A Spectral Analysis Based Channel Estimation Method for Time Diversity Combining in Helicopter Satellite Communications

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***Abstract*—Helicopter satellite communications have to overcome the periodic blockage of the signal caused by rotor blades. To solve this problem, time diversity with maximal ratio combining (MRC) is promising. Although the conventional channel estimation method to perform MRC is accurate, its computational complexity is relatively high because of fast Fourier transform (FFT) and inverse FFT (IFFT). This paper proposes a novel spectral analysis based channel estimation method. The proposed method estimates the channel gain without IFFT. The computer simulation results confirmed that the proposed method reduces the FFT size by 75% without degrading the BER performance compared with the conventional method using IFFT.**

I. Introduction

Helicopters play essential roles in news reports, rescue operations, and so on. They need to communicate with the base station in these missions. Although helicopter satellite communications are the only way to provide reliable communication when surrounding environments obstruct line-of-sight communication, they are required to overcome the periodic blockage (PB) of the signal due to rotor blades [1]–[7]. Time diversity is an attractive solution for low speed data transmission because of its simplicity [1]. To improve the bit error rate (BER) performance of the time diversity system, maximal ratio combining (MRC) is effective [3]. Performing MRC requires channel information.

In a previous study [5], an accurate channel estimation method for helicopter satellite communications was proposed. This conventional method extracts the line spectra of the PB from the received signal with fast Fourier transform (FFT) and reconstructs the PB as the channel information from the line spectra with inverse FFT (IFFT). The time diversity system employing the conventional method achieves the excellent BER performance very close to that of the ideal MRC using the known channel information even under a low carrier-to-noise power ratio (CNR) environment. However, the conventional method requires a large FFT/IFFT size, and therefore the computational complexity is relatively high.

In this paper, we propose a novel channel estimation method based on spectral analysis to reduce the computational complexity. The proposed method estimates the blockage ratio, defined as the ratio of the duration to the period of the PB, from the line spectra extracted with FFT and reconstructs the PB without IFFT by using the blockage ratio. Thus the proposed method eliminates IFFT and its computational complexity. Additionally, the results of the computer simulation showed that the proposed method reduces the FFT size by 75% without degrading the BER performance of the time diversity system. That is, the proposed method is significantly superior to the conventional method in terms of computational complexity.

The rest of the paper is organized as follows: Section II describes the periodic blockage and its line spectra. The communication system model is shown in Section III. Section IV explains the channel estimator and the conventional method. The proposed method is presented in Section V. Section VI reports and discusses the results of the computer simulation. Finally, Section VII concludes the paper.

II. Periodic blockage and its line spectra

Letting T_p and T_d denote the period and the duration of the PB, respectively, the following equation expresses the channel gain function $h(t)$ of the PB channel [1]:

$$h(t) = \sum_{\ell=-\infty}^{\infty} g(t - \ell T_p) \tag{1}$$

where

$$g(t) = \begin{cases} 1 & (T_d \leq t \leq T_p) \\ 0 & (\text{otherwise}) \end{cases}. \tag{2}$$

The period T_p is known and constant, while the duration T_d is unknown. Therefore, the blockage ratio defined as T_d/T_p is also unknown. The channel gain function $h(t)$ has line spectra because of its periodicity. The Fourier series expansion of $h(t)$ is as follows [5]:

$$h(t) = 1 - \rho + 2\rho \sum_{n=1}^{\infty} \frac{\sin n\pi\rho}{n\pi\rho} \cos(2\pi n f_0 t - n\pi\rho) \tag{3}$$

where ρ denotes the blockage ratio and $f_0 = 1/T_p$. From (3), we can see that the line spectra of $h(t)$ has the following envelope $\Gamma(f)$:

$$\Gamma(f) = 2\rho \frac{\sin(\pi T_d f)}{\pi T_d f} \quad (f > 0). \tag{4}$$

Letting the amplitude of the n-th line spectrum be denoted by A_n, we have

$$A_n = |\Gamma(n f_0)| = 2\rho \left| \frac{\sin n\pi\rho}{n\pi\rho} \right|. \tag{5}$$

Thus the amplitude A_n depends on the blockage ratio ρ.

