行政院國家科學委員會專題研究計畫 成果報告

音響定位之研究
研究成果報告(精簡版)

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行政院國家科學委員會專題研究計畫成果報告

音響定位之研究

Study on Acoustic Localization

計畫編號: 98-2221-E-009-094
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摘要

GPS 技術的應用相當普遍，但若使用者位於室內或上方有遮蔽物時，就無法收到 GPS 所發射的訊號，亦即無法達成定位的目的。本計畫我們使用聲音的展頻訊號來作室內測距。我們試圖以“參考麥克風”找出同步和聲速的校準方法。此外，我們考慮衰減及非直線效應(NLOS)之下的定位系統，我們估測初始的位置，以減少 NLOS 所造成的定位誤差。由電腦的模擬可以証明我們提出的方法確實使 NLOS 之下的定位有很大的改善。此外我們也討論合作式定位，在此系統中，除了利用本來多個固定端與多個待測物的量測距離外，增加了待測物彼此之間的量測距離來做定位，達到較精準的定位效能並有初步的電腦模擬結果。

ABSTRACT

GPS is widely used for localization. However it can fail in case of indoor or obstacles because of blocked signals. A wideband spread-spectrum pseudo random acoustic signal is used for indoor localization. A reference microphone is used for synchronization and calibration of sound speed. We also consider the effects of attenuation and non-line-of-sight(NLOS).Computer simulations validate our NLOS-remedy approach is effective. Cooperative localization will be also studied.

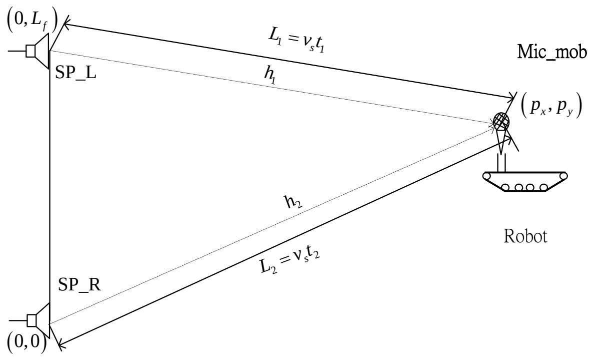
關鍵詞: 音響，定位，機器人，同步，最小方差法，泰勒級數展開，合作式定位。

Keywords: Acoustic, Localization, Robot, Synchronization, Least-squares, Taylor-series expansion, Cooperative localization.

一、緣由與目的

GPS 技術[1][2]在這幾年的應用相當普遍，但是當使用者位於室內或上方有遮蔽物時，就無法收到 GPS 所發射的訊號，亦即無法達成定位的目的，但室內做定位的應用[3][4]日益廣泛。例如室內清潔機器人，一般設計大多是隨機的去打掃房間，如果使用者知道目前機器人的所在位置，就能進一步控制它下一步的動作，有效率的達到清潔效果，減少不必要的能源消耗。另外對於醫療方面，病人的行蹤將可以由 UWB 定位系統[11]定來做安全性監控。

二維定位的基本原理是利用三角定位法，如圖一，其中放置兩個固定端的喇叭，



圖一、簡單的定位系統示意圖

其之間的距離為 L_f。建立座標系統，令下方

喇叭為座標原點，則上方喇叭座標為(0, L_f)。另外我們在機器人上放置麥克風，用來接收喇叭所送出的聲音。將麥克風收到的訊號經過估算後，得到訊號從喇叭到麥克風的飛行時間 (Time-Of-Arrival) TOA [6]。把飛行時間乘上聲速即可得到喇叭與麥克風之間的距離。不過要得到麥克風與喇叭之間的距離有二個主要的困難點：第(一)如果要知道聲音從喇叭放出到麥克風收音的這段飛行時間，則聲音從喇叭輸出的正確時間要知道。即喇叭放出聲音的同時麥克風開始收音要真正的同步 [5]。第(二) 計算距離必須用到聲速，由於聲速與當時當地的大氣條件(溫度、壓力、相對濕度)有關 [7]，因此在測量距離前必須先做聲速的校準[9]。而最近的研究發現，在喇叭與麥克風間若不存在直接路徑[8] (direct path)，則測量到的飛行距離將會大於真正的實際值，造成位置估算有較大的偏差。克服同步及聲速校準的困難是我們研究的重點。

本計畫我們試圖找出低成本，低複雜度的收送同步和聲速的校準方法，這個方法主要的關鍵是在喇叭附近，放置一個參考麥克風，來估算喇叭送出聲音前的這段隨機的時間延遲。另外由於參考麥克風與喇叭之間的距離已知，因此可用來做聲速的校準。最後，我們考慮考慮衰減及非直線效應(NLOS)之下的定位系統，它主要的目的，在於研究如何減少 NLOS 所造成的定位誤差，以確保我們的定位效能在可以接受範圍之內。我們使用的方法是: 首先產生一個初始的位置估測，由此初始的位置估測，得到新的一組與固定(傳送)端的距離。接著經由比對來估計這些 NLOS 的誤差量。再用這些估得的誤差量去修正我們先前的測量距離，或者去修正我們一開始粗略的定位估測，由電腦的模擬可以証明我們提出的方法確實使 NLOS 之下的定位有很大的改善。此外我們也討論合作式定位，在此系統中，除了利用本來多個固定端與多個待測物的量測距離外，增加了待測物彼此之間的量測距離來做定位，一般用 joint Newton's method 但它是 nonlinear function 而且運算量很高. 目前我們嘗試用 joint Taylor-series expansion method and divide-and-conquer algorithm. 來降低運算量，達到較精準的定位效能，並有初步的電腦模擬結果

二、音響定位之同步研究

圖 二圖九為二維定位系統架構圖。此系統包含了一張雙頻道的音效卡，兩個喇叭且彼此相距L_f，在機器人上的麥克風裝置，以及估算距離的模組裝置(錄音機及 correlator)。我們挑選了兩個正交的 PN-sequence [12]，同時由音效卡播放出來。房間的脈衝響應(room impulse response)h_1和h_2將由測距模組中的 recorder 和 correlator 求得。其中 timestamp 和h_1或h_2之直接路徑所對應的時間差則是估出來的飛行時間。將得到的兩個飛行時間乘上聲速可以分別得到L_1和L_2。

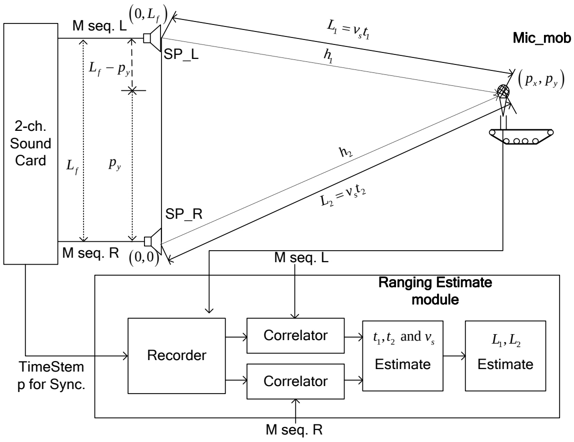


圖 二 2 維定位系統架構

我們可以由這幾段已知的長度來列出數學式子：

$$L_1{}^2 - (L_f - p_y)^2 = L_2{}^2 - p_y \qquad (1)$$

$$p_x = \sqrt{L_2{}^2 - p_y{}^2} \qquad (2)$$

從式(1)(2)我們可以求出機器人的座標為(p_x, p_y)，其中

$$p_y = \frac{L_f^2 + L_2^2 - L_1^2}{2L_f}, p_x = \sqrt{L_2^2 - p_y^2} \qquad (3)$$

(p_x, p_y)的精準度則取決於飛行時間估算的精準度。而精準的飛行時間估算則是建立在傳送接收有精準的同步。同步的重要性我們由圖十來說明。首先由於電腦的隨機延

遲，使得音效卡產生訊號後不會馬上從喇叭播放出來，我們把這段延遲(latency)稱為τ。假設喇叭送出的聲音為$s(t-\tau)$，麥克風收到的信號表為：

$$r(t) = \alpha s(t-\tau-t_1) + n(t) \qquad (4)$$

其中α為聲音在傳送過程中的衰減係數，t_1為我們要估算的信號真正飛行時間，$n(t)$為麥克風端所接收到的雜訊，一般會將之 model 成高斯隨機變數。因為麥克風端事先知道喇叭傳送訊號的原形，所以估算t_1可以藉由 match filter(correlator)來達成，其數學式表為：

$$h(k) = \int r(t)s(t-k)dt \qquad (5)$$

$h(k)$的最大值所對應的時間即為飛行時間的估算$\tilde{t}_1$。

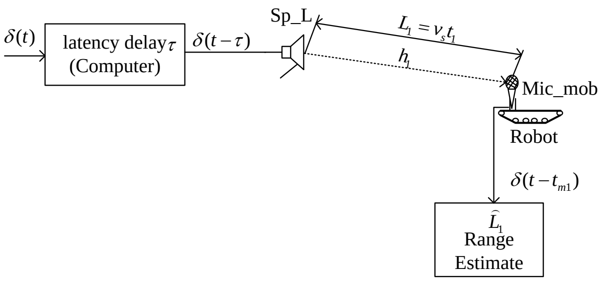


圖 三 電腦的遲滯時間所造成的同步問題

假設(5)式沒有因為雜訊產生的估算誤差，則估測的$\tilde{t}_1 = \tau + t_1$，並不等於真正的飛行時間t_1。因為(5)式中所估算的是音效卡送出訊號到麥克風接收到所需的時間，因此會多出τ這段延遲。

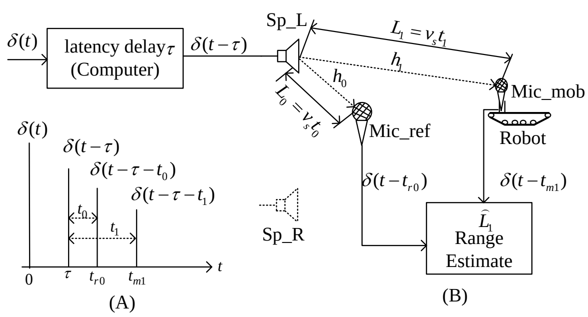


圖 四 加入參考麥克風之測距系統

為了解決喇叭與麥克風間的同步問題，我們提出的方案如圖 四 所示。為了方便起見，我們先只分析上方喇叭的距離估測。假設電腦的音效卡送出的訊號為$\delta(t)$，電腦的延遲時間為τ，喇叭的輸入訊號為$\delta(t-\tau)$。假設第一個到達麥克風的路徑為$\delta(t-t_{m1})$，其中$t_{m1} = \tau + t_1$。所以我們要找的飛行時間表為$t_1 = t_{m1} - \tau$然而這個延遲的時間是一個未知的隨機變數，必須找到有效率的方法將之估算出來。以下為我們估算τ的方法：

