{7}$$

where P_{PA} accounts for the power consumption at the transmitting cognitive radio and $P_{LNA} + P_{ADC}$ accounts for the power consumption at the receiving cognitive radio. Using (6) and (7), the total energy use per packet for cognitive radio is

$$E_t = E_{ss} + E_{dt} = P_{PA}(T - T_s) + (P_{LNA} + P_{ADC})T. \tag{8}$$

Next, we derive the objective functions for optimization.

Assume that $P(H_0)$ is the *a priori* probability that the licensed user is not active and $P(H_1) = 1 - P(H_0)$ is the *a priori* probability that the licensed user is active. Also, denote P_{fa} as the probability of false alarm, and P_d as the probability of detection.

Cognitive radio only transmits data when the interested frequency band is detected free. This has two scenarios. In the first scenario, the interested frequency band is detected free while the licensed user is actually absent. This happens with a probability of $P(H_0)(1 - P_{fa})$. In this scenario, the signal-to-noise ratio is the same as that in (5) and the achievable rate is the same as that in (4). In the second scenario, the interested frequency band is detected free while the licensed user is actually present. This happens with a probability of $P(H_1)(1 - P_d)$, where a miss-detection occurs. In this scenario, the signal-to-noise ratio needs to take the interference from the licensed user into account, giving

$$SNR_p = \frac{G_c \alpha^2 \sigma_x^2}{G_c \sigma_{nPA}^2 + N_F \cdot B \cdot N_0 + \sigma_p^2} \tag{9}$$

where σ_p^2 is the average interference power from the licensed user. The achievable rate in this scenario is therefore given by

$$R(SNR_p) \approx B \log_2 \left(\frac{1 + SNR_p}{1 + SNR_p \cdot 2^{-2b}} \right). \tag{10}$$

Putting these two scenarios together, the effective achievable number of bits per packet, obtained by multiplying the effective achievable rate with the packet duration, is

$$I = (T - T_s) \left[P(H_0)(1 - P_{fa}) \cdot B \cdot \log_2 \left(\frac{1 + SNR}{1 + \frac{SNR}{2^{2b}}} \right) + P(H_1)(1 - P_d) \cdot B \cdot \log_2 \left(\frac{1 + SNR_p}{1 + \frac{SNR_p}{2^{2b}}} \right) \right] \tag{11}$$

where the term $(T - T_s)$ takes the penalty of spectrum sensing on throughput into account. Using the technique in [8] by multiplying the nominator and denominator of (11) with E_t in (8) and minimizing the energy use of E_t for a fixed I, one has to maximize

$$f = \frac{T - T_s}{(P_{PA}(T - T_s) + (P_{LNA} + P_{ADC})T)} \cdot \left[P(H_0)(1 - P_{fa}) \cdot B \cdot \log_2 \left(\frac{1 + SNR}{1 + \frac{SNR}{2^{2b}}} \right) + P(H_1)(1 - P_d) \cdot B \cdot \log_2 \left(\frac{1 + SNR_p}{1 + \frac{SNR_p}{2^{2b}}} \right) \right]. \tag{12}$$

Further, denote $r_1 = \frac{P_{PA}}{P_{ADC}}$ and $r_2 = \frac{P_{PA}}{P_{ADC}}$. After some mathematical manipulations, one has

$$SNR = \frac{aG_c}{dG_c + c_1} \tag{13}$$

and

$$SNR_p = \frac{aG_c}{dG_c + c_2} \tag{14}$$

where $a = [1 - e^{-z^2} + \frac{\sqrt{\pi}}{2} z \cdot \text{erfc}(z)]^2 \eta r_1 / (1 - e^{-z^2})$, $d = [1 - e^{-z^2} - (1 - e^{-z^2} + \frac{\sqrt{\pi}}{2} z \cdot \text{erfc}(z))^2] \eta r_1 / (1 - e^{-z^2})$, $c_1 = \frac{1}{2c_{ADC}2^{2b}}(1 + \frac{G_{LNA}}{2r_2 c_{ADC} 2^{2b} FOM_{LNA}})$ and $c_2 = c_1 + \frac{\gamma_p}{2c_{ADC}2^{2b}}$. Essentially, SNR and SNR_p are functions of r_1, r_2, b and z.

