A LOW-COMPLEXITY ANTENNA DIVERSITY RECEIVER SUITABLE FOR TDMA HANDSET IMPLEMENTATION

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Abstract— **This paper introduces a low-complexity antenna diversity receiver suitable for TDMA handset implementation. The receiver employs two branches of diversity, and is capable of adaptively choosing among three different diversity techniques which are implemented in a single design which we refer to as a multi-diversity receiver. It consists of a single conventional wireless digital receiver chain augmented with a few additional low-cost passive RF components and minor control circuits. We present algorithms for efficient co-phasing and equal-gain combining, as well as a novel co-channel interference-reduction combining algorithm. These are implemented in a system design simulation which conforms to the PACS standard. We then present simulation link performance for selection diversity (SD), equal-gain combining (EGC) and interference-reduction combining (IRC) under Raleigh fading, flat fading with co-channel interference and selective fading conditions. The results show that EGC and IRC yield a signal-to-noise ratio (SNR) improvement of 1 dB and a signal-to-interference ratio (SIR) improvement 4 dB~5.5 dB, respectively, compared with SD. The receiver also lowers the irreducible word error rate due to multipath delay spread when the Doppler shift is small. These results are compared with the results from other research.**

I. Introduction

ANTENNA diversity can both improve the quality of communications in a wireless environment, and also yield increased system capacity. Recent work by Cox and Wong has shown that two-antenna optimum-combining diversity (OC) produces a signal-to-interference (SIR) improvement of at least 3dB over conventional two-antenna selection diversity in Personal Access Communication Systems (PACS) [1]-[3]. Furthermore, in a line-of-sight (LOS) environment, selection diversity provides no gain since the two branches are correlated, while combining diversity can still cancel co-channel interference and boost the desired signal. Qualitatively speaking, in an LOS environment the OC receiver adjusts the joint signal of several antennas, resulting in an adaptive joint antenna pattern or polarization which attenuates co-channel interference while amplifying the desired signal. In a multipath environment, the two antennas may be receiving signals from separate paths and this picture is not entirely applicable, but the concept is the same.

System complexity, with the associated power consumption, cost and size, is a significant barrier to implementing diversity techniques in handsets. In particular, most proposed schemes require one receiver chain for each branch of diversity [3]. For a two-branch diversity system, this doubles the receiver circuits from RF to baseband, and is not an easy tradeoff to make where reducing system complexity is essential. Furthermore, many of the so-called adaptive antenna array solutions also rely on algorithms requiring antenna patterns which are well-characterized. In contrast, handset antennas possess patterns which are not carefully controlled and quite dependent on the position of the antenna with respect to the user's hand and head. As such, most existing antenna diversity research to date has been limited to either simple selection diversity [4], or has targeted basestation implementation [3] where complexity issues are not significant compared with handset implementations.

The receiver proposed in this paper meets some important design constraints which make it suitable for handset implementation employing TDMA architecture. First, the system utilizes a single receiver chain and baseband combining processor together with standard baseband processing techniques. The additional circuit components required to implement the system are low-cost passive RF front-end components which combine the antenna signals at RF. The system is sufficiently robust to handle poorly-defined, user-dependent antenna patterns. Furthermore, the system is able to implement several modes of diversity without changing the hardware or baseband processing, and can choose the most appropriate mode given the type of mobile usage and signal environment. We therefore refer to this as a multi-diversity receiver. Finally, the techniques can be applied to a larger number of antennas at the cost of decreased mobility and lower tolerance to fading.

Section II introduces two-antenna diversity receiver. Section III briefly describes the fully digital TDMA burst demodulator. Section IV briefly reviews the selection diversity, then presents an efficient co-phasing method for equal-gain combining and a novel co-channel interference-reduction combining (IRC) algorithm. Numerical results are shown and compared in section V.

II. Two-Antenna Multi-Diversity Receiver

The multi-diversity system presented here implements selection diversity (SD), equal-gain combining (EGC) and interference-reduction combining (IRC) using a single receiver chain design.