我們在喇叭附近放置了一個參考麥克風，喇叭與之的距離為L_0，且它們之間的房間脈衝響應為h_0。令h_0的第一個到達路徑為$\delta(t-t_{r0})$，其中$t_{r0} = \tau + t_0$，如圖 四所示。因為L_0和v_s(聲速)已知，$t_0 = \frac{L_0}{v_s}$，則電腦延遲時間可表為：

$$\tau = t_{r0} - t_0 = t_{r0} - \frac{L_0}{v_s} \qquad (6)$$

可以得到

$$t_1 = t_{m1} - t_{r0} + \frac{L_0}{v_s} \qquad (7)$$

同理，我們考慮下方喇叭時，可以得到其飛行時間：

$$t_2 = t_{m2} - t_{r0} + \frac{L_0}{v_s} \qquad (8)$$

所以L_1和L_2可以由 $L_i = v_s t_i, i = 1,2$ 算出，再將之帶入式(3)，我們就可以得到麥克風(機器人)的位置估算(p_x, p_y)。我們將利用在同步時所加的參考麥克風來做聲速的校準。如圖 五 所示，L_0和L_3可以事先測量得到，其分別又可以表示成 $L_i = v_s t_i, i = 0,3$。假設h_0和h_3的第一個到達路徑分別為$\delta(t-t_{r0})$ 和$\delta(t-t_{r3})$，其中$t_{r0} = \tau + t_0$，$t_{r0} = \tau + t_3$ 我們可以推得:

$$t_{r3} - t_{r0} = t_3 - t_0 = \frac{L_3 - L_0}{v_s} \qquad (9)$$

因此，我們估算出來的聲速$\hat{v}_s$可以表示為：

$$\hat{v}_s = \frac{L_3 - L_0}{t_{r3} - t_{r0}} \qquad (10)$$

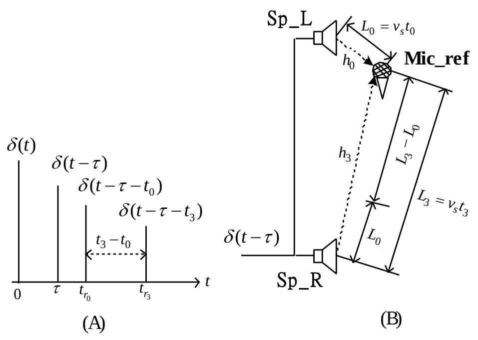


圖 五 聲速校準的方法

我們以圖十二做說明。假設喇叭到參考麥克風之間的 delay 為t_0，是一個已知數。這個方法是將參考麥克風所收到的訊號$\delta(t-t_{r0})$在時間軸上往前修正t_0，把它當成是喇叭所發出的信號$s_l(t)$，表示為

$$s_l(t) = \delta(t - t_{r0} + t_0) \qquad (11)$$

再與機器人上擺的麥克風訊號$r(t)$做一次 correlation，output 訊號表示為

$$s_l(t) \otimes r(t) \qquad (12)$$

其中$\otimes$表示 cross-correlation，其最大值所對到的平移時間則為估計的飛行時間。

三、衰減及非直線效應之下的定位系統

如同前面所述，為了使定位系統更 robust，我們將會放置3個以上固定端的傳送器(喇叭或天線)。圖 六為典型的多喇叭室內定位系統

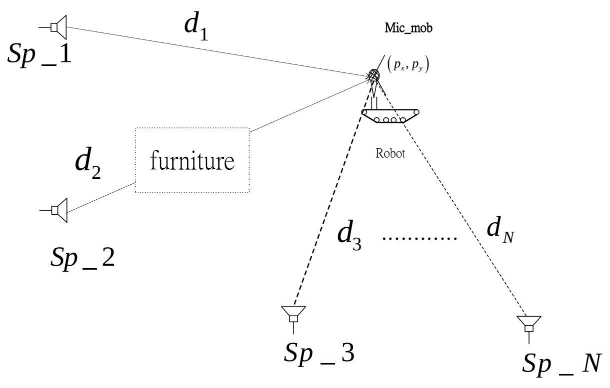


圖 六 典型的多喇叭室內定位系統

在此假設我們用了三個喇叭(N=3)來做距離量測。d_1、d_2、d_3分別為三個喇叭到機器人之間的真正距離，距離量測分別為為L_1、L_2、L_3。但因為在室內的環境之下，有背景雜訊以及 NLOS 誤差，使得L_1、L_2、L_3未必全等於d_1、d_2、d_3。由於我們有三個數學近似式，未知數只有兩個(二維定位)，其關係式可表示為：

$$\begin{aligned} d_i &= \sqrt{(x-x_i)^2 + (y-y_i)^2} \\ &= L_i + n_i \quad, i = 1,2,3 \end{aligned} \qquad (13)$$

其中(x_i, y_i)為喇叭編號i之座標，(x, y)為待求的機器人座標，d_i為麥克風到喇叭之真正距離。n_i為測距誤差。我們採用 Least Square(LS) Solution，其數學式表為：

$$(\tilde{x}, \tilde{y}) = \arg\min_{x,y} \sum_{i=1}^{N} \left(\sqrt{(x_i - x)^2 + (y_i - y)^2} - L_i\right)^2$$

……………………………………………(14)

其中 $c(x, y) = \sum_{i=1}^{N} \left(\sqrt{(x_i - x)^2 + (y_i - y)^2} - L_i\right)^2$

稱為成本函數(cost function)，N為喇叭個數。我們可以發現，式(14)對於待求的未知數來說，是一個非線性方程式，無法得到簡單的 LS 公式解。一般使用 Taylor-series-based Linearized Least Squares 來解，我們以式(11)(14)為基礎，定義

$$F_i(\theta) = \sqrt{(x-x_i)^2 + (y-y_i)^2} \qquad (15)$$

而泰勒展開式可表為:

$$F(\theta) = F(\theta_0) + \nabla F(\theta_0)(\theta - \theta_0) + \text{high order term} \qquad (16)$$

假設我們忽略高階項，結合式(13)，我們可以寫成：

$$F_i(\theta) \approx F_i(\theta_0) + \frac{\partial F_i}{\partial x}(x - x_0) + \frac{\partial F_i}{\partial y}(y - y_0)$$

………………………………(17)

其中

$$F_i(\theta_0) = \sqrt{(x_0 - x_i)^2 + (y_0 - y_i)^2} = d_{i,0} \qquad (18)$$

$$\frac{\partial F_i}{\partial x} = \frac{x_0 - x_i}{d_{i,0}}, \quad \frac{\partial F}{\partial y} = \frac{y_0 - y_i}{d_{i,0}} \qquad (19)$$

將式(16)結合式(13)可以改寫成

$$\begin{aligned} F_i(\theta) \approx d_{i,0} &+ \frac{x_0 - x_i}{d_{i,0}}(x - x_0) \\ &+ \frac{y_0 - y_i}{d_{i,0}}(y - y_0) \approx L_i \end{aligned} \qquad (20)$$

將式(20)未知數整理至等號左邊

$$(\frac{x_0 - x_i}{d_{i,0}})x + (\frac{y_0 - y_i}{d_{i,0}})y \approx L_i - \tilde{d}_{i,0} \quad ....(21)$$

其中

$$\tilde{d}_{i,0} = d_{i,0} - (\frac{x_0 - x_i}{d_{i,0}})x_0 - (\frac{y_0 - y_i}{d_{i,0}})y_0 \quad (22)$$

因為式(13)中有三個近似式，我們把式(21)寫成矩陣形式:

$$H \times \theta = \begin{bmatrix} \frac{x_0 - x_1}{d_{1,0}} & \frac{y_0 - y_1}{d_{1,0}} \\ \frac{x_0 - x_2}{d_{2,0}} & \frac{y_0 - y_2}{d_{2,0}} \\ \frac{x_0 - x_3}{d_{3,0}} & \frac{y_0 - y_3}{d_{3,0}} \end{bmatrix} \times \begin{bmatrix} x \\ y \end{bmatrix} \approx \begin{bmatrix} L_1 - \tilde{d}_{1,0} \\ L_2 - \tilde{d}_{2,0} \\ L_3 - \tilde{d}_{3,0} \end{bmatrix} = b$$

…………………………………(23)

由式(23)我們可以得到線性 least squares solution 表為

$$\theta = (H^T H)^{-1} H^T b \qquad (24)$$

由式(15)到式(24)的推導，我們可以得知用泰勒展開式的好處在於它可以使得 least square 的數學式”線性化”。經由化簡整理，我們才能夠得到如式(23)這種乾淨簡單的解。可惜的是，泰勒展開在式(16)是用”近似”的方法，所以當初始猜測座標 θ_0 不夠接近待測物(robot)時，其展開式就不會以”近似”的形式呈現，造成式(24)的解跟待測物座標有相當大的偏差。因此我們可以仿效疊代的方法，將式(24)的解代回式(16)中的 θ_0 ，重覆數次，直到 $\|\theta_{i+1} - \theta_i\| < \delta$ 。

接下來我們要考慮兩個路徑的效應，分別是 NLOS(Non-line-of-Sight)以及路徑衰減。我們將重點放 NLOS Identification 之後的訊號處理。NLOS 發生時，我們的 range measurement 將會有過長(positive bias)的現象，我們把這個正的偏差 model 成 exponential 分布。假設在房間裡放置了Ｎ個喇叭。假設喇叭與麥克風的真正距離為

$$d_i = \sqrt{(x - x_i)^2 + (y - y_i)^2} \quad i = 1, 2......, N \quad (25)$$

則 range measurements 可表為

$$L_i = d_i + \varepsilon_i \qquad (26)$$

其中 $\varepsilon_i = \begin{cases} n_i & i = 1, 2..k.. \\ N_i + n_i & i = k+1, ..N, \end{cases}$

而 n_i 通常 model 成高斯分佈，為 time of arrival 的 error。N_i 則為 NLOS error。有了這些 range measurements 和 model，我們由式(24)的方法先得到一個初估的座標 $(\tilde{x}, \tilde{y})$。有了初估的座標後，我們希望找到一個方向 (δ_x, δ_y) ，使得 $(\tilde{x} + \delta_x, \tilde{y} + \delta_y)$ 更加精確。首先我們去重算這個初估座標 $(\tilde{x}, \tilde{y})$ 和喇叭的距離，稱為 $\tilde{L}_i$。一般而言我們相信對有 NLOS 誤差的喇叭來說，重算後的距離將會比原來更準確。因此把測到的距離修改成新的：

$$L_i = \tilde{L}_i \text{ ， } i = k+1, ..., N \qquad (27)$$

利用泰勒的方法把式(11)對 $(\tilde{x}, \tilde{y})$ 展開，我們可以得到

$$(\frac{\tilde{x} - x_i}{\tilde{L}_i})\delta_x + (\frac{\tilde{y} - y_i}{\tilde{L}_i})\delta_y \approx L_i - \tilde{L}_i \text{ ， } i = 1, 2....., k$$