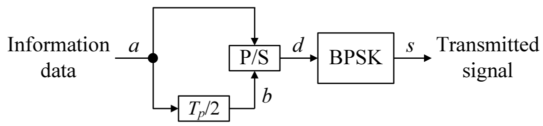

Fig. 1. Configuration of the transmitter.

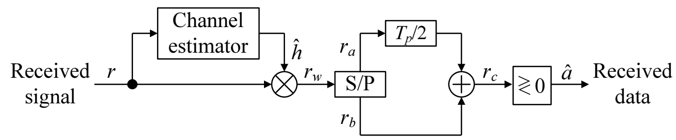

Fig. 2. Configuration of the receiver.

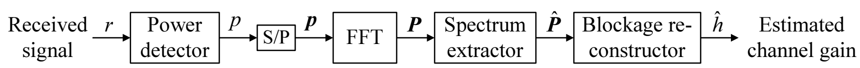

Fig. 3. Configuration of the channel estimator.

III. Communication system model

Figures 1 and 2 depict the configuration of the transmitter and the receiver of a time diversity system, respectively. The diversity factor, its value is two in Figs. 1 and 2, is represented by M. Let E_b and T_b denote the energy and the duration of the information bit, respectively.

A. Transmitter

In Fig. 1, the information bit sequence $\{a_i\}$ is delayed for T_p/M, resulting in the delayed sequence $\{b_i\}$. The parallel-to-serial converter (P/S) multiplexes the two sequences $\{a_i\}$ and $\{b_i\}$ into the transmitted data sequence $\{d_i\}$. The binary phase-shift keying (BPSK) modulator generates the transmitted signal s_i as follows:

$$s_i = \sqrt{\frac{E_b}{M}} \cos d_i \pi. \tag{6}$$

B. Channel

In the PB channel, the received signal r_i is described by the following equation.

$$r_i = h_i s_i + n_i \tag{7}$$

where $h_i = h(iT_b/M)$ is the channel gain and n_i denotes additive white Gaussian noise (AWGN) with the one-sided power spectral density N_0. Note that the value of the channel gain h_i is one or zero and thus ${h_i}^2 = h_i$.

C. Receiver

In Fig. 2, the channel estimator supplied the received signal r_i outputs the estimated channel gain $\hat{h}_i$. Multiplying the received signal r_i by the estimated channel gain $\hat{h}_i$ as the weighting factor yields the weighted signal $r_{w\,i}$. The serial-to-parallel converter (S/P) separates the weighted signal sequence $\{r_{w\,i}\}$ into the two signal sequences $\{r_{a\,i}\}$ and $\{r_{b\,i}\}$ corresponding to $\{a_i\}$ and $\{b_i\}$ in the transmitter, respectively. The signal sequence $\{r_{a\,i}\}$ is delayed for T_p/M and added to the signal sequence $\{r_{b\,i}\}$. This processing performs MRC and results in the combined signal sequence $\{r_{c\,i}\}$. The received data $\hat{a}_i$ is decided based on the value of the combined received signal $r_{c\,i}$.

IV. Channel estimator and conventional method

Figure 3 illustrates the configuration of the channel estimator based on spectrum analysis. In Fig. 3, the power detector detects the received signal power $p_i = |r_i|^2$. The S/P forms the following received power sequence $\boldsymbol{p}$:

$$\boldsymbol{p} = \begin{bmatrix} p_1 & p_2 & \cdots & p_N \end{bmatrix}^{\mathrm{T}} \tag{8}$$

where N denotes the FFT size. The FFT processor transforms the received power sequence $\boldsymbol{p}$ into the discrete Fourier transform (DFT) $\boldsymbol{P}$ represented by the following equation:

$$\boldsymbol{P} = \begin{bmatrix} P_1 & P_2 & \cdots & P_N \end{bmatrix}^{\mathrm{T}} = \boldsymbol{W}\boldsymbol{p} \tag{9}$$

where $\boldsymbol{W}$ is the DFT matrix of order N. From (6), (7), and the fact that ${h_i}^2 = h_i$, we have

$$p_i = r_i\, {r_i}^* = \frac{E_b}{M} h_i + 2 h_i s_i \,\mathrm{Re}(n_i) + |n_i|^2. \tag{10}$$