The system marginally increases the complexity of existing receivers to a level comparable with selection diversity, while achieving the performance gains of more complex systems under quasi-static multipath channel and interferer conditions. The receiver consists of the following components (see Fig. 1): the two antennas may consist of any type of antenna, as long as they meet spatial or polarization diversity

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conditions, i.e. they are uncorrelated and with same average received signal power in a multipath environment. In an LOS environment, selection diversity loses its effectiveness while combining techniques still provide substantial gain over single-antenna systems, since selection depends on uncorrelated signals at the two antennas while combining techniques effectively yield an adaptive antenna array with an improved pattern.

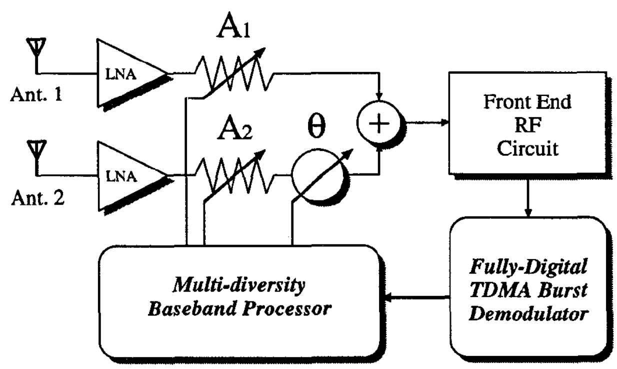

Fig. 1. Block diagram of two-branch antenna multi-diversity receiver.

The most significant hardware modifications needed to adapt a conventional receiver system are restricted to simple passive RF front-end components:

- Two voltage-controlled RF attenuators (0~-20 dB)
- One voltage-controlled RF phase shifter (0~360°)
- An RF signal combiner

Each diversity branch utilizes an attenuator for amplitude scaling. One of the branches contains the RF phase shifter. The resulting signals from each branch are summed to enhance the received signal quality and fed into the remaining receiver chain. The RF front-end converts the signal down to a low IF which is quantized by an analog-to-digital (A/D) converter. The digitized signal is processed by a fully digital TDMA burst demodulator. The baseband signal is fed into the multi-diversity baseband processor, which performs SD, EGC and IRC algorithms and adaptively adjusts the attenuators and phase shifter to optimize the received signal quality.

III. Fully-Digital TDMA Burst Demodulator

The receiver is designed to operate in a TDMA environment for PCS communications, such as PACS. We have implemented a practical system simulation which conforms to the PACS standard. The PACS standard uses $\frac{\pi}{4}$DQPSK modulation, 384kbps channel bit rate, 120 bits per time slot, 8 time slots per frame, 312.5uSec burst duration and 2.5mSec frame duration [5][6]. A fully digital demodulation architecture which is suitable for VLSI implementation has been proposed by Chuang and Sollenberger [4][7]. In this proposed implementation, $\frac{\pi}{4}$DQPSK modulation and square root raised cosine ($\alpha = 0.5$) pulse shaping is used. The receiver also uses two square root raised cosine ($\alpha = 0.5$) filters for in-phase (I) and quadrature (Q) baseband signal to match the transmitter for optimal performance in AWGN environments [9]. Other digital phase modulation systems can be treated similarly.

We adopt the fully-digital coherent demodulation technique proposed by Chuang and Sollenberger [4][7]. It jointly estimates symbol timing and carrier frequency offset by operating on an individual TDMA burst without requiring a training sequence. These estimates produce a signal quality (SQ) measurement which is a good indicator of the degree of signal impairment caused by noise, delay spread or interference, which closes the eye-opening of the detected signals [3][4]. Unlike using maximum average eye opening as symbol timing in [4], we use a square-law symbol timing scheme [11] to estimate timing, then use the values of I and Q at the sampling point to calculate the SQ and carrier phase (ϕ). A novel low-complexity diversity combining processor is added into the receiver to control the combining circuits. At the same time, the signal strength is measured through received signal strength indicator (RSSI) circuits.