(NLOS) ……………………………(28)

$$(\frac{\tilde{x} - x_i}{\tilde{L}_i})\delta_x + (\frac{\tilde{y} - y_i}{\tilde{L}_i})\delta_y \approx 0 \text{ ， } i = k+1, ..., N$$

(LOS)………………................(29)

因為式(27)的結果使得式(29)近似式右方為 0。將式(28)(29)結合，可以得到一組 least square 的解 (δ_x, δ_y) 。我們可以利用這個新的座標估測 $(\tilde{x} + \delta_x, \tilde{y} + \delta_y)$ 對式(13)做泰勒展開，即用疊代的方法求解，直到 (δ_x, δ_y) 收斂到某個極小的值。

圖 七 是我們提出的演算法流程圖，分為三個步驟：

圖 七 減緩 NLOS 誤差流程圖

第（一）Initial LS estimation，我們先得到一個初估的座標 $(\tilde{x}, \tilde{y})$ 。第(二) NLOS tradeoff，從式(27)中，我們會有新的 NLOS 測距 $\tilde{L}_i$ ，也有一開始的測距 L_i，$i = k+1, ..., N$ 。由於新的測距不見得完全可靠，我們在新舊測距中做一點折衷：

$$\hat{L}_i = \mu \tilde{L}_i + (1 - \mu) L_i \text{ ， } i = k+1, ..., N \qquad (30)$$

其中μ為 0~1 之間的常數。

第(三) LS estimation，利用 LOS 的測距L_i，$i=1,...,k$，和式(30)的測距$\hat{L}_i$，$i=k+1,...,N$，我們可以用任一種 LS 方法來估測待測物的座標。接著我們考慮路徑衰減(path loss)的效應。上述方法並未考慮式(4)中接收訊號的 gain α之效應。[13][14]提出的方法就是利用信號的能量來估測距離。信號在傳送過程中，若飛行距離相對遠，則收到的信號能量則相對小。反之，若飛行距離相對近，則收到的信號能量則相對大。其方法是在測距前先建立 path-loss–model(PLM)，可以表為

$$\frac{P_r}{P_0}=k_0 d^{-n} \qquad (31)$$

其中P_r為接收訊號之能量，P_0為傳送訊號之能量，d為傳送距離。k_0、n為衰減常數，將由統計資料而求得。 有了這個 model，我們只要根據傳送訊號及接收訊號算出P_r、P_0，就可以得到距離估測

$$\tilde{d}=\left(\frac{P_r}{k_0 P_0}\right)^{\frac{-1}{n}} \qquad (32)$$

接下來我們想在飛行時間估測的過程當中，把 correlator 加入 power 的概念，如圖 八。

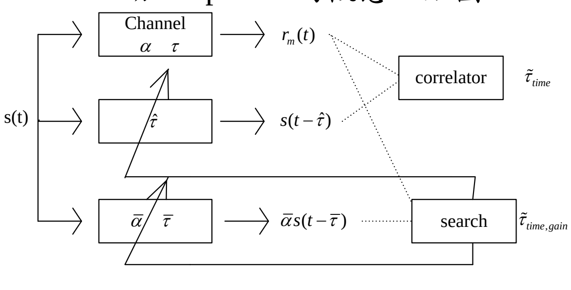


圖 八 用 path-loss 輔助估測飛行時間

我們先考慮單一的喇叭測距，收到的訊號為

$$r_m(t)=\alpha s(t-\tau)+n(t) \qquad (33)$$

其中$s(t)$為發射訊號，$n(t)$為背景雜訊，τ為我們要估測的飛行時間。而從 path loss 統計上的 model，我們可以得知衰減系數

$$\alpha=k_0\tau^{-n} \qquad (34)$$

其中k_0、p可以由事前統計上的 model 得知，圖 八 中的$\bar{\alpha}$將等於$k_0\bar{\tau}^{-n}$。則式(33)可以改寫成

$$r_m(t)=k_0\tau^{-n}s(t-\tau)+n(t) \qquad (35)$$

我們的估測將會變成

$$\min_{\tau}\left\|r(t)-k_0\tau^{-n}s(t-\tau)\right\|^2 \qquad (36)$$

其中$r(t)$是實際在麥克風收到的訊號。式(36)將會是一個高非線性度的 LS 問題，最佳解必須用 Iterative Nonlinear Least Square 來得到。但其運算複雜度太高，我們將用其它次佳的方法來找出式(36)的解。首先我們用傳統的 correlator 先估得一個飛行時間$\tilde{\tau}_{time}$。接下來我們來解釋圖十八下半部 search 所做的處理。將$\tilde{\tau}_{time}$代入式(36)中的$s(t-\tau)$，可以得到

$$E_1=\left\|r(t)-k_0\tilde{\tau}_{time}{}^{-n}s(t-\tilde{\tau}_{time})\right\|^2 \qquad (37)$$

$\tilde{\tau}_{time}$的估計可能會有誤差，因此式(28)的值不一定會最小。所以我們在此引進式(25)衰減系數的效應。我們在$\tilde{\tau}_{time}$之前後開一個時間的 window，我們將此 window 代表為：

$$\bar{\tau}\in[\tilde{\tau}_{time}-t_w,\ \tilde{\tau}_{time}+t_w] \qquad (38)$$

在這個時間 window 裡，希望找到某個時間估測$\tilde{\tau}_{time,gain}\in\bar{\tau}$代入式(36)找到其最小值。換句話說，我們是想要利用 gain 來修正 correlator 時間估測的不準確。

四、合作式定位系統

合作式定位系統中，除了利用本來多個固定端與多個待測物的量測距離外，增加了待測物彼此之間的量測距離來做定位。根據合作對象的不同可大致分為：待測物之間的合作[21]、待測物與移動傳送器間的合作[19]和移動接收器之間的合作[20]等。而傳統的定位系統，如圖 九，則是利用多個固定端與待測物的量測距離來作定位。

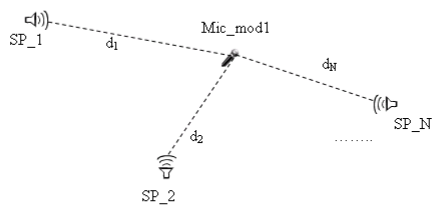


圖 九 典型的定位系統

合作式定位系統[17][18]較傳統的定位系統複雜許多，對於合作式定位，問題衍生

成如何將待測物之間的量測距離與傳統定位中的量測距離做整合，來完成較精準的定位效能。在傳統的定位系統中，過去我們是用 Maximum likelihood (ML)演算法去估計待測物的座標，為了降低這種非線性解法的運算複雜度，會將之處理成線性化。對於合作式定位系統亦產生更複雜的非線性方程式。圖十 是合作式定位系統的示意圖。我們假設有 N 個已知感測器(sensors or microphones) 的位置. 以及 M 個未知行動台(mobiles or robot)的位置。我們除了量測行動台 i 到感測器 j 之間的距離 $d_{i\bar{j}}$ 之外，同時也量測了行動台 i 和 行動台 j 之間的距離 d_{ij} 假設 r_{ij} 是行動台 i 和 行動台 j 之間的真正距離。則我們將以 $d_{i\bar{j}}$，d_{ij} 和 r_{ij} 這些參數來估計 M 個行動台的位置

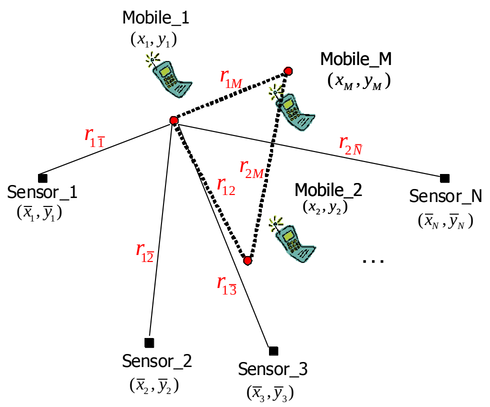


圖 十合作式定位系統.

中虛線表示合作式行動台和行動台之間距離。量測行動台 i 到感測器 j 之間的距離 $d_{i\bar{j}}$ 可表為: $d_{i\bar{j}} = r_{i\bar{j}} + n_{i\bar{j}},\quad i = 1 \sim M,\quad j = 1 \sim N$ 而行動台 i 和 行動台 j 之間的距離 $d_{ij} = r_{ij} + n_{ij}$ 其中 $n_{i\bar{j}}$ 和 n_{ij} 為 AWGN $n_{i\bar{j}} \sim N\left(0, \sigma_{i\bar{j}}^2\right)$ 及 $n_{ij} \sim N\left(0, \sigma_{ij}^2\right)$。我們有 M 個未知行動台的位置需要被估計。合作式 likelihood function 可以表為

$$p\left(\mathbf{d}_{Uncoop}, \mathbf{d}_{Coop} \mid \mathbf{x}\right) = \underbrace{\prod_{i=1}^{M}\prod_{j=1}^{N}\frac{1}{\sqrt{2\sigma_{i\bar{j}}^2}}\exp\left(-\frac{\left(d_{i\bar{j}} - \left\|\mathbf{x}_i - \bar{\mathbf{x}}_j\right\|\right)^2}{2\sigma_{i\bar{j}}^2}\right)}_{Uncooperation} \cdot \underbrace{\prod_{\substack{i,j=1\\ j>i}}^{M}\frac{1}{\sqrt{2\sigma_{ij}^2}}\exp\left(-\frac{\left(d_{ij} - \left\|\mathbf{x}_i - \mathbf{x}_j\right\|\right)^2}{2\sigma_{ij}^2}\right)}_{Cooperation} \quad (39)$$

其中非合作式的能被觀察的距離的集合表為 $\mathbf{d}_{Coop} = \left\{d_{12}, d_{13}, \cdots, d_{1M}, d_{23}, d_{24}, \cdots d_{2M}, \cdots d_{M-1,M}\right\}$

行動台的位置表為 $\mathbf{x} = \left[\mathbf{x}_1 \quad \mathbf{x}_2 \quad \cdots \quad \mathbf{x}_M\right]^T$

而要被估計的行動台的位置為 $\hat{\mathbf{x}} = \left[\hat{\mathbf{x}}_1 \quad \hat{\mathbf{x}}_2 \quad \cdots \quad \hat{\mathbf{x}}_M\right]^T$

$\hat{\mathbf{x}}$ 以 ML criterion searches 表示如下

$$\min_{\hat{\mathbf{x}}}\left\{\begin{array}{l}\underbrace{\sum_{i=1}^{M}\sum_{j=1}^{N}\frac{1}{2\sigma_{i\bar{j}}^2}\left(d_{i\bar{j}} - \left\|\mathbf{x}_i - \bar{\mathbf{x}}_j\right\|\right)^2}_{Noncooperation} + \\ \underbrace{\sum_{\substack{i=1\\ j>i}}^{M}\frac{1}{2\sigma_{ij}^2}\left(d_{ij} - \left\|\mathbf{x}_i - \mathbf{x}_j\right\|\right)^2}_{cooperation}\end{array}\right\} \quad (40)$$

解(40)式一般用 joint Newton’s method 但它是 nonlinear function 而且運算量很高. 目前我們嘗試用 joint Taylor-series expansion method and divide-and-conquer algorithm.來降低運算量，初步結果如下一節電腦模擬所示.