This equation clearly shows that the DFT $\boldsymbol{P}$ contains the line spectra of $h(t)$. The spectrum extractor selects the extracted line spectra $\hat{\boldsymbol{P}}$ from the DFT $\boldsymbol{P}$ as follows:

$$\hat{\boldsymbol{P}} = \begin{bmatrix} \hat{P}_1 & \hat{P}_2 & \cdots & \hat{P}_N \end{bmatrix}^{\mathrm{T}} \tag{11}$$

where

$$\hat{P}_k = \begin{cases} P_k & (|P_k| > P_{\mathrm{th}}) \\ 0 & (\text{otherwise}) \end{cases} \tag{12}$$

and P_{th} is the predetermined threshold. The blockage reconstructor produces the estimated channel gain $\hat{h}_i$ from the extracted line spectra $\hat{\boldsymbol{P}}$.

Figure 4 shows the configuration of the blockage reconstructor using the conventional method. The conventional method uses IFFT to reconstruct the PB from the extracted line spectra $\hat{\boldsymbol{P}}$. Let the IFFT of the the extracted line spectra $\hat{\boldsymbol{P}}$, i.e., the reconstructed PB, be represented as

$$\hat{\boldsymbol{p}} = \begin{bmatrix} \hat{p}_1 & \hat{p}_2 & \cdots & \hat{p}_N \end{bmatrix}^{\mathrm{T}}. \tag{13}$$

Binarizing $\hat{p}_i$ results in the estimated channel gain $\hat{h}_i$.

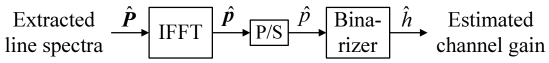

Fig. 4. Configuration of the conventional method.

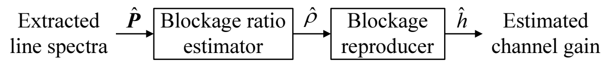

Fig. 5. Configuration of the proposed method.

V. Proposed method

The proposed method estimate the blockage ratio ρ from the amplitude of the extracted line spectra $\hat{\boldsymbol{P}}$ to reconstruct the PB without IFFT. Since the period T_p is known, estimating the blockage ratio ρ is nothing but estimating the duration T_d. Therefore, we can reconstruct the PB by (1) and (2).

As shown in (5), the amplitude of the line spectra of $h(t)$ only depends on the blockage ratio ρ. Equation (10) shows, however, that the amplitude of the the extracted line spectra $\hat{\boldsymbol{P}}$ depends not only on the blockage ratio ρ but also on the signal energy E_b. The following equation derived from (5) has no dependence on the signal energy:

$$\frac{A_{2n}}{A_n} = \frac{|\sin 2n\pi\rho|}{2|\sin n\pi\rho|} = |\cos n\pi\rho|. \tag{14}$$

Particularly, when $n\pi\rho < \pi/2$, we have

$$\frac{A_{2n}}{A_n} = \cos n\pi\rho. \tag{15}$$

Solving (15) for the blockage ratio ρ yields

$$\rho = \frac{1}{n\pi} \arccos \frac{A_{2n}}{A_n}. \tag{16}$$

Since A_n is the amplitude of the line spectrum at $f = nf_0 = n/T_p$, the prerequisite $n\pi\rho < \pi/2$ for (16) is equivalent to the following one:

$$f < \frac{1}{2T_d}. \tag{17}$$

The fact that the envelope $\Gamma(f)$ has the first zero at $f = 1/T_d$ assists in discriminating the range shown in (17).

Figure 5 shows the configuration of the proposed method. The blockage ratio estimator calculates the estimated blockage ratio $\hat{\rho}$ by the following procedure:

1) Let $\hat{A}_n$ be the amplitude of the element of the extracted line spectra $\hat{\boldsymbol{P}}$ corresponding to $f = nf_0$.
2) Find the following ordinal m:

$$m = \min\left(\arg\min_n \hat{A}_n\right). \tag{18}$$

Note that, from (12), the minimum value of $\hat{A}_n$ is zero. Equation (18) is therefore equivalent to searching the first zero of the envelope $\Gamma(f)$. In other words, the ordinal m has the relation $m\pi\rho = \pi$.
3) Let the ordinal L be $\lfloor m/2 \rfloor$. The ordinal $n \leq L$ satisfies the prerequisite $n\pi\rho < \pi/2$ for (16).
4) Calculate the estimated blockage ratio $\hat{\rho}$ by the following equation corresponding to (16):

$$\hat{\rho} = \frac{1}{L} \sum_{n=1}^{L} \frac{1}{n\pi} \arccos \frac{\hat{A}_{2n}}{\hat{A}_n}. \tag{19}$$

The blockage reproducer finds the lag τ maximizing the cross-correlation between the received power sequence $\boldsymbol{p}$ and the following $\hat{h}(t)$:

$$\hat{h}(t) = \sum_{\ell=0}^{\lfloor NT_b/2T_p \rfloor} \hat{g}(t - \ell T_p) \tag{20}$$

where

$$\hat{g}(t) = \begin{cases} 1 & (\hat{T}_d \leq t \leq T_p) \\ 0 & (\text{otherwise}) \end{cases} \tag{21}$$

and $\hat{T}_d = \hat{\rho} T_p$ is the estimated blockage duration. As $\hat{h}(t+\tau)$ is a replica of $h(t)$, the following equation gives the estimated channel gain $\hat{h}_i$:

$$\hat{h}_i = \hat{h}(iT_b/2 + \tau). \tag{22}$$

Clearly, the proposed method is simpler than the conventional one.

VI. Computer simulation

We evaluate the proposed method by computer simulation. Table I summarizes the simulation conditions. The typical and worst values of the blockage ratio ρ are 0.086 and 0.321, respectively. First, we evaluate the root mean square error (RMSE) of the estimated blockage ratio as the estimation accuracy. Second, we compare the BER performance of the proposed method with that of the conventional one and both the theoretical curves of equal gain combining (EGC) and MRC. Letting P_{EGC} and P_{MRC} denote the theoretical BER of EGC and MRC, respectively, we can derive them as follows:

$$P_{\text{EGC}} = \frac{1}{2}\left[2\rho \operatorname{erfc}\sqrt{\frac{E_b}{4N_0}} + (1-2\rho)\operatorname{erfc}\sqrt{\frac{E_b}{N_0}}\right]$$
$$P_{\text{MRC}} = \frac{1}{2}\left[2\rho \operatorname{erfc}\sqrt{\frac{E_b}{2N_0}} + (1-2\rho)\operatorname{erfc}\sqrt{\frac{E_b}{N_0}}\right].$$

A. RMSE of estimated blockage ratio

The accuracy of the estimated blockage ratio $\hat{\rho}$ depends on the FFT size N. Figure 6 shows the RMSE of the estimated blockage ratio $\hat{\rho}$ vs. the FFT size N characteristics with $\rho = 0.086$ at $E_b/N_0 = 0$ dB. We can see that the RMSE decreases with increasing the FFT size N. However, it does not pay to

TABLE I
Simulation conditions

Modulation	BPSK
Diversity factor	2
Information data rate	4096 bps
Blockage period	31.25 msec

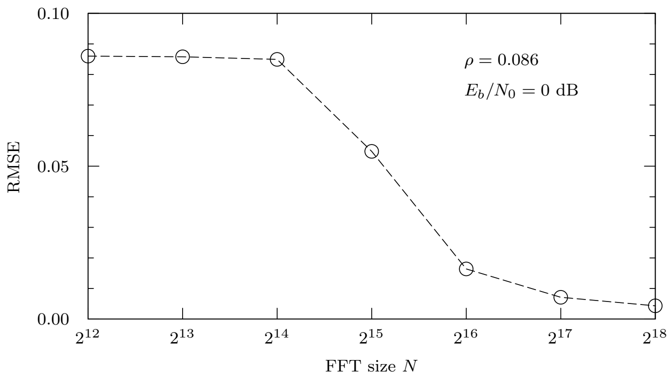

Fig. 6. RMSE of the estimated blockage ratio vs. FFT size characteristics.

