A Coarse Frequency Offset Estimation Based on the Differential Correlation in DAB Systems

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Abstract

This paper presents a new and robust technique for a coarse frequency offset estimation in OFDM systems. As an evaluation of the proposed algorithm, we apply it to Eureka 147 DAB system. The proposed coarse frequency offset estimation algorithm is based on the differential detection technique between adjacent subcarriers to eliminate the phase shift effects of symbol timing offset and fractional frequency offset. A coarse frequency offset is determined from the correlation outputs between a received intercarrier differential phase reference symbol and several locally generated but frequency-shifted intercarrier differential phase reference symbols. The performance of our estimation algorithm is evaluated by means of computer simulation and is compared with those of previous proposed algorithms for DAB transmission modes Ⅰ, Ⅱ, Ⅲ, and Ⅳ. Simulation results show that the proposed algorithm generates extremely accurate estimates with low complexity irrespective of the symbol timing offset.

Ⅰ. INTRODUCTION

Digital audio broadcasting (DAB) system developed by the Eureka 147 project uses orthogonal frequency division multiplexing (OFDM) modulation to alleviate the effect of multipath fading[1]~[3]. However, OFDM modulation scheme is known as being sensitive to carrier frequency offset due to oscillator instability or Doppler shift[4]~[6].

A number of coarse frequency offset estimation methods have been proposed for OFDM applications[2],[6]~[8]. A simplified coarse frequency offset estimation method is proposed by Norbert in [7]. When a symbol timing offset exists, however, this method cannot achieve satisfactory performance. In [2],[7],[8], an estimation method using a channel impulse response (CIR) is proposed, which is widely used. This method can exactly estimate the coarse frequency offset irrespective of the symbol timing offset and channel environment. However, too much computation is required in this method.

In this paper, we propose a new and robust algorithm to estimate coarse frequency offset with low complexity irrespective of the symbol timing offset. The proposed scheme is based on the differential detection technique between subcarriers to eliminate the phase shift effects of symbol timing offset and fractional frequency offset. In order to demonstrate the validity of the proposed synchronization algorithm, we apply it to DAB system.

In Section 2, we first discuss the DAB system model and the existing coarse frequency offset estimation algorithms. In Section 3, the proposed coarse frequency offset estimation algorithm is presented in detail. The performance of the proposed algorithm is evaluated by computer simulation and the simulation results are presented in Section 4. Finally, we give a brief conclusion in Section 5.

Ⅱ. DAB SYSTEM MODEL AND EXISTING COARSE FREQUENCY OFFSET SYNCHRONIZATION ALGORITHMS

2-1 DAB System Model

The DAB broadcast signal is arranged into a transmission frame as shown in Fig. 1, which consists of consecutive OFDM symbols[1],[8]. Each OFDM symbol consists of a set of equally-spaced subcarriers with a subcarrier spacing equal to $1/T_u$, where T_u is a useful symbol duration. Symbol duration is the sum of useful symbol duration T_u and guard interval duration Δ. The number of OFDM symbols, the number of subcarriers and other various parameters are dependent on the transmission mode, as shown in Table 1.

The first OFDM symbol of a transmission frame is the null symbol, which is a duration of no signal transmitted. It is used to perform coarse time (frame) synchronization by envelope detection of the signal[6]. The second OFDM symbol is the phase reference symbol (PRS), which has fixed magnitudes and known phases on each subcarrier[2],[6],[8]. It gives the precise reference for symbol timing and frequency offset synchroniza-

Manuscript received April 18, 2001; revised May 17, 2001.