五、電腦模擬及結果:

第（一）NLOS model

我們實驗模擬的定位幾何結構。我們用了五個固定式的傳送器，用來測量與待測物之間的距離。房間的尺寸為 $30 \times 30(m)$，傳送器分別置於四個角落及中正央，而機器人將在房間內隨機走動。測距誤差方面，我們將 Time of arrival error model 成平均為零的高斯分布，variance 為 0.5 公尺。而 NLOS 誤差則 model 成 Rayleigh 分布，平均值為 4 公尺，variance 為 2 公尺。評估效能的標準為方均根誤差(RMS error)，模擬圖中每一點的模擬次數為 5000 次。圖 十一為定位效能和 NLOS

接收個數的關係圖。NLOS 誤差的個數越多，定位的精確度越差。從圖中可以清楚的得知，我們提出的方法($\mu \fallingdotseq 0.8$)確實使 NLOS 之下的定位有很大的改善，而且比[31]提出的方法在準確度上平均改善了約 1 公尺。另外，不做任何 NLOS 處理的定位效果在 NLOS 個數等於 2 時，方均根誤差將會大於 6 公尺，為了清楚比較減緩 NLOS 誤差的效能，我們只放上改善後的兩條線。

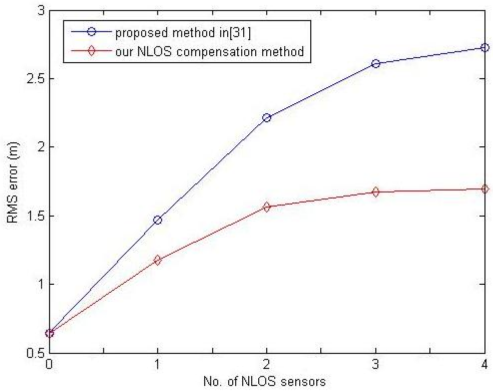


圖 十一定位效能與 NLOS 個數的曲線圖

第(二)簡單辨別 NLOS 的方法

首先我們由 LOS 和 NLOS 測距結合的 LS estimation 得到一個座標的初估。利用初估座標去算出與喇叭編號 i 的距離為 $\tilde{L}_i$。一般我們認為對 NLOS 測距來說，$\tilde{L}_i$ 會比初始測距 L_i 要來得準確。因此我們的檢測方法為：

如果 $L_i - \tilde{L}_i > \delta$，則我們就判定它是 NLOS。其中 δ 是一個稍微大於零的數，且必須適當設定的參數。圖 十二是用我們提出的檢測方法做定位的效能結果。雖然結果會比直接用單純 least-squares 解座標來得好一點，但是與真正知道 NLOS 資訊後做處理的定位結果有一段差距。

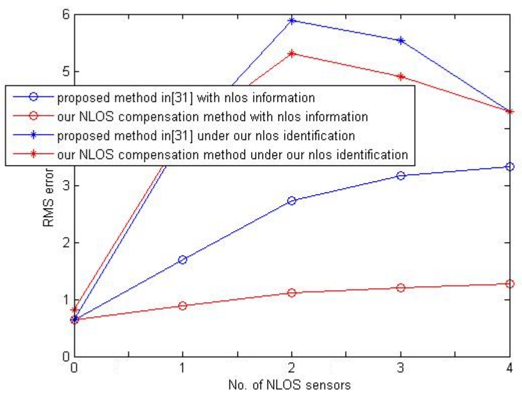


圖 十二 檢測 NLOS 減緩定位誤差

(一) Weighted-Least-Squares 模擬

我們假設機器人的位置為(30,15)，我們由事前的統計，可以得知每一組傳輸路徑大概的可靠度，使其 least squares 的權重等於它測距變異之倒數，如式**錯誤! 找不到參照來源。**。以這個座標來說，會有兩個路徑相對較遠，很可能成為相對不可靠的測距。所以不可靠的測距會有 0~2 個，對有(無)加權重的定位精準度引響如圖 十三所示。其中我們求助於 MATLAB 內建的方程式 lsqnonlin 來找出 LS 的解。因為是牛頓法，我們將初始座標猜測設定為房間的幾何中心(15,15)。從圖中我們可以看出，WLS 的方法比一般統一權重的 LS 來得精準，而且隨著不可靠的測距數增加，兩者間的準確度差別越大。所以 WLS 確實可以明顯改善定位的精確度。

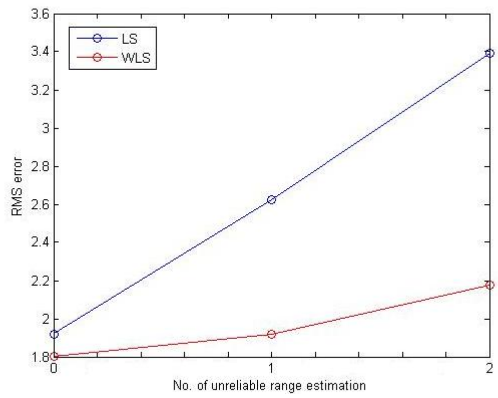


圖 十三 圖 WLS/LS 之定位效能比較

圖 十四 是合作式定位系統和另外兩個系統,divided TS, 及 joint TS 在 MSE vs. noise variance 的比較. 在低的 noise variance 時, joint TS 較 divided TS 好. 但在低的 noise variance 時則反之

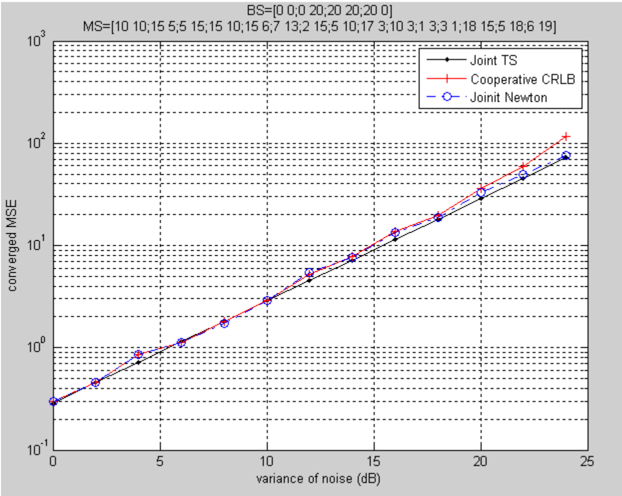


圖 十四 合作式定位系統和,divided TS, 及 joint TS 的比較

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國科會出國報告書

2009 年 10 月 9 日

<table>
<tr><td rowspan="2">報告人姓名</td><td rowspan="2">謝世福</td><td rowspan="2">申請單位
（學生請加註系級）</td><td rowspan="2">交通大學電信系</td><td>職稱</td><td>副教授</td></tr>
<tr><td>電話</td><td>03-5731974</td></tr>
<tr><td>出國目的/發表論文題目</td><td colspan="5">參加國際學術論文會議，發表兩篇論文:
1. A Diagonally Weighted Space-Time Block Code OFDM with Channel Estimation
2. Maximum SNR Detection for Selection-Relaying Cooperative System</td></tr>
</table>

一、參加經過

第一天在北海道大學舉辦 tutorials 課程，接下來三天安排四類議程: plenary talk, panel discussion, oral sessions, poster sessions.

二、心得（可含照片）

此次會議以亞太地區為主，希望可以凝聚此區之訊號處理學者，提升創新研究領域.
下次預定於新加坡舉行.

三、攜回資料名稱及內容

會議論文集 CD 一片，議程摘要一本.

四、發表論文如下:

A DIAGONALLY WEIGHTED SPACE-TIME BLOCK CODE OFDM WITH CHANNEL ESTIMATION

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ABSTRACT

We focus on space-time block code (STBC) OFDM system with four transmit antennas. A block diagonal (BD) non-orthogonal STBC is proposed for full-rate transmission of complex data. To estimate the channel response, a BD STBC matrix has to be nonsingular for all possible transmitted BD STBC data matrices. A diagonal weighting constant is proposed to assure the nonsingularity. A phase-direct technique can be used to further improve the subspace-based semi-blind channel estimation. From computer simulation the proposed channel estimation is shown to be effective and we can also see that a large weighting constant improves channel estimation at the cost of a degraded bit error rate.

Index Terms — space-time code, OFDM, channel estimation

1. INTRODUCTION

Space-Time Block Code (STBC) Orthogonal Frequency Division Multiplexing (OFDM) has been popular in wireless communications for its advantages of transmit and time diversity to combat fading [1]. In case of four-transmit-antenna, non-orthogonal STBC [3] can achieve full transmission rate for complex data symbol, such as QPSK. Similar to previously proposed two non-orthogonal STBC's, a new block diagonal (BD) STBC is proposed and it can be shown to be the remaining one yet to be found.

At its receiver end, channel estimation is necessary for decoding and equalization. In case of a complex QPSK STBC OFDM system, the STBC data matrix can possibly be singular. As a result, its matrix inverse does not exist and the channel estimation fails. To overcome this singularity issue, we propose to impose a diagonal weighting constant on the STBC matrix.

A subspace-based semi-blind channel estimation has been proposed [4]. A phase-direct technique [5,6] can be employed to enhance its performance, but it only deals with the case of two-transmit-one-receive antenna and a real BPSK data. We will apply phase-direct channel estimation to the case of four-by-one antenna non-orthogonal STBC and QPSK data.

This paper is organized as follows. After presenting the STBC system model in section 2, we propose a diagonally weighted block diagonal STBC and the phase-direct channel estimation in section 3. Section 4 derives the performance analysis from which we can see effects of the weighting constant on channel estimation error and bit error rate. Section 5 shows our simulation results. Finally, our conclusions are summarized in section 6.

2. STBC MODEL

Fig. 1 shows a complete STBC OFDM transceiver in time domain. A block precoder is needed to apply the subspace-based channel estimation [4]. $\mathbf{W}_{IFFT}$ and $\mathbf{W}_{FFT}$ denote Fourier transform matrices, while $\mathbf{A}_{CP}$ and $\mathbf{R}_{CP}$ means the cyclic prefix insertion and deletion.