It was also shown in [1] that, if energy detection is used for spectrum sensing based on the Neyman-Pearson rule, the probability of false alarm for fixed probability of detection is [1, eq. (13)]

$$P_f = Q(\sqrt{2\gamma_p + 1} Q^{-1}(\bar{P}_d) + \sqrt{T_s f_s} \gamma_p) \tag{15}$$

where $\gamma_p = \frac{\sigma_p^2}{N_0 B}$ is the signal-to-noise ratio of the licensed user signal, $\bar{P}_d$ is the fixed probability of detection, f_s is the sampling frequency with $f_s = B$ for Nyquist sampling, $Q(x) = \frac{1}{\sqrt{2\pi}} \int_x^\infty e^{-\frac{t^2}{2}} dt$ is the Gaussian Q function and $Q^{-1}(x)$ is the inverse function of $Q(x)$.

Using (13)- (15) in (12), one has

$$f = \frac{T - T_s}{2c_{ADC} \cdot N_0 (r_1(T - T_s) + (r_2 + 1)T) 2^{2b}} \cdot \Big[P(H_0)(1 - Q(\sqrt{2\gamma_p + 1} Q^{-1}(\bar{P}_d) + \sqrt{T_s f_s} \gamma_p)) \cdot \log_2 \left(\frac{(a + d)G_c + c_1}{(a \cdot 2^{-2b} + d)G_c + c_1} \right) + P(H_1) \left(1 - \bar{P}_d\right) \cdot \log_2 \left(\frac{(a + d)G_c + c_2}{(a \cdot 2^{-2b} + d)G_c + c_2} \right) \Big]. \tag{16}$$

For Rayleigh fading channels, one has G_c as an exponential random variable with probability density function of $p(G_c) = \frac{1}{2\sigma_G^2} e^{-\frac{G_c}{2\sigma_G^2}}$. The average achievable rate can be derived by

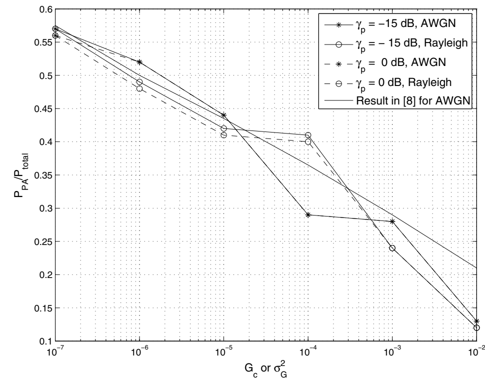

Fig. 2. $s_1 = P_{PA}/P_{total}$ vs. G_c or σ_G^2.

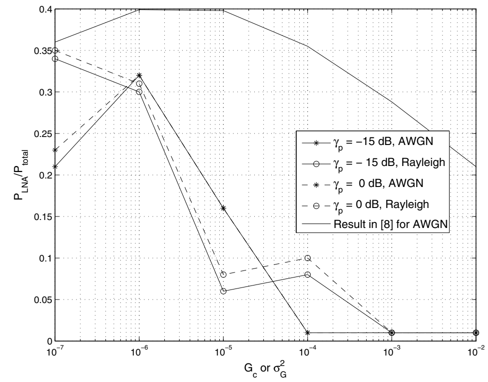

Fig. 3. $s_2 = P_{LNA}/P_{total}$ vs. G_c or σ_G^2.

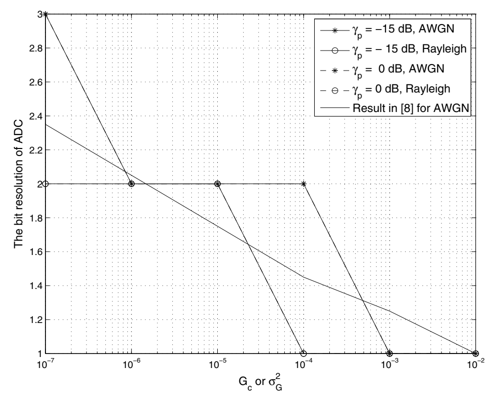

Fig. 4. b vs. G_c or σ_G^2.