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Table 1. Parameters for tranmission modes Ⅰ, Ⅱ, Ⅲ, Ⅳ

Parameters		DAB transmission mode			
		Ⅰ	Ⅱ	Ⅲ	Ⅳ
L	No. of symbols/frame	76	76	153	76
K	No. of subcarriers/symbol	1536	384	192	768
Δ	No. of samples/guard interval	504	126	63	252
N	No. of IFFT samples/symbol	2048	512	256	1024

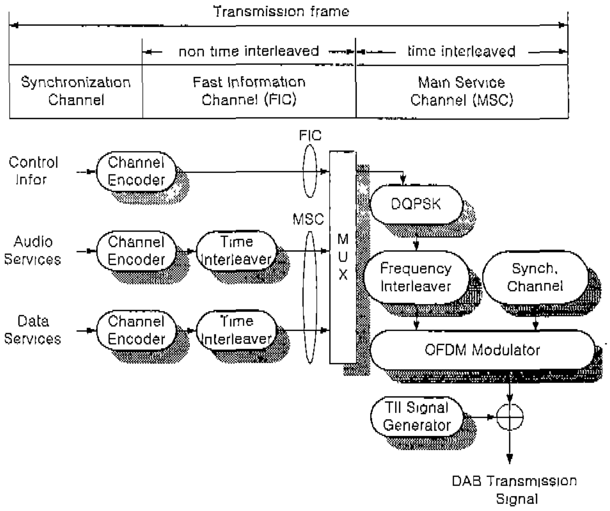

Fig. 1. Transmission frame structure and block diagram for DAB.

tion since the contents of this symbol are standardized and fixed. These two symbols comprise the synchronization channel (SC). The main service channel (MSC) is used to transmit each service, which is individually encoded and time-interleaved. The fast information channel (FIC) is used to transmit the control data which is necessary for demultiplexing and decoding of the MSC part in each transmission frame. A Eureka 147 OFDM symbol consists of N sinusoidal subcarriers with frequency spacing $1/T_u$, and each subcarrier is differential quadrature phase shift keying (DQPSK) modulated on a symbol-by-symbol basis to avoid channel estimation, which is based on simple and inexpensive receiver for consumer[1],[3],[6]. During a useful symbol duration T_u, the complex envelope of a transmitted OFDM symbol can be expressed as

$$x(n) = \frac{1}{\sqrt{N}} \sum_{\substack{k=-K/2 \\ k\neq 0}}^{K/2} X(k) e^{j2\pi nk/N}, \quad n=0,1,\cdots,N-1$$
$$= \frac{1}{\sqrt{N}} \sum_{k=0}^{N-1} X(k) e^{j2\pi nk/N} \tag{1}$$

where $X(k)$ represents the complex signal in the k-th subcarrier at the PRS or the frequency interleaved complex DQPSK signal in the k-th subcarrier at the FIC and MSC, and N is the number of subcarriers. N is larger than K, and it is typically chosen as a power of two.

After passing through a bandpass channel, the complex envelope of the received symbol can be expressed as

$$y(n) = \frac{1}{\sqrt{N}} \left[\sum_{k=0}^{N-1} X(k) H(k) e^{j2\pi n(k+\Delta f)/N} \right] + w(n) \tag{2}$$

where $H(k)$ is the transfer function of the channel at frequency of the k-th subcarrier and Δf is the relative frequency offset (the ratio of the actual frequency offset to the intercarrier spacing). Frequency offset Δf is divided into integer part Δf_i and fractional part Δf_f. $w(n)$ is the complex envelope of a additive white Gaussian noise (AWGN).

At the receiver, the demodulator performs a fast Fourier transform (FFT) in order to obtain the multiple OFDM subcarriers. In an ideal case, the received and demodulated symbol $Y(k)$ will be

$$Y(k) = \frac{1}{\sqrt{N}} \sum_{n=0}^{N-1} y(n) e^{-j2\pi kn/N} \tag{3}$$

where $y(n)$ is the received OFDM symbol. But if there is a symbol timing offset τ, the operation of the FFT within an OFDM symbol period including a guard interval Δ is defined by

$$Y(k) = \frac{1}{\sqrt{N}} \sum_{n=0}^{N-1} y(n) e^{-j2\pi k(n-\tau)/N}$$
$$= \frac{e^{j2\pi k\tau/N}}{\sqrt{N}} \sum_{n=0}^{N-1} y(n) e^{-j2\pi kn/N} \tag{4}$$

where k is sample index of current symbol and τ is symbol timing offset in samples. The k-th subcarrier of the FFT can be expressed as

$$Y(k) = H(k) X(k-\Delta f_i) e^{j2\pi(k-\Delta f_i)\tau/N} \cdot e^{j\pi \Delta f_f \frac{N-1}{N}}$$
$$\cdot \frac{1}{N} \frac{\sin(\Delta f_f \pi)}{\sin(\Delta f_f \pi/N)} + I(k) + W(k) \tag{5}$$