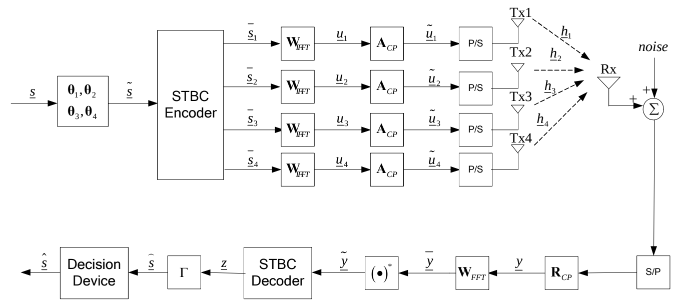

Fig. 1. Four-transmit-antenna STBC OFDM transceiver model with block precoders

For simplicity, a STBC transceiver system model in frequency domain is shown in Fig. 2. Suppose *4* transmit antennas are used. $\mathbf{S}$ denotes the STBC transmission matrix, made up by *4* symbol vectors s_1, s_2, s_3, s_4 and their conjugates. The channel frequency response vector and the AWGN are denoted by $\underline{\mathbf{h}}$ and $\underline{\mathbf{n}}$, respectively.

First, The ST encoder takes OFDM symbols to compose the transmission matrix $\mathbf{S}$, which is then fed into the channel. The received data vector $\underline{\mathbf{y}}$ can be written in frequency domain as:

$$\underline{\mathbf{y}} = \mathbf{S} * \underline{\mathbf{h}} + \underline{\mathbf{n}} \tag{3}$$

which can be useful for channel estimation. Alternatively, we can express the complex conjugate of $\underline{\mathbf{y}}$ as

$$\underline{\mathbf{y}}' = \mathbf{H} * \underline{\mathbf{s}} + \underline{\mathbf{n}}' \tag{4}$$

$\mathbf{H}$ is the channel state matrix in which h_1, h_2, h_3, h_4 and their conjugates form its elements. In Eq. (4), we can recover $\underline{\mathbf{s}}$ from $\underline{y}$ by

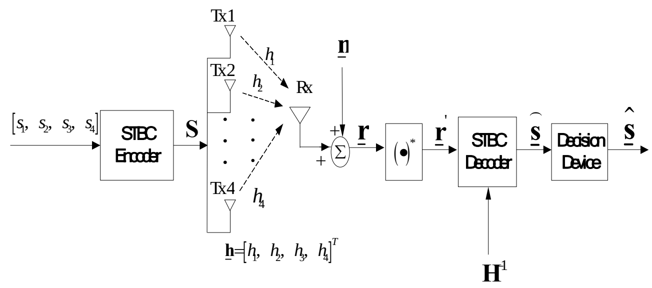

Fig. 2. Basic STBC transceiver model in frequency domain

$$\hat{\underline{\mathbf{s}}}=\mathbf{H}^{-1}*\underline{\mathbf{y}}' \tag{6}$$

For real data symbol, orthogonal STBC matrix can achieve full transmission rate. However, for complex data symbol, such as QPSK, only non-orthogonal STBC data matrix can maintain full transmission rate. There have been two non-orthogonal STBC schemes proposed previously [2,3]. Here, we propose a block-diagonal complex non-orthogonal STBC matrix as

$$\mathbf{S}=\begin{bmatrix} s_1 & s_2 & s_3 & s_4 \\ -s_3^* & -s_4^* & s_1^* & s_2^* \\ s_2 & s_1 & s_4 & s_3 \\ -s_4^* & -s_3^* & s_2^* & s_1^* \end{bmatrix} \tag{7}$$

We note that $\mathbf{S}^H*\mathbf{S}$ is of block diagonal form:

$$\mathbf{S}^H*\mathbf{S}=\begin{bmatrix} a_6 & b_6 & 0 & 0 \\ b_6 & a_6 & 0 & 0 \\ 0 & 0 & a_6 & b_6 \\ 0 & 0 & b_6 & a_6 \end{bmatrix} \tag{8}$$

where

$$a_6=\sum_{i=1}^{4}|s_i|^2 \tag{9}$$

$$b_6=s_1s_2^*+s_2s_1^*+s_3s_4^*+s_4s_3^*=2\,\mathrm{Re}[s_1s_2^*+s_3s_4^*] \tag{10}$$

$\mathbf{S}^H*\mathbf{S}$ of the other two non-orthogonal STBC matrices are similar except for the locations of the nonzero off-diagonal elements that account for the non-orthogonality between columns of S. Eq. (8) shows that the two non-orthogonal pairs of $\mathbf{S}$ are the 1^{st} and 2^{nd} columns, the 3^{rd} and 4^{th} columns. Since there are only 3 possibilities of non-orthogonal pairs among 4 data columns, our proposed block diagonal form is indeed the only remaining non-orthogonal STBC other than $\{(1^{st},3^{rd})(2^{nd},4^{th})\}$ and $\{1^{st},4^{th}\},(2^{nd},3^{rd})\}$ pairings proposed by [2] and [3], respectively.

3. CHANNEL ESTIMATION FOR STBC OFDM

Previously, phase direct (PD) for OFDM in [1] was incorporated with the subspace method [2] in [3] to enhance channel estimation. We will extend this PD technique to BD STBC OFDM and explain why diagonally weighted STBC is necessary for nonsingular channel estimation.

3.1 Diagonally weighted STBC

Consider the mth subcarrier in Eq. (3),

$$\underline{\mathbf{H}}_m=\left[\mathbf{S_m}\right]^{-1}*\underline{\mathbf{y}}_m \tag{11}$$

where $\underline{\mathbf{H}}_m$ is the 4x1 channel frequency response vector at the mth subcarrier, and $\mathbf{S_m}$ and $\underline{\mathbf{y}}_\mathbf{m}$ are transmission matrices corresponding to all possible data and received data vector, respectively.

We note that there exists a problem of $\mathbf{S_m}$ in Eq. (11). For some corresponding $\mathbf{S_m}$ of possible data in BD when BPSK or QPSK is used, $\mathbf{S_m}$ could be singular. In order to prevent $\mathbf{S_m}$ from being singular for all possible data, a diagonally weighted method is therefore proposed as follows:

$$\mathbf{S}=\begin{bmatrix} k*s_1 & s_2 & s_3 & s_4 \\ -s_3^* & k*(-s_4^*) & s_1^* & s_2^* \\ s_2 & s_1 & k*s_4 & s_3 \\ -s_4^* & -s_3^* & s_2^* & k*s_1^* \end{bmatrix} \tag{12}$$

with

$$\mathbf{S}^H*\mathbf{S}=\begin{bmatrix} a_{61} & b_{61} & c_{61} & 0 \\ b_{62} & a_{62} & 0 & -c_{62} \\ c_{61} & 0 & a_{62} & b_{61} \\ 0 & -c_{62} & b_{62} & a_{61} \end{bmatrix} \tag{13}$$

where

$$a_{61}=k^2|s_1|^2+\sum_{i=2}^{4}|s_i|^2 \tag{14}$$

$$a_{62}=\sum_{i=1}^{3}|s_i|^2+k^2|s_4|^2 \tag{15}$$

$$b_{61}=k(s_1^*s_2+s_3s_4^*)+(s_1s_2^*+s_3^*s_4) \tag{16}$$

$$b_{62}=k(s_1s_2^*+s_3^*s_4)+(s_1^*s_2+s_3s_4^*) \tag{17}$$

$$c_{61}=(k-1)(s_1^*s_3+s_2^*s_4) \tag{18}$$

$$c_{62}=(k-1)(s_1s_3^*+s_2s_4^*) \tag{19}$$

We assume that the diagonal weighting constant k is a real positive number. When $k\neq 1$, the matrix can be shown to be nonsingular for any possible symbol data.

3.2 Phase Direct (PD) --- An improved method for Subspace Channel Estimation

PD is to resolve channel phase ambiguity after getting channel power response. In conventional OFDM system, it is easy to obtain channel power response by simple computation. But in STBC OFDM it is quite different since channel power consists of several different data symbols.

Fig. 3 shows a flowchart of channel power estimation. Suppose we have an initial channel estimate $\mathbf{H}_{est,m}$ obtained from the subspace-based algorithm. Here we aim to find out channel power response on the m th subcarrier by solving the minimization:

$$\min_{\underline{\mathbf{H}}_m=\mathbf{S}^{-1}*\underline{\mathbf{y}}}\|\underline{\mathbf{H}}_{est,m}^P-\underline{\mathbf{H}}_m^P\|^2 \tag{12}$$

in which all possible data symbol vectors $\mathbf{S_m}$, corresponding to channel gain $\underline{\mathbf{H}}_m$ as given in (11), have to be considered in the minimization problem. Here the choice for the *P*-th power depends on the signal constellation. For BPSK *P*=2, and for QPSK, *P*=4. $\underline{\mathbf{H}}_m^P$ denotes the 4x1 channel power vector with each component being taken to its *P*-th power.

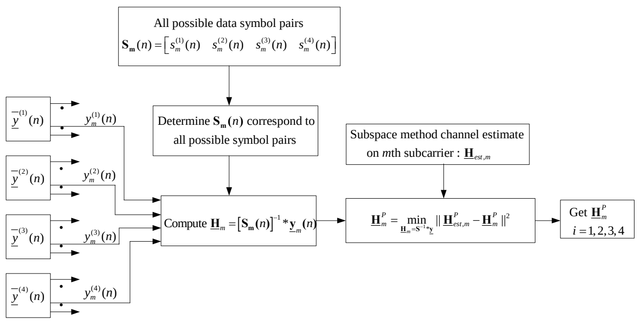

Fig. 3. Channel power response estimation in four-antenna STBC OFDM

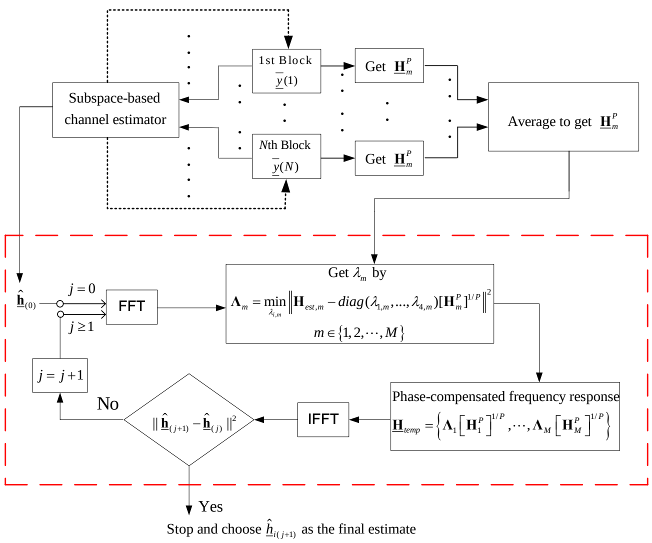

Fig. 4. Phase Direct in four- antenna STBC OFDM

After receiving a total of N data blocks in this algorithm, we can have the time-averaged channel power response**.** The phase ambiguity can be solved by

$$\mathbf{\Lambda}_m = \min_{\lambda_{i,m}} \left\| \mathbf{H}_{est,m} - diag(\lambda_{1,m},\ldots,\lambda_{4,m})[\mathbf{H}_m^P]^{1/P} \right\|^2 \tag{14}$$

where $\lambda_{i,m} \in \left\{ e^{j2\pi k/P} \mid k = 1,2,\cdots,P \right\}$ on the m th subcarrier for i i th antenna. A 4xM spatial-frequency channel response matrix becomes:

$$\underline{\mathbf{H}}_{temp} = \left\{ \mathbf{\Lambda}_1 \left[\mathbf{H}_1^P \right]^{1/P}, \cdots, \mathbf{\Lambda}_M \left[\mathbf{H}_M^P \right]^{1/P} \right\} \tag{15}$$

where M is the number of subcarriers. After IFFT, we can perform a denoising process by truncating the time domain channel $\hat{\underline{h}}$. With FFT, a new channel frequency response can replace the initial subspace-based estimate $\mathbf{H}_{est,m}$, which is repeated until $\| \hat{\underline{\mathbf{h}}}_{(j+1)} - \hat{\underline{\mathbf{h}}}_{(j)} \|^2$ converges as shown in Fig. 4.