A low IF bandpass signal at 768kHz (4 times the symbol rate) is sampled with an A/D at 3.072MHz, resulting in an oversample of 16 samples per symbol. This is required to achieve symbol timing recovery, signal quality measure, frequency offset estimation and carrier phase recovery without overhead [4]. Using the same downconverter architecture, coherent and differential detection can be achieved for $\frac{\pi}{4}$DQPSK. Frequency offset estimation is not addressed in this implementation. It can be removed either through RF frequency synthesizer or baseband frequency estimation [4]. Since PACS bursts are very short relative to channel variation, a quasi-static channel approximation which means the channel is static during a burst period is realistic applied to a flat fading study. All of the following simulation work for this receiver is based on these assumptions.

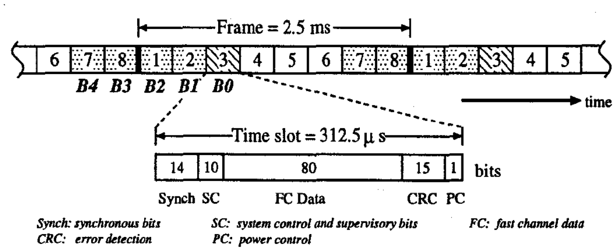

Fig. 2. PACS downlink frame structure.

IV. Antenna Diversity Methods

The PACS downlink utilizes continuous time division multiplex (TDM) transmission. As such, we take advantage of bursts **B1** ~ **B4** preceding the desired burst **B0** (shown in Fig. 2) to measure the channel and set the combining diversity parameters A_1, A_2, and θ, refer to Fig. 1. SD, EGC and IRC can be implemented via the same combining circuit and receiver chain. The system can then select the appropriate diversity technique most suited to the prevailing channel conditions. A_1, A_2, and θ are amplitudes of attenuators and phase of phase shifter, respectively. P_i, SQ_i, ϕ_i are the received signal power, signal quality and carrier phase of i'th burst, respectively. $R = \frac{A_1}{A_2}$ is the ratio of attenuation factors. We can set the larger of A_1 and A_2 to be equal to 1 and set the other equal to R or 1/R, limiting the value to the range $0.1 \sim 1$.

A. Selection Diversity (SD)

SD uses two bursts (**B2**, **B1**) to calculate the signal quality of two branch and select the better one for demodulating the consecutive desired burst (**B0**). The following steps are used:

- **B2**: Select Ant.1 by setting $A_1 = 1, A_2 = 0.1, \theta = 0°$ Receive the burst and calculate SQ_1.
- **B1**: Select Ant.2 by setting $A_1 = 0.1, A_2 = 1, \theta = 0°$ Receive the burst and calculate SQ_2.
- **B0**: Select the antenna with larger SQ, receive and demodulate desired burst.

B. Equal-Gain Combining (EGC)

Because attenuators rather than variable gain amplifiers (VGA) are used, EGC is applied in order to minimize noise figure. Therefore, A_1 and A_2 are set to 1, and the key processing which remains is to cophase the two branches.

B.1 Co-phasing

The local crystal oscillator has very high short-term stability. As such, we can assume its frequency and phase are constant during several bursts, and use it as a reference. First, the system selects antenna 1 to receive a burst (**B3**) and recover phase ϕ_1. The system then selects antenna 2 during burst (**B2**) and obtains phase ϕ_2. ϕ_1 and ϕ_2 have an ambiguity equal to an integer multiple of 90° introduced during the phase recovery process. This ambiguity causes no problems during coherent detection because differential decoding can remove it. However, we need to obtain the absolute phase difference between the two branches in order to cophase them properly. Let $\phi = \phi_1 - \phi_2$ (with ambiguity), and θ equals the desired phase difference, with no ambiguity. To remove the ambiguity, the system employs an additional burst (**B1**) which is divided into four equal periods. During these periods, the four possible phases are tested for combined signal power using the RSSI circuit. The phase yielding the smallest power is then phase-inverted (+180°) to yield cophasing.

B.2 Algorithm

- **B3**: Select Ant.1 by setting $A_1 = 1, A_2 = 0.1, \theta = 0°$. Receive the burst and calculate SQ_1, ϕ_1.
- **B2**: Select Ant.2 by setting $A_1 = 0.1, A_2 = 1, \theta = 0°$. Receive the burst and calculate SQ_2, ϕ_2.
- **B1**: Combine the two branches and cophase to get θ.
- **B0**: Set $A_1 = 1, A_2 = 1$, and θ, then receive and demodulate desired burst.