An integer frequency offset Δf_i will only lead to a shift of all subcarriers while the signal orthogonality remains. Because of the fractional frequency offset Δf_f, the magnitude of the desired signal is attenuated by both $H(k)$ and $\sin(\Delta f_f \pi)/\sin(\Delta f_f \pi/N)$. Besides, the term $\exp\{j\pi\Delta f_f(N-1)/N\}$ introduces a constant phase shift of each of the subcarriers. If $\Delta f_f \neq 0$, the fractional frequency offset causes intercarrier interference (ICI) between subcarriers. The second term $I(k)$ in eqn. (5) is the ICI that is the sum of components dependent on each of values $X(l)$, $l \neq k$. The contribution of each $X(l)$ can be approximated as a Gaussian noise according to the central limit theorem if more subcarriers are allowed [9]. The effect of symbol timing offset will produce a linear phase shift in the

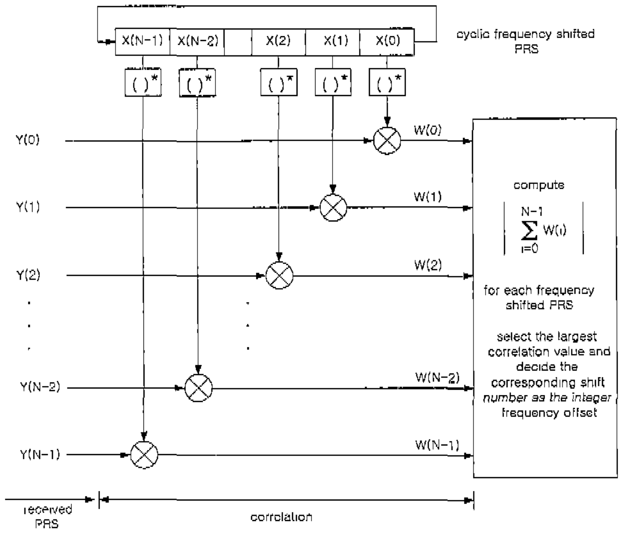

Fig. 2. Configuration of the frequency offset estimator using APRS.

frequency domain, which is proportional to the multiplication of symbol timing offset τ and subcarrier index k.

2-2 The Existing Coarse frequency Offset Synchronization Algorithms

2-2-1 The Algorithm using the Phase Reference Symbol (APRS)

If there is an integer offset in the subcarrier positions of the demodulated phase reference symbol, this technique performs the correlation between a sampled reference symbol and the frequency-shifted original reference symbol, as shown in Fig. 2. The amount of shift which provides the maximum correlation is considered as an integer carrier offset such as

$$\widehat{\Delta f_i} = \max_{d} \left| \sum_{k=0}^{N-1} Y(k) \cdot X^*(k+d)_N \right| \tag{6}$$

where d is the amount of cyclic shift and $(k+d)_N$ is the modulo-N addition operation. This algorithm is more simple than other algorithms, but when a symbol timing offset exists, satisfactory performance cannot be achieved.

2-2-2 The Algorithm using the Channel Impulse Response (ACIR)

The configuration of this algorithm is shown in Fig. 3. This algorithm searches for the highest peak in the IFFT output to find the integer carrier offset. The amount of shift which gives the highest peak can be interpreted as the carrier frequency offset such as

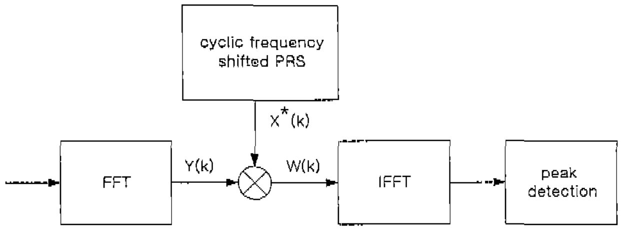

Fig. 3. Configuration of the frequency offset estimator using ACIR.

$$\widehat{\Delta f_i} = \max_{d} \left\{ \max_{amp} IFFT \{ Y(k) \cdot X^*(k+d)_N \} \right\} \tag{7}$$

This algorithm can accurately estimate the integer frequency offset irrespective of the symbol timing offset and channel environment, which requires too much computation.