4. PERFORMANCE ANALYSIS

The theoretical mean square error of channel estimation for the subspace-based method [2] can be written as.

$$E(\| \hat{\underline{\mathbf{h}}} - \underline{\mathbf{h}} \|^2) = E(\| \triangle \underline{\mathbf{h}} \|^2) \approx \frac{\sigma_n^2 \cdot \| \mathbf{Q}^+ \|^2}{\sigma_s^2 N} \tag{20}$$

where

$$\mathbf{Q} = \begin{bmatrix} \tilde{\mathbf{U}}_q & \tilde{\mathbf{U}}_{nq} \end{bmatrix} \begin{bmatrix} \tilde{\sum}_q & \mathbf{0} \\ \mathbf{0} & \tilde{\sum}_{nq} \end{bmatrix} \begin{bmatrix} \tilde{\mathbf{V}}_q^H \\ \tilde{\mathbf{V}}_{nq}^H \end{bmatrix} \tag{21}$$

and

$$\mathbf{Q}^+ = \tilde{\mathbf{V}}_q \left(\tilde{\sum}_q \right)^{-1} \tilde{\mathbf{U}}_q^H \tag{22}$$

σ_s^2 / σ_n^2 is SNR. $\mathbf{Q}$ is an important information matrix comprising singular vectors of the received data symbol $\underline{\mathbf{y}}$ [4].

When the diagonal weighting constant k is larger, the diagonal elements of $\tilde{\Sigma}_q$ becomes larger and $(\tilde{\Sigma}_q)^{-1}$ in (22) will become smaller, which will lead to a smaller estimated mean square error in (20).

However, an increased diagonal weight k also increases the transmitted power as well. For same SNR, the noise power σ_n^2 at the receiver needs to be increased, and bit error rate for symbol detection will degrade accordingly.

From Eq. (3), replacing true $\underline{\mathbf{h}}$ with $\hat{\underline{\mathbf{h}}} = \underline{\mathbf{h}} + \triangle \underline{\mathbf{h}}$, we have

$$\underline{\mathbf{y}} = \mathbf{S} * \hat{\underline{\mathbf{h}}} + \underline{\mathbf{n}} = \mathbf{S} * \underline{\mathbf{h}} + (\mathbf{S} * \triangle \underline{\mathbf{h}} + \underline{\mathbf{n}}) \tag{23}$$

We can see that correct detection of the transmitted symbol s from the received data $\underline{\mathbf{y}}$ depends on the channel error $\triangle \underline{\mathbf{h}}$ and noise. Suppose they are independent, the equivalent overall perturbation power can be written as

$$P_{\text{total}}(k) = P_{\mathbf{S}*\triangle \underline{\mathbf{h}}}(k) + P_{\underline{\mathbf{n}}}(k) \tag{24}$$

which is a function of the weighting constant k.

Now we will examine how the increased diagonal weight k decreases the channel estimation error $\triangle \underline{\mathbf{h}}$ but also increases the noise $\underline{\mathbf{n}}$. In case of BPSK with P=2, we note that $E\left[\mathbf{S}^H \cdot \mathbf{S} \right] = \left(k^2 + 3 \right) \cdot I$, then the equivalent channel estimation error power can be shown to be

$$\begin{aligned} P_{\mathbf{S}*\triangle \underline{\mathbf{h}}}(k) &= E\left[| \mathbf{S} \cdot \triangle \underline{\mathbf{h}} |^2 \right] = \left(k^2 + 3 \right) \cdot E(\| \triangle \underline{\mathbf{h}} \|^2) \\ &\approx \left(k^2 + 3 \right) \cdot \frac{\sigma_n^2 \| \mathbf{Q}^+(k) \|^2}{\sigma_s^2 N} \end{aligned} \tag{25}$$

Notice that $\| \mathbf{Q}^+(k) \|^2$ decreases as k increases**.** Assume that all channels are normalized and uncorrelated, $h_i^* * h_j = \delta_{i,j}$, then the average noise power can be shown to be:

$$P_{\underline{\mathbf{n}}}(k) = \frac{k^2 + 3}{4} \cdot \frac{\sigma_n^2}{\sigma_s^2} \tag{26}$$

where σ_s^2 / σ_n^2 is the signal-to-noise ratio.

Finally we have

$$P_{\text{total}}(k) = (k^2 + 3) \cdot \frac{\sigma_n^2}{\sigma_s^2} \cdot \left[\frac{\| \mathbf{Q}^+(k) \|^2}{N} + \frac{1}{4} \right] \tag{27}$$

It is now clear that the choice of the diagonal weighting constant k is a tradeoff between channel estimation error in (25) and noise in (26).

5. COMPUTER SIMULATION

Our simulation parameters are: the block size N=100, no. of subcarriers M=32, and four independent 5-ray Rayleigh fading channels. We compare the theoretical and simulated normalized mean square channel errors (NMSCE) for BD STBC BPSK system using the subspace-based channel estimation in Fig. 5. We can see that a large weighting constant *k* improves channel estimation and the simulated curves approach theoretical ones in (20) at high SNR.

Fig. 6 shows the improved NMSCE performance if the subspace-based method is further enhanced by the phase direct technique.

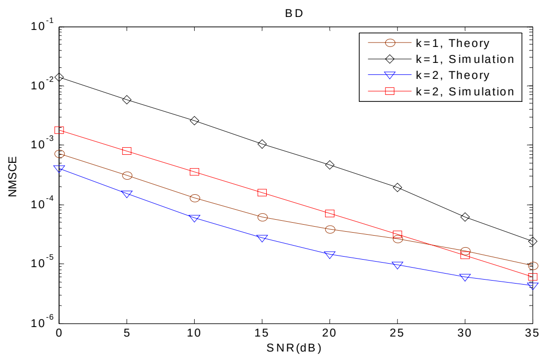

Fig. 5. Theoretical and simulated NMSCE for subspace-based channel estimation

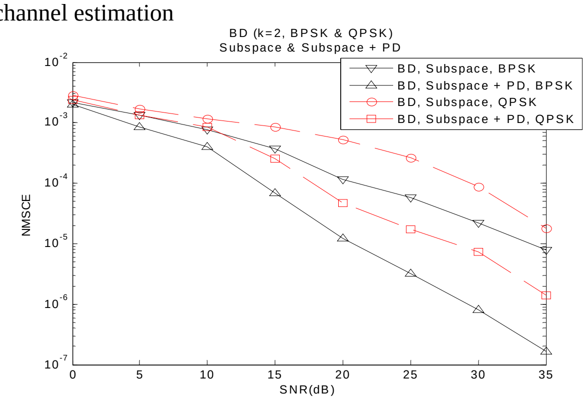

Fig. 6. Improved performances by the phase-direct technique

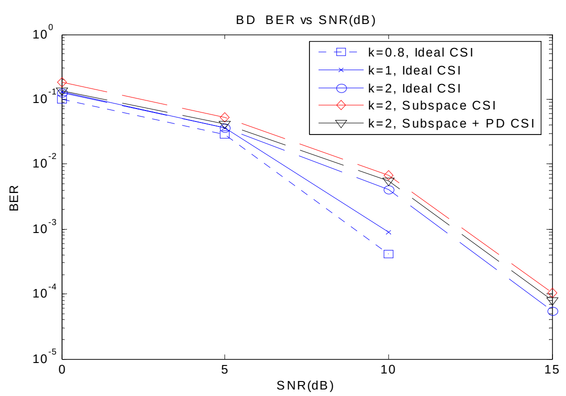

Fig. 7. Bit error rate comparison when k =0.8, 1, 2.

Fig. 7 compares bit error rates when k =0.8, 1, 2. If the channel state information (CSI) is ideal, a larger k degrades the bit error rate. In the case of *k=2*, again, we can see the phase-direct technique approaches the ideal CSI case.

The results of simulated noise power $P_{\underline{\mathbf{n}}}(k)$, equivalent channel estimation error power $P_{\mathbf{S}^*\Delta\underline{\mathbf{h}}}(k)$, both simulated and theoretical total perturbation powers $P_{\text{total}}(k)$ with SNR=10, 15dB are shown in Fig. 8. We can see that as *k* increases, the noise curve increases too, while the equivalent channel estimation error decreases, and the total perturbation power seems to be dominated by the noise power. It appears that the weighting constant can fall into (0.1,0.5) to have the small total perturbation power.

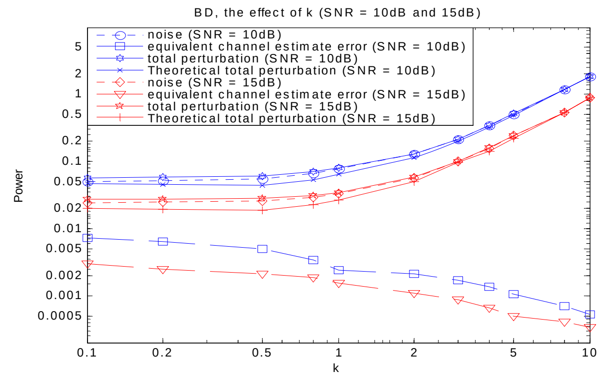

Fig. 8. Theoretical and simulated total perturbation powers

6. CONCLUSIONS

In this paper, we propose a block diagonal non-orthogonal SBTC scheme with diagonal weighting to facilitate the phase-direct channel estimation. We have derived and simulated its mean square channel estimation error and the bit error performances. Large weighting constant improves channel estimation accuracy. However, to compensate the increased transmitted signal power due to an enlarged diagonal weighted signal, the noise power is also increased so that the SNR can remain unchanged for purpose of fair comparison. In such case BER degrades. Simulation indicates a slightly weighting constant with $k \in (0.1,0.5)$ is proper in terms of a better channel estimation and symbol recovery.