C. Interference-Reduction Combining (IRC)

In a high capacity PCS, for a given bandwidth, co-channel interference (CCI) limits the system capacity [1]. Usually, CCI is dominated by one co-channel interferer because of

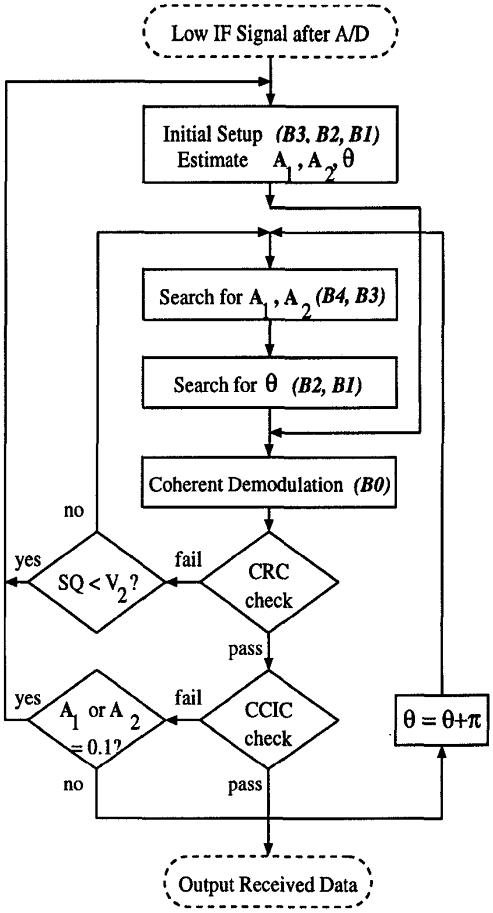

Fig. 3. IRC algorithm with two antennas.

the shadowing phenomenon, which has a log-normally distributed local mean of received signal power. In order to cancel the primary source of CCI, the system must adjust A_1 and A_2 to make the received interference I_1 and I_2 from each of the branches equal in amplitude, and then add them out of phase. Through system simulation, we obtain an approximate relationship between SIR and SQ when SIR is between 7~13 dB.. That is,

$$SIR = \frac{S}{I} \approx 4 + 38SQ \tag{1}$$

This curve is similar to the result presented in [3]. Then SIR_1 and SIR_2 can be calculate from (1).

$$SIR_1 = \frac{S_1}{I_1}, P_1 = S_1 + I_1 = I_1(1 + SIR_1) \tag{2}$$

$$SIR_2 = \frac{S_2}{I_2}, P_2 = S_2 + I_2 = I_2(1 + SIR_2) \tag{3}$$

After attenuating, the interference becomes,

$$I_1' = A_1 I_1, I_2' = A_2 I_2 \tag{4}$$

To make $I'_1 = I'_2$, we get

$$R = \frac{A_1}{A_2} = \frac{I_2}{I_1} = \frac{P_2(1+SIR_1)}{P_1(1+SIR_2)} \quad (5)$$

The IRC algorithm is divided into three major steps: initial setup, optimizing search, and coherent demodulation. Initial setup utilizes the received signal power P_i and signal quality SQ_i to estimate the amplitudes A_i of attenuators from (1)~(5) then cophase. This provides the system with a very good starting point from which to begin interference-reduction. The optimizing search determines parameters for interference-reduction and tracks channel variations. Coherent detection recovers the received data and check the cyclic redundancy check (CRC) bits and co-channel interference control (CCIC) bits to determine how to next proceed. This is now described in detail (see Fig. 3):

C.1 Initial Setup

Define two SQ thresholds, V_1 and V_2, which are used to detect certain conditions in the algorithm flow.