integrating (4) and (10) over G_c. Using [10, eq. (4.337.2)] and after some manipulations, one has

$$\begin{aligned} f = &\frac{T - T_s}{2c_{ADC} \cdot N_0(r_1(T - T_s) + (r_2 + 1)T)2^{2b}} \\ &\cdot[P(H_0)(1 - Q(\sqrt{2\gamma_p + 1}Q^{-1}(\bar{P}_d) + \sqrt{T_s f_s \gamma_p})) \\ &\cdot[e^{\frac{c_1}{2\sigma_G^2(a\cdot 2^{-2b}+d)}} \mathrm{Ei}\left(\frac{-c_1}{2\sigma_G^2(a\cdot 2^{-2b}+d)}\right) \\ &- e^{\frac{c_1}{2\sigma_G^2(a+d)}} \mathrm{Ei}\left(\frac{-c_1}{2\sigma_G^2(a+d)}\right)] \\ &+ P(H_1)\left(1 - \bar{P}_d\right) \\ &\cdot[e^{\frac{c_2}{2\sigma_G^2(a\cdot 2^{-2b}+d)}} \mathrm{Ei}\left(\frac{-c_2}{2\sigma_G^2(a\cdot 2^{-2b}+d)}\right) \\ &- e^{\frac{c_2}{2\sigma_G^2(a+d)}} \mathrm{Ei}\left(\frac{-c_2}{2\sigma_G^2(a+d)}\right)]]. \end{aligned} \quad (17)$$

Thus, one needs to maximize the functions in (16) and (17) with respect to T_s, r_1, r_2, b and z for fixed values of other parameters. No closed-form expressions for the optimal values of T_s, r_1, r_2, b and z are tractable. In the next section, numerical examples will be presented.

IV. Numerical Results and Discussion

In this section, numerical examples are provided to show the optimal values of s_1, s_2, b, z and T_s that minimize the energy use in cognitive radio considering the circuit energy. The values of $s_1 = \frac{P_{PA}}{P_{total}} = \frac{r_1}{1+r_1+r_2}$ and $s_2 = \frac{P_{LNA}}{P_{total}} = \frac{r_2}{1+r_1+r_2}$ are used instead of r_1 and r_2 so that one can compare the result with that in [8], where $P_{total} = P_{PA} + P_{ADC} + P_{LNA}$. Also, other parameters are set to be the same as those in [1] and [8] for comparison such that $P(H_1) = 0.2$, $\bar{P}_d = 0.9$, $f_s = 6\ MHz$, $T = 100\ ms$ and $c_{ADC} = 96000$, $G_{LNA} = 10$, $FOM_{LNA} = 10^{-7}$.

Figs. 2 and 3 show the optimal values of s_1 and s_2 for different values of G_c in AWGN channels and σ_G^2 in Rayleigh fading channels when $\gamma_p = 0\ dB$ and $\gamma_p = -15\ dB$, respectively. Several observations can be made. First, for the same channel conditions, one sees that a larger value of γ_p leads to a lower value of s_1 and a higher value of s_2, suggesting that for stronger primary user signals, less transmission energy and more circuit energy are required, as P_{PA} is linear to the transmission energy [8, eq. (3)] while P_{LNA} determines the circuit energy. From (11) and (15), a larger γ_p leads to a larger value of $1 - P_{fa}$. To have a fixed achievable rate, SNR and SNR_p in (11) have to be reduced, leading to a smaller transmission energy. Thus, the proportion of the circuit energy increases. Second, for the same primary user signal, one sees that the channel condition does not change the optimal values of s_1 and s_2 much. For some ranges the values of s_1 and s_2 in AWGN channels are larger than those in fading channels, while for other ranges they are smaller. This is because G_c determines the AWGN channel condition and σ_G^2 determines the fading channel condition. When they are the same, the channel condition is roughly the same. Third, for large values of G_c or σ_G^2, the circuit energy dominates. For example, at $G_c = 10^{-2}$, the circuit energy consumes almost 90% of the total energy. This agrees with that observed in [8], as less transmission energy is required for short-range communications with higher channel gains such that the proportion of the circuit energy increases.