Ⅲ. A NEW COARSE FREQUENCY OFFSET ESTIMATION ALGORITHM

3-1 Proposed Coarse Frequency Offset Estimation Algorithm using the Intercarrier Differential Correlation (AIDC)

From eqn. (5), the amount of phase rotation of $Y(k)$ is proportional to the fractional frequency offset Δf_f and the product of symbol timing offset τ and subcarrier index k. By using the differential detection technique between adjacent subcarriers, the effect of phase rotation caused by fractional frequency offset Δf_f and subcarrier index k can be neglected. The output of the k-th intercarrier differential detector can be expressed as

$$\begin{aligned} \underline{Y}(k) &= Y(k) \cdot Y^*(k-1) \\ &\approx X(k-\Delta f_i) X^*(k-1-\Delta f_i) e^{j2\pi\tau/N} \left[\frac{1}{N} \frac{\sin(\Delta f_f \pi)}{\sin(\Delta f_f \pi / N)} \right]^2 + \eta(k) \end{aligned} \tag{8}$$

where underbar means intercarrier differential detection between the k-th and the $(k-1)$-th subcarriers and η is a noise component. The last equality holds if the channel responses $H(k)$ and $H(k-1)$ is quasistationary during two adjacent subcarriers, and if the noise component is sufficiently small. Futhermore, the quasistationary assumption can be valid thanks to the slowly-varying frequency selective channl property and the effect of channel response can be ignored using the differential detection. In the absence of noise, eqn. (8) can be expressed as

$$\underline{Y}(k) = \underline{X}(k-\Delta f_i) e^{j2\pi\tau/N} \left[\frac{1}{N} \frac{\sin(\Delta f_f \pi)}{\sin(\Delta f_f \pi / N)} \right]^2 \tag{9}$$

where $\underline{X}(k-\Delta f_i)$ is a intercarrier differential phase reference symbol shifted by the amount of integer offset. From eqn. (9), the amount of phase rotation of $\underline{Y}(k)$ is only related to the term $\exp\{j2\pi\tau/N\}$ which is independent on the index k, i.e., that term is a constant.

To estimate the coarse frequency offset (integer offset), the correlation between $\underline{Y}(k)$ and the cyclic shifted intercarrier differential symbol $\underline{X}(k)=X(k)X^*(k-1)$ which is calculated from the original phase reference symbol is computed and the shift value that provides maximum correlation is decided as an estimated coarse frequency offset such as

$$\widehat{\Delta f}_i = \max_{d} \left| \sum_{k=0}^{N-1} \underline{Y}(k) \cdot \underline{X}^*(k+d)_N \right| \tag{10}$$

If the shift value of d should be the same as the integer frequency offset Δf_i, in other words, $d=\Delta f_i$, then eqn. (10) becomes

$$\begin{aligned}
&\left| \sum_{k=0}^{N-1} \underline{X}(k-\Delta f_i) e^{j2\pi\tau/N} \left[\frac{1}{N} \frac{\sin(\Delta f_f \pi)}{\sin(\Delta f_f \pi/N)} \right]^2 \underline{X}^*(k-d)_N \right|_{d=\Delta f_i} \\
&= \left| \sum_{k=0}^{N-1} \underline{X}(k-\Delta f_i) e^{j2\pi\tau/N} \left[\frac{1}{N} \frac{\sin(\Delta f_f \pi)}{\sin(\Delta f_f \pi/N)} \right]^2 \underline{X}^*(k-\Delta f_i) \right| \\
&= \left| \sum_{k=0}^{N-1} X(k-\Delta f_i) X^*(k-1-\Delta f_i) e^{j2\pi\tau/N} \right. \\
&\quad \left. \left[\frac{1}{N} \frac{\sin(\Delta f_f \pi)}{\sin(\Delta f_f \pi/N)} \right]^2 X^*(k-\Delta f_i) X(k-1-\Delta f_i) \right| \\
&= \left| \sum_{k=0}^{N-1} |X(k-\Delta f_i)|^2 \cdot |X(k-1-\Delta f_i)|^2 \cdot \right. \\
&\quad \left. e^{j2\pi\tau/N} \left[\frac{1}{N} \frac{\sin(\Delta f_f \pi)}{\sin(\Delta f_f \pi/N)} \right]^2 \right| \\
&= \left[\frac{1}{N} \frac{\sin(\Delta f_f \pi)}{\sin(\Delta f_f \pi/N)} \right]^2 \cdot |e^{j2\pi\tau/N}| \\
&\quad \cdot \left| \sum_{k=0}^{N-1} |X(k-\Delta f_i)|^2 \cdot |X(k-1-\Delta f_i)|^2 \right|
\end{aligned} \tag{11}$$