7. REFERENCES

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Maximum SNR Detection for Selection-Relaying Cooperative System

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ABSTRACT

The major issue of the cooperative system is how to design a system scheme whose performance is as good as that of a ML receiver at the Destination node. Here, we propose to use a threshold-selection relay and a correction weighting at the destination. With this proposed scheme, we could improve the performance during small threshold value utilization. The theoretical BER of the proposed scheme with BPSK signals is derived and shown to be tight to the simulated results. We also use the high SNR approximation to simplify the theoretical BER from which the diversity order of the proposed scheme is shown to be 1.5~2.

Index Terms—Cooperative communication, Selective relay, diversity order

1. INTRODUCTION

As we know, the MIMO system is a very popular technology for wireless communication. However, it is impossible to place 2 or more antennas in the portable device due to the limited space. In [1]-[2], the cooperative communication emerges as a promising alternative to form a virtual MIMO system and combat fading in a wireless channels. The basic idea is that users or nodes in a wireless network share their information and transmit cooperatively as a virtual antenna array, thus providing diversity without the requirement of additional antennas at each node. There are two fundamental user cooperation protocols, i.e. Amplify-and-Forward protocol and Decode-and-Forward protocol in [3]. Here, we simply discuss about the Decode-and-Forward protocol.

There have been many works carried out in this protocol. In [3], for destination receiver scheme, they find out the optimum performance can be achieved by employing a maximum likelihood (ML) receiver rather than a maximal ratio combining (MRC) receiver at the destination. Due to the complexity of ML receiver, they develop the maximum SNR receiver can reduce the complexity but suffers performance degradation of 1~2dB as a tradeoff. In [4], each relay applies a threshold value to evaluate the signal quality. According to this quality, the relay would function well or not. This scheme could improve the MRC performance and is close to the ML performance during the large threshold value utilization. But the optimum performance of selection relay with ML receiver occurs during small threshold-value utilization.

Hence, we propose a threshold-selection relay with a correction weighting at the destination. With this scheme, we improve the performance in [4] during small threshold value utilization. We derive the theoretical Bit-Error-Rate (BER) analysis for the proposed scheme with BPSK modulation. We also develop an approximation for theoretical BER which helps us know the diversity order of the proposed scheme. Simulation results validate performance improvement of the proposed scheme.

2. SYSTEM MODEL OF PROPOSED SCHEME

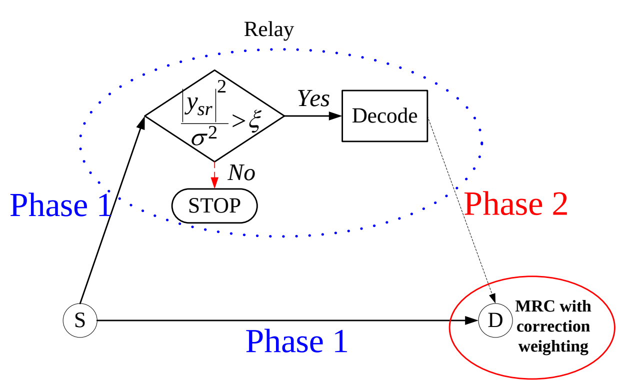

Fig. 1: Illustration of proposed cooperative system

We could use Fig. 1 to explain our proposed scheme. In Phase 1, the source (S) broadcasts the information to the destination and the relay. The received signals y_{sr} and y_{sd} can be written as

$$y_{sr} = \sqrt{P_1} h_{sr} x_s + n_{sr}, \tag{1}$$

$$y_{sd} = \sqrt{P_1} h_{sd} x_s + n_{sd} \tag{2}$$

where $x_s \in \{-1,1\}$ is a BPSK information symbol, P_1 is the transmitted power at the source, and n_{sr} and n_{sd} are additive noises. In (1) and (2), h_{sr} and h_{sd} are channel coefficients from the source to the destination and the relay respectively. In Phase 2, the relay will not work when the received signal power is lower than the threshold value. Otherwise, the relay works under the Decode-and-Forward protocol. Therefore the received signal y_{rd} can be written as

$$y_{rd} = \begin{cases} \sqrt{P_2} h_{rd} \hat{x} + n_{rd}, & |y_{sr}|^2 / \sigma^2 > \xi \\ n_{rd}, & \text{others} \end{cases} \quad (3)$$

where $\hat{x} = x$ if the relay decodes the transmitted signal correctly. Otherwise, $\hat{x} \neq x$ i.e. $\hat{x} = -x$ in case of BPSK signal. Hence, we use the method in [3] to overcome the error propagation phenomenon. We assume $\hat{x}$ could be shown as

$$\hat{x} = x_s + e \quad (4)$$

where e is a random variable that accounts for the effect of relay decision errors. With some calculations, we know

$$E[e \mid x_s] = \begin{cases} -2p & \text{if } x_s = 1 \\ 2p & \text{if } x_s = -1 \end{cases} \quad (5)$$

$$\sigma_e^2 = 4p, \quad (6)$$

where p represents the probability of relay decision error. According to (4), we know the constellation of relay transmitted signal has been changed to

$$\begin{aligned} \hat{x}_1 &= (1-p)x_1 + px_{-1} = 1-2p \\ \hat{x}_{-1} &= (1-p)x_{-1} + px_1 = -1+2p \end{aligned} \quad (7)$$

With the result of (7), we could rewrite (4) as $\hat{x} = \tilde{x}_s + \tilde{e}$, where $\tilde{x}_s = x_s + E[e \mid x_s]$ and $\tilde{e} = e - E[e \mid x_s]$. Based on these results, the received signal y_{rd} could be rewritten as

$$y_{rd} = \begin{cases} \sqrt{P_2} h_{rd} (\tilde{x}_s + \tilde{e}) + n_{rd}, & |y_{sr}|^2 / \sigma^2 > \xi \\ n_{rd}, & \text{others} \end{cases} \quad (8)$$

With (2) and (8), the weightings for relay received power large than threshold value can be shown to be

$$w_{sd} = \frac{\sqrt{P_1} h_{sd}^*}{\sigma^2} \quad w_{rd}^{'} = \frac{\sqrt{P_2} h_{rd}^* (1-2p)}{P_2 |h_{rd}|^2 \sigma_e^2 + \sigma^2} \quad (9)$$

Compared to the traditional MRC method, we could know $w_{rd}^{'}$ consists of traditional w_{rd} followed by the additional correction weighting α, i.e.

$$w_{rd} = \frac{\sqrt{P_2} h_{rd}^*}{\sigma^2} \qquad \alpha = \frac{(1-2p)}{\gamma_{rd} \sigma_e^2 + 1}. \quad (10)$$

In (10), it could be seen as a weighted-MRC method. The weighting w_{rd} would change to 0 when the relay received power is lower than threshold value. According to the above receiver weighting results, the destination receiver scheme could be shown in the Fig. 2.

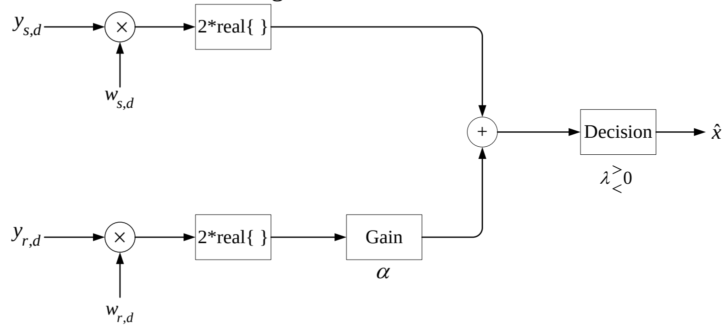

Fig. 2 The weighted-MRC receiver scheme

3. BER ANALYSIS

[4] has performed the BER analysis of the threshold selection relay with traditional MRC receiver in destination. Here, we try to analyze the BER of our proposed system by following the similar methods in [4] and [6].

We could formulate the BER of the proposed scheme by analyzing the receiver scheme in Fig. 2. Its BER depends on the relay received signal power. If the received signal power is larger than the relay threshold value, the relay would transmit the decoded signal to the destination. Therefore, the destination receiver scheme is the same as Fig. 2. On the contrary, the destination receiver scheme only works in the upper part. With these two phenomena, the BER of the proposed scheme is written as

$$P_e = P(\phi_1) P_{e|\phi_1} + P(\bar{\phi}_1) P_{e|\bar{\phi}_1} \quad (11)$$

where $P(\phi_1)$ means the probability of the event that the received signal power falls below the relay threshold value, $P_{e|\phi_1}$ means the error probability of this event. $P(\bar{\phi}_1)$ and $P_{e|\bar{\phi}_1}$ have similar meanings with $\bar{\phi}$ denoting the event that the received signal power falls above the relay threshold value.

We use (1) to derive the occurrence of these two events. Because h_{sr} and n_{sr} are complex Gaussian random variables, the received signal power $|y_{sr}|^2$ is exponentially distributive. Hence, the $P(\phi_1)$ is

$$P(\phi_1) = P\left[|y_{sr}|^2 \le \xi\sigma^2\right] = 1 - e^{-\frac{\xi\sigma^2}{P_1\sigma_{sr}^2 + \sigma^2}} \quad (12)$$

Now we proceed to derive the error probability of ϕ_1, i.e. $P_{e|\phi_1}$. Since the received signal power is lower than relay threshold value, the destination receiver only works in the upper part of Fig. 2. Hence, $P_{e|\phi_1}$ could be derived by taking the conditional expectation on the channel effect first and then averaging the effect in the end. From [7], we know

$P_{e|\phi_1}$ is known to be

$$P_{e|\phi_1} = \frac{1}{2}\left(1-\sqrt{\frac{\overline{\gamma}_{sd}}{1+\overline{\gamma}_{sd}}}\right). \quad (13)$$

Before derivation of $P_{e|\overline{\phi}_1}$, we should know the components of $P_{e|\overline{\phi}_1}$. Due to the relay error decision, the transmitted signal would have a probability p to represent the relay error decision. Hence,

$$P_{e|\overline{\phi}_1} = (1-p)P_{b1} + pP_{b2} \quad (14)$$

The derivation of p is similar to $P_{e|\phi_1}$ with different parameter γ_{sr}. The only different part occurs when averaging the channel effect. From (1), the received signal power is $P_1|h_{sr}|^2+\sigma^2$. Under the condition that the relay received power is larger than the threshold value, γ_{sr} needs to satisfy the condition $\gamma_{sr} \geq \xi - 1$. The lower limit would be shown as

$$\text{lower\_limit} = \begin{cases} 0 & \text{if } \xi \leq 1 \\ \xi - 1 & \text{if } \xi > 1 \end{cases} \quad (15)$$