Fig. 4 shows the optimal value of the bit resolution b of the ADC vs. G_c or σ_G^2. One sees that the bit resolution decreases when G_c or σ_G^2 increase. This is because large values of G_c or σ_G^2 lead to better signal qualities such that the ambiguity caused by noise and fading is reduced and less bits are required to represent the signal. In most cases, a 1-bit or 2-bit resolution is enough for the ADC. For the same channel conditions, the optimal bit resolution does not change when γ_p changes. On the other hand, for the same primary user signal, the optimal

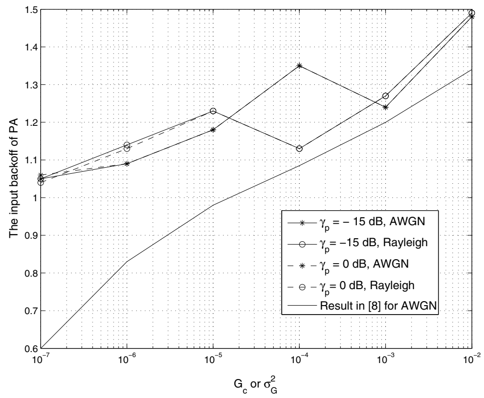

Fig. 5. z vs. G_c or σ_G^2.

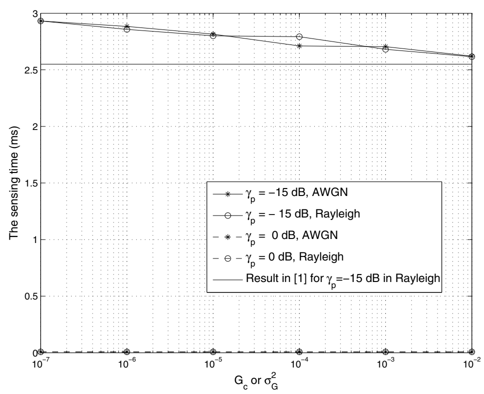

Fig. 6. T_s vs. G_c or σ_G^2.

bit resolution is smaller in Rayleigh fading channels than that in AWGN channels. Fig. 5 shows the optimal value of the input backoff z of the PA vs. G_c or σ_G^2. In this figure, the optimal value of z is almost identical for different channel conditions and different values of γ_p when G_c or σ_G^2 are large, while for small values of G_c and σ_G^2, the optimal value of z is larger for Rayleigh fading channels than for AWGN channels, as fading signals have larger dynamic ranges. Comparing our result with that in [8], one also sees that s_1 for CR is similar to that in [8], s_2 for CR is smaller than that in [8], b for CR is similar to that in [8] if it is quantitized, while z for CR is larger than that in [8]. This suggests that the ADC in cognitive radio consumes more energy than those in traditional radio, especially when G_c or σ_G^2 are small, and that it is advantageous to operate the PA at high input backoffs for cognitive radios.

Fig. 6 shows the optimal value of T_s vs. G_c or σ_G^2. In this case, the primary user signal is the dominant factor. For $\gamma_p = -15\ dB$, the optimal value of T_s decreases between 3 ms and 2.5 ms when G_c or σ_G^2 increase. For $\gamma_p = 0\ dB$, the optimal value of T_s is very small and is almost constant for different values of G_c or σ_G^2. This agrees with intuition that when the primary user signal is strong, less sensing time is required to detect it. In all cases, there is little difference between AWGN channels and Rayleigh fading channels. Comparing Fig. 6 with that in [1], one concludes that the optimal value of T_s that considers the circuit energy as in this letter is always larger. This suggests that the circuit energy has significant impact on the choice of the sensing time and that it cannot be ignored in the tradeoff between sensing accuracy and data throughput for cognitive radios as was in previous works. Note that the optimization is conducted by performing a multi-dimensional search over all the parameters in Figs. 2 - 6. Thus, in each figure, other searched parameters are not fixed and are not shown. Note also that the amount of power consumption is system-dependent. We use the systems considered in [1] and [8] so that one can compare our result with theirs to find out the difference between cognitive and traditional radios as well as the difference between cognitive radios with and without circuit energy consideration. However, our derivation in Section III is general enough to cover other systems, such as [11], too.

V. Conclusion

The tradeoff between sensing and throughput has been studied by considering the circuit energy using the method in [8]. Numerical results have shown that the optimal sensing time increases when the circuit energy of the cognitive radio is considered. They have also shown that it is advantageous to operate the PA at high input backoffs for cognitive radio.

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