The reference symbol in the DAB contains a constant amplitude, in other words, $|X(k)|=1$. Therefore,

$$\begin{aligned}
&\left| \sum_{k=0}^{N-1} \underline{X}(k-\Delta f_i) e^{j2\pi\tau/N} \left[\frac{1}{N} \frac{\sin(\Delta f_f \pi)}{\sin(\Delta f_f \pi/N)} \right]^2 \underline{X}^*(k-d)_N \right|_{d=\Delta f_i} \\
&= \left[\frac{1}{N} \frac{\sin(\Delta f_f \pi)}{\sin(\Delta f_f \pi/N)} \right]^2 \cdot N \\
&= \frac{1}{N} \frac{\sin^2(\Delta f_f \pi)}{\sin^2(\Delta f_f \pi/N)}
\end{aligned} \tag{12}$$

If the shift value of d is not the integer frequency offset Δf_i, in other words, $d=\Delta f_j \neq \Delta f_i$, then eqn. (10) becomes

$$\begin{aligned}
&= \left[\frac{1}{N} \frac{\sin(\Delta f_f \pi)}{\sin(\Delta f_f \pi/N)} \right]^2 \\
&\quad \cdot \left| \sum_{k=0}^{N-1} X(k-\Delta f_i) X^*(k-\Delta f_j) X(k-1-\Delta f_j) X^*(k-1-\Delta f_i) \right| \\
&< \frac{1}{N} \frac{\sin^2(\Delta f_f \pi)}{\sin^2(\Delta f_f \pi/N)}
\end{aligned} \tag{13}$$

Since the phase reference symbol (PRS) in Eureka 147 has the property that it is orthogonal to itself with any integer frequency offset, the correlation value of $d=\Delta f_j \neq \Delta f_i$ is always less than that of $d=\Delta f_i$. So, the proposed coarse frequency offset estimation algorithm has the maximum correlation value at the integer frequency offset irrespective of the symbol timing offset because the term $e^{j2\pi\tau/N}$ is eliminated, that is, $|e^{j2\pi\tau/N}|=1$. In other words, there is no problem in estimating the coarse frequency offset provided that the time reference begins in the guard interval. The configuration of the proposed frequency offset estimator based on eqn. (10) is shown in Fig. 4.

3-2 Comparisons of Complexity

In this paper, we define the computational complexity as the number of the complex multiplication. The existing algorithm using the phase reference symbol has a disadvantage that it can not be applied when the symbol timing offset exists. So, the algorithm using the channel impulse response is widely used because it is insensitive to a symbol timing offset. But the algorithm using the channel impulse response has a disadvantage that it requires too much computation as shown in Table 2. The complexity of the proposed algorithm is nearly proportional to N^2, that is the same complexity as the algorithm using the phase reference symbol. On the other hand, the complexity of the algorithm using the channel impulse response is proportional to $N\times(N+(N/2)\log_2 N)$. So, the proposed algorithm is inse-