- **B3**: Select Ant.1 by setting $A_1 = 1, A_2 = 0.1, \theta = 0°$ Receive the burst and calculate P_1, SQ_1, ϕ_1.
- **B2**: Select Ant.2 by setting $A_1 = 0.1, A_2 = 1, \theta = 0°$ Receive the burst and calculate P_1, SQ_2, ϕ_2.
- Using P_1, P_2, SQ_1, SQ_2 to determine A_1, A_2
- **B1**: Combine the two branches and cophase to get θ.
- When $SQ > V_1$, selection diversity will be used.
- Proceed to detection step.

C.2 Optimizing Search

- **B4**: $R_1 = R + 3dB$, set A_1 and A_2 to get SQ_1.
- **B3**: $R_2 = R - 3dB$, set A_1 and A_2 to get SQ_2.
- Select one of R, R_1, and R_2 with max SQ.
- **B2**: $\theta_1 = \theta + 45°$, set θ_1 to get SQ_1.
- **B1**: $\theta_2 = \theta - 45°$, set θ_2 to get SQ_2.
- Select one of θ, θ_1, and θ_2 with max SQ.

C.3 Coherent Detection

- **B0**: Coherent detection and check CRC & CCIC.
- If CRC check fails and $SQ < V_2$ then restart a initial step else goto search step
- When CRC check is passed and CCIC check fails, this means the combiner locks to a stronger inteferer. If the A_1 or A_2 equal 0.1 (only one antenna being used for receiving), a initial setup step will be executed and the phase θ will be inverted to cancel the stronger interferer. Otherwise, simply invert the phase θ and do searching.
- If CRC & CCIC are all passed, the receiver outputs recovered data.

V. Simulation Results

The following conditions, in addition to those described in the previous sections, are used in the computer simulation:

- Coherent or differential detection
- Jakes model [10] is used for generating correlated slow Rayleigh fading channel.
- A single dominant interferer is assumed in co-channel interference environment [8]
- Delay spread effects are accounted for by using two-ray model [3][4]
- Quantization is not considered but 40 dB SNR is used as noise floor.

The system is simulated with uniform distributed symbol timing and a maximum Doppler shift of fd=6 Hz which is corresponds to a traveling speed of 2 mph at 2GHz. RSSI and SQ provide a useful indicator of the channel conditions, and are used by the multi-diversity receiver to switch between the two combining schemes automatically. Below a defined threshold, the environment is noise-limited and EGC is used. Otherwise the environment is assumed to be CCI-limited and IRC is employed. In voice and data communication, word error rate (WER) is good performance indicator than bit error rate (BER) in a bursty error environment. All the simulation will be given in WER.

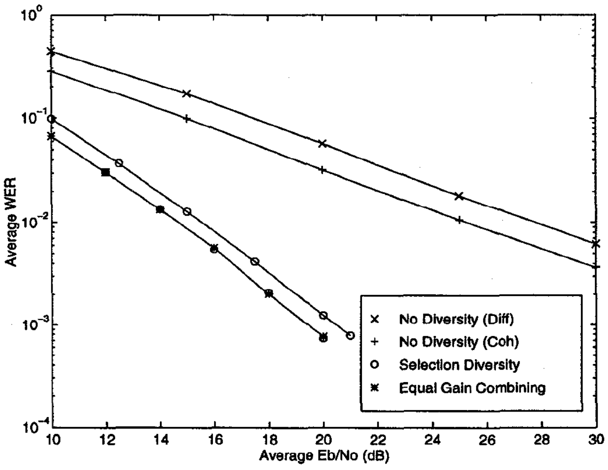

Fig. 4. Link performance under flat fading

A. Flat Fading Environment

When there is no diversity, the coherent detection is about 2.5 dB better than differential detection, see Fig. 4. Therefore coherent detection is adopted for remaining simulation. Under flat fading condition, the system uses EGC since programmable attenuators, not VGA, are used. Noise is dominant and at a constant level, so $A_1 = A_2 = 1$ yields the best noise figure. The two branches are then cophased. Simulation results are presented in Fig. 4, and show that EGC provides a 1 dB improvement over SD at WER=0.01. The results for SD performance agree closely with those presented in [8].