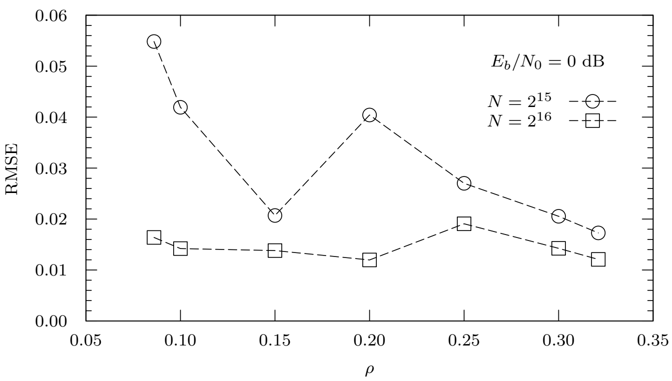

Fig. 7. RMSE vs. blockage ratio characteristics.

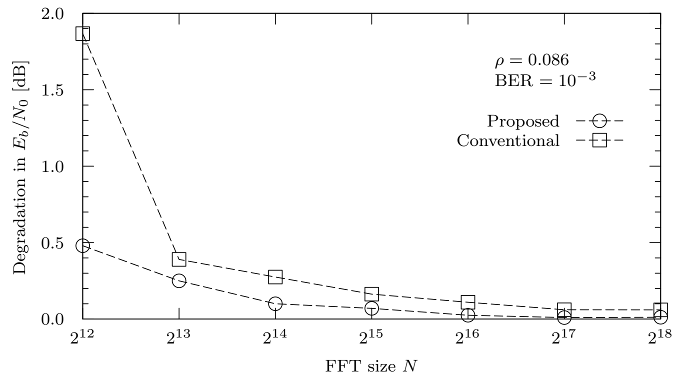

Fig. 8. Dependence of BER performance degradation on FFT size.

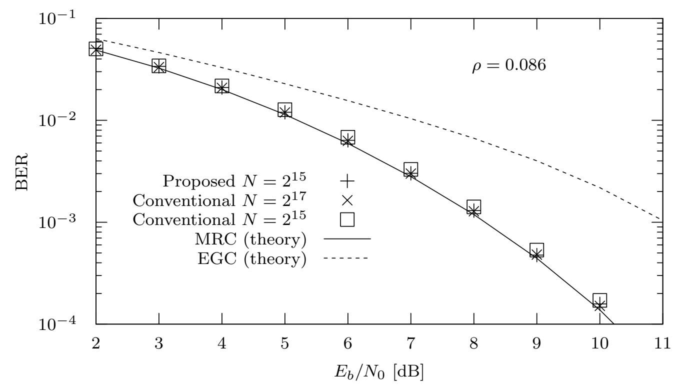

Fig. 9. BER performance.

increase the FFT size N to 2^{17} or more because decreasing the RMSE becomes smaller.

Figure 7 shows the RMSE from the typical blockage ratio $\rho = 0.086$ to the worst one $\rho = 0.321$. The RMSE with the FFT size $N = 2^{16}$ is independent of ρ and almost constant. Although halving the FFT size approximately doubles the RMSE on average, it reduces the computational complexity.

B. BER performance

Figure 8 shows the dependence of the degradation from the theoretical curve of MRC on the FFT size N at BER $= 10^{-3}$. The FFT sizes reducing the degradation to less than 0.1 dB are 2^{15} in the proposed method and 2^{17} in the conventional one. That is, the proposed one reduces the FFT size by 75%.

Figure 9 shows the BER performance. The proposed method with the FFT size $N = 2^{15}$ achieves the BER performance very close to the theoretical curve of MRC. This demonstrates that the proposed method almost perfectly estimates the channel gain. The proposed method outperforms the conventional one with the same FFT size. To achieve the same performance as the proposed one, the conventional one requires quadrupling the FFT size. Thus, the proposed method does not only employ no IFFT but also reduces the FFT size significantly. Therefore, the computational complexity of the proposed one is remarkably low compared with the conventional one.

VII. Conclusion

We have proposed a spectral analysis based channel estimation method for time diversity combining in helicopter satellite communications. The proposed method estimates the channel gain without IFFT by using the amplitude of the line spectra of the PB. The computer simulation results confirmed that the proposed method significantly reduces the computational complexity without degrading the BER performance.

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