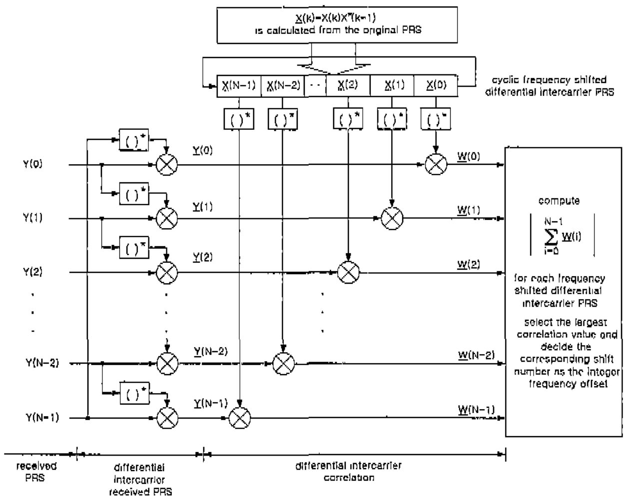

Fig. 4. Configuration of the proposed frequency offset estimator.

Table 2. Comparisons of complexity of coarse frequency offset algorithms

Algorithms	Relation to timing offset	Complexity (N : the number of subcarriers)
APRS	dependent	N^2
ACIR	independent	$N\times(N+(N/2)\log_2 N)$
AIDC	independent	$N^2+N \approx N^2$

nsitive to a symbol timing offset with reduced complexity.

Ⅳ. SIMULATION RESULTS

In order to demonstrate the effectiveness of the proposed synchronization algorithm, the effect of timing offset on the frequency offset estimation is simulated. The performance of our estimation algorithm is evaluated by means of computer simulation and compared with that of previous works for DAB transmission modes Ⅰ, Ⅱ, Ⅲ, and Ⅳ. Table 1 is used as parameters for DAB transmission modes Ⅰ, Ⅱ, Ⅲ and Ⅳ. Computer simulations are performed in a frequency selective Rayleigh fading channel, which is implemented by using channel parameters specified in European Telecommunications Standard Institute (ETSI)[10]. The multipath delay profile is shown in Fig. 5.

In this simulation, the carrier frequency offset and the symbol timing offset are set to 2.4 subcarrier spacings and 10 samples. So, the exact integer frequency offset of 2 have to estimated and the estimation error is obtained by running 1000 frames at each SNR value. Fig. 6 shows the integer frequency offset estimation error for DAB transmission mode I, as a function of SNR. As mentioned in Section 2, we see that the APRS does not obtain the integer frequency offset estimation accurately, as long as the symbol timing offset exists. It can also be observed that the proposed algorithm can perfectly estimate the coarse frequency offset such as the ACIR only if the SNR is greater than 2[dB]. Fig. 7, 8 and 9 show a plot of the integer frequency offset

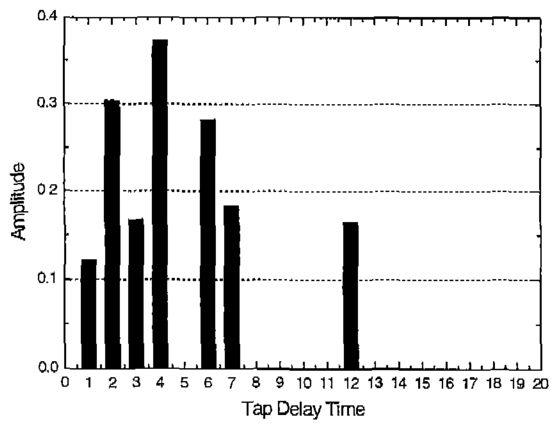

Fig. 5. Multipath delay profile specified in ETSI.

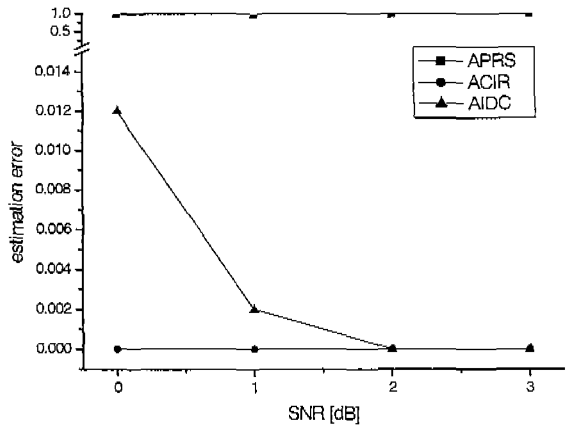

Fig. 6. Estimation error versus SNR for Mode Ⅰ.