B. Flat Fading with CCI Environment

The IRC is employed to combat CCI. Fig. 5 shows that the performance of SD and EGC are close, and a substantial improvement of 4 dB over SD and EGC at a WER of 0.01 with a Doppler frequency fd=6 Hz. The IRC is sensitive

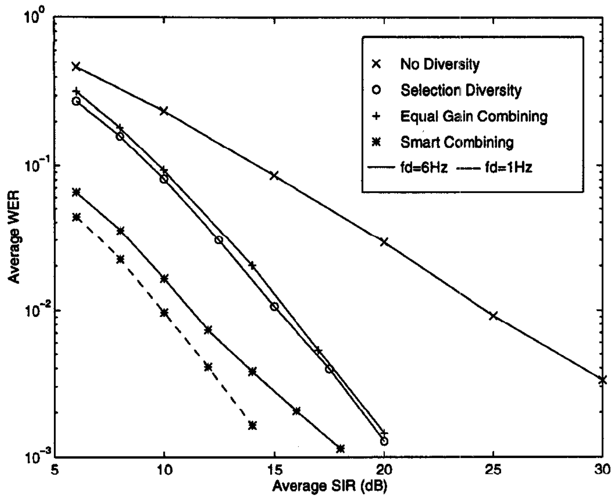

Fig. 5. Link performance under flat fading with co-channel interference

to Doppler shift because of the delay introduced by channel measure. When the Doppler frequency is 1 Hz, the SIR improvement increases to 5.5 dB compared with EGC.

A key feature of the IRC technique is its insensitivity to the accuracy of attenuators and phase shifter. The attenuator should have a range of 0~20 dB with ± 1 dB accuracy, and the phase shifter should have a range of 0~ 360° range with ±10° accuracy.

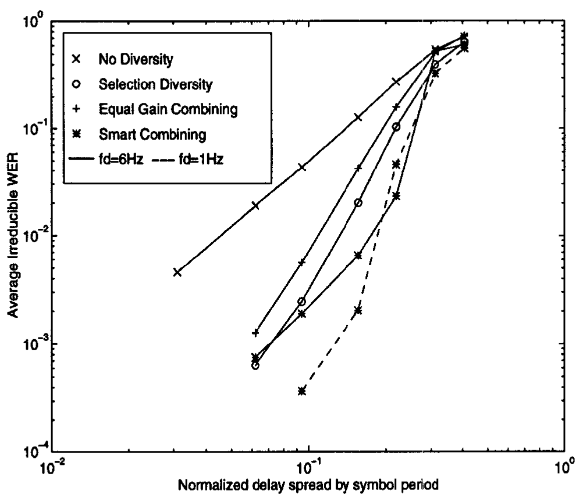

Fig. 6. Link performance under frequency selective fading

C. *Frequency Selective Fading Environment*

Because we focus on cancelling interference, the IRC may be not suitable for dealing with multipath delay spread, just as cophasing is ineffective in dealing with CCI and delay spread. Fig. 6 shows that the IRC's irreducible WER performance in frequency selective fading environment is better than SD and EGC. Normally, delay spread can be a significant factor even when signal power is high. Under such conditions, IRC is used, treating multipath as CCI.

VI. Conclusion

This paper introduces a low-complexity antenna diversity receiver suitable for TDMA handset implementation. The receiver employs two branches of diversity, and is capable of adaptively choosing among selection, equal gain combining and interference reduction combining techniques, which are implemented in a single multi-diversity recever design. The receiver employs a single conventional wireless digital receiver chain augmented with a few additional low-cost passive RF components and minor control circuits. Furthermore, the algorithm presented is insensitive to antenna pattern variations. In this paper we have presented simulation link performance of the receiver under Raleigh fading, flat fading with co-channel interference and selective fading conditions. The results show that EGC and IRC yield a signal-to-interference ratio improvement of 1 dB and 4 dB~5.5 dB, respectively, compared with selection diversity. The receiver also lowers the irreducible word error rate due to multipath delay spread, especially when the Doppler shift is small. The performance and robust, low-complexity design of this receiver makes it an excellent candidate for practical handset implementations.

Acknowledgments

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