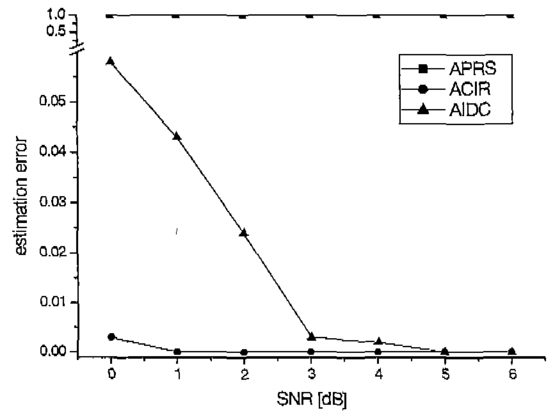

Fig. 7. Estimation error versus SNR for Mode Ⅱ.

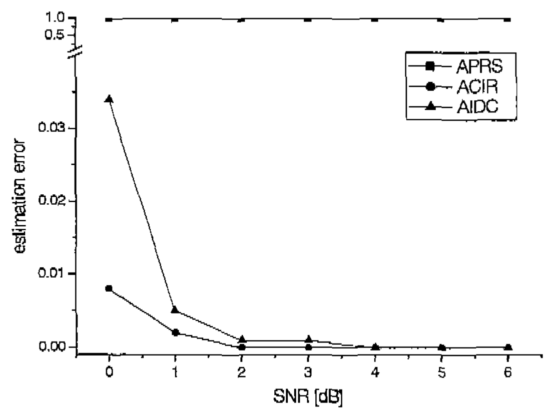

Fig. 8. Estimation error versus SNR for Mode Ⅲ.

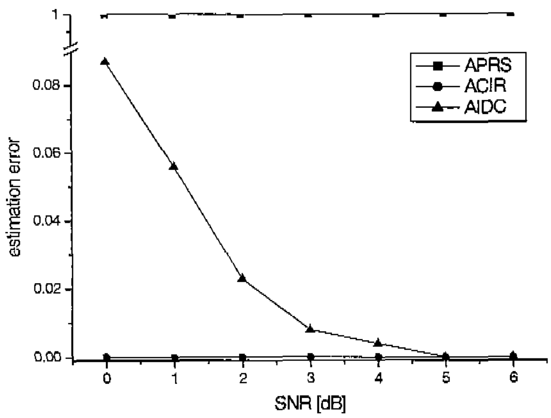

Fig. 9. Estimation error versus SNR for Mode Ⅳ.

estimation error versus SNR for DAB transmission mode Ⅱ, Ⅲ and Ⅳ, respectively. We know that the APRS can not estimate the accurate frequency offset for each transmission mode if the timing offset should exists. Simulation results show that the estimation error performance of our algorithm coincides with the ACIR case when the SNR is greater than 5 [dB]. In other words, the proposed algorithm and the ACIR can exactly estimate the coarse frequency offset (perfect coarse frequency offset estimation, 0 estimation error) irrespective of the symbol timing offset.

It can be observed that the proposed algorithm outperforms the APRS introduced in [7] against the accuracy of frequency offset with the nearly same computational complexity. And the proposed algorithm shows the same accuracy as the ACIR proposed in [8] in other words, each algorithm is insensitive to a symbol timing offset, but the complexity is reduced by 80～85%. Therefore, the proposed algorithm can be used effectively for each DAB transmission mode with low computational complexity.

V. CONCLUSIONS

In this paper, we have described a technique for the robust frequency offset estimation for OFDM system. To get rid of the phase effects of symbol timing offset and fractional frequency offset, the algorithm using intercarrier differential phase is proposed. The proposed synchronization algorithm estimates an accurate frequency offset with low complexity irrespective of symbol timing offset by using intercarrier differential phase correlation. We have observed that by applying the proposed algorithm to DAB system the proposed algorithm can accurately estimate the coarse frequency offset much faster than the existing algorithms because of the reduced computation.

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