With the result of (15), we could derive p as

$$\begin{aligned} p &= \frac{1}{2}\left(1-\sqrt{\frac{\overline{\gamma}_{sr}}{1+\overline{\gamma}_{sr}}}\right) \quad \xi \leq 1 \\ &= \frac{1}{2}\left[\frac{erf(\sqrt{a(b+1)})}{\sqrt{b+1}} + e^{-ba} erfc\left(\sqrt{a}\right) - \frac{1}{\sqrt{b+1}}\right] \quad \xi > 1 \end{aligned} \quad (16)$$

where $a = \xi - 1$ and $b = (\overline{\gamma}_{sr})^{-1}$. From (16), we could derive P_{b1} and P_{b2} individually. The conditional $P_{b1}^{h_{sd},h_{rd}}$ is similar to $P_{e|\phi_1}^{h_{sd}}$ with different parameters, i.e.

$$P_{b1}^{h_{sd},h_{rd}} = Q\left(\frac{\sqrt{2}(\gamma_{sd}+\alpha\gamma_{rd})}{\sqrt{\gamma_{sd}+\alpha^2\gamma_{rd}}}\right) \quad (17)$$

Similarly, $P_{b2}^{h_{sd},h_{rd}}$ could be shown to be

$$P_{b2}^{h_{sd},h_{rd}} = Q\left(\frac{\sqrt{2}(\gamma_{sd}-\alpha\gamma_{rd})}{\sqrt{\gamma_{sd}+\alpha^2\gamma_{rd}}}\right) \quad (18)$$

After averaging the channel effect of (17) and (18), P_{b1} and P_{b2} could be derived. With the result (12), (13), (16) and averaging result of (17) and (18), BER in (11) can be shown to be

$$\begin{aligned} P_e = &\left(1 - e^{-\frac{\xi\sigma^2}{P_1\sigma_w^2+\sigma^2}}\right)\left(\frac{1}{2}\left(1-\sqrt{\frac{\overline{\gamma}_{sd}}{1+\overline{\gamma}_{sd}}}\right)\right) + \\ &e^{-\frac{\xi\sigma^2}{P_1\sigma_w^2+\sigma^2}}\left(\begin{aligned}&(1-p)E\left[Q\left(\frac{\sqrt{2}(\gamma_{sd}+\alpha\gamma_{rd})}{\sqrt{\gamma_{sd}+\alpha^2\gamma_{rd}}}\right)\right] + \\ &pE\left[Q\left(\frac{\sqrt{2}(\gamma_{sd}-\alpha\gamma_{rd})}{\sqrt{\gamma_{sd}+\alpha^2\gamma_{rd}}}\right)\right]\end{aligned}\right) \end{aligned} \quad (19)$$

4. HIGH-SNR APPROXIMATION OF BER

We could simplify (11) to get the diversity order by assuming the system has a high SNR. It is well known that the diversity order is defined as the negative exponent of the average BER plotted in log-log scale. Let us parameterize three SNRs first, $\overline{\gamma}_{sd} = c_1\overline{\gamma}_s$, $\overline{\gamma}_{rd} = c_2\overline{\gamma}_s$ and $\overline{\gamma}_{rd} = c_3\overline{\gamma}_s$ where $\overline{\gamma}_s$ denotes the averaged SNR. We also assume the relay is close to the source. With this assumption, p is almost equal to zero. Therefore, the correction weighting α is very close to 1. With these approximations, (11) becomes

$$P_e \approx P(\phi_1)P_{e|\phi_1} + (P_{b1} + pP_{b2}) \quad (20)$$

By the Taylor expansion and some calculations, we have

$$\begin{aligned} P_e \approx &\frac{\xi}{(1+\overline{\gamma}_{sr})(4\overline{\gamma}_{sd})} + \\ &\left(\begin{aligned}&\left(\frac{1}{2(\overline{\gamma}_{sd}-\overline{\gamma}_{rd})}\left[-\frac{\overline{\gamma}_{sd}^{1.5}}{\sqrt{\overline{\gamma}_{sd}+1}} + (\overline{\gamma}_{sd}-\overline{\gamma}_{rd}) + \frac{\overline{\gamma}_{rd}^{1.5}}{\sqrt{\overline{\gamma}_{rd}+1}}\right]\right) \\ &+pE\left[Q\left(\frac{\gamma_{sd}-\gamma_{rd}}{\sqrt{\gamma_{sd}+\gamma_{rd}}}\right)\right]\end{aligned}\right) \end{aligned} \quad (21)$$

From (21), we only know the diversity of first term is 2 but cannot know the diversity order in second term. Therefore, we use a similar approximation method in [6] to derive the second term. With these assumptions and the Chernoff bound, P_{b1} is shown to be

$$P_{b1} \leq \frac{1}{1+\overline{\gamma}_{sd}}\frac{1}{1+\overline{\gamma}_{rd}} \propto (\overline{\gamma}_s)^{-2} \quad (22)$$

From (22), the diversity of P_{b1} is 2. We derive the diversity of p by using $Q(x) \leq \frac{1}{2}e^{-\frac{x^2}{2}}$. It is shown to be

$$p \leq \frac{1}{1+\overline{\gamma}_{sr}}e^{-d} \propto (\overline{\gamma}_s)^{-1} \quad (23)$$

where $d = 0$ or $\xi - 1$. From (23), the diversity order is 1. We also use this Q-function property to derive diversity order of P_{b2}. With $P_{b2} \le E\left[\frac{1}{2} e^{-\left(\gamma_{sd}+\gamma_{rd}-\frac{4\gamma_{sd}\gamma_{rd}}{\gamma_{sd}+\gamma_{rd}}\right)}\right]$ and the decreasing property of the exponential term, we change $\frac{4\gamma_{sd}\gamma_{rd}}{\gamma_{sd}+\gamma_{rd}}$ to $2\sqrt{\gamma_{sd}\gamma_{rd}}$. With some calculations and approximations,

$$P_{b2} \le \frac{1}{(1+\bar{\gamma}_{sd})(1+\bar{\gamma}_{rd})} + \frac{(\bar{\gamma}_{sd}\bar{\gamma}_{rd})^{1/2}}{(\bar{\gamma}_{sd}+\bar{\gamma}_{rd}+1)^{3/2}} \propto (\bar{\gamma}_s)^{-1/2} \qquad (24)$$

From (24), the diversity order is 0.5. From (21)-(24), we could conclude the diversity order of the (20) is 1.5~2. This result agrees with that in [5], i.e. for the uncoded relay, the diversity is 1.5~2 under decode-and-forward protocol.

5. SIMULATION RESULTS

Computer simulations are performed to validate our derivations. Here we assume the relay is placed in the middle. With this assumption, the channel coefficient's variance we used is the same, i.e. $\sigma_{sr}^2 = \sigma_{sd}^2 = \sigma_{rd}^2 = 1$. The additive noise is zero mean with $\sigma^2 = 1$. The total power for system is $P = P_1 + P_2$. Both transmitted power P_1 and P_2 are set to equal. We also assume the relay is placed close to source. With this assumption, we change the channel coefficient h_{sr} to 10. From Fig. 3, we know the performance of proposed scheme greatly improves the threshold-selection relay scheme and also is close to the destination with ML receiver without using any non-linear mapping function. Besides, the curve based on the theoretical BER matches closely with the simulated curve. From Fig. 4, we know the curve of the simplified BER matches those of the simulated and theoretical BERs.

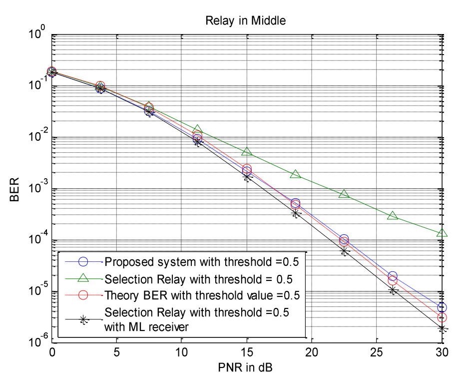

Fig. 3 Performance comparison with theoretical BER

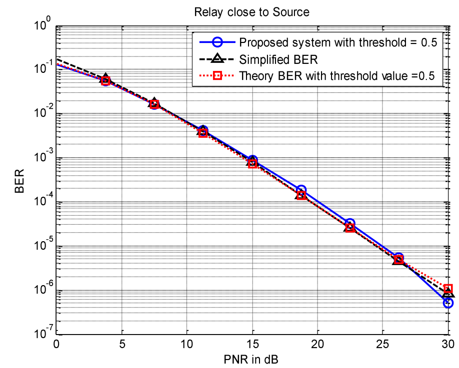

Fig. 4 Performance comparison of theoretical and approximated BERs

6. CONCLUSIONS

In this paper, we propose a threshold-selection relay with correction weighting at the destination. It improves the performance during small threshold value utilization and successfully reduces the threshold value. The optimum threshold value is 0.5 when the relay is placed in the middle. BER performance is derived theoretically and the diversity order is shown be 1.5~2.

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無衍生研發成果推廣資料

98 年度專題研究計畫研究成果彙整表

計畫主持人：謝世福			計畫編號：98-2221-E-009-094-				
計畫名稱：音響定位之研究							
成果項目			量化			單位	備註（質化說明：如數個計畫共同成果、成果列為該期刊之封面故事...等）
			實際已達成數（被接受或已發表）	預期總達成數(含實際已達成數)	本計畫實際貢獻百分比		
國內	論文著作	期刊論文	0	0	100%	篇	
		研究報告/技術報告	0	0	100%		
		研討會論文	0	0	100%		
		專書	0	0	100%		
	專利	申請中件數	0	0	100%	件	
		已獲得件數	0	0	100%		
	技術移轉	件數	0	0	100%	件	
		權利金	0	0	100%	千元	
	參與計畫人力（本國籍）	碩士生	0	0	100%	人次	
		博士生	0	0	100%		
		博士後研究員	0	0	100%		
		專任助理	0	0	100%		
國外	論文著作	期刊論文	0	0	100%	篇	
		研究報告/技術報告	0	0	100%		
		研討會論文	0	2	100%		
		專書	0	0	100%	章/本	
	專利	申請中件數	0	0	100%	件	
		已獲得件數	0	0	100%		
	技術移轉	件數	0	0	100%	件	
		權利金	0	0	100%	千元	
	參與計畫人力（外國籍）	碩士生	0	2	100%	人次	
		博士生	0	0	100%		
		博士後研究員	0	0	100%		
		專任助理	0	0	100%		

其他成果 (無法以量化表達之成果如辦理學術活動、獲得獎項、重要國際合作、研究成果國際影響力及其他協助產業技術發展之具體效益事項等，請以文字敘述填列。)	無

	成果項目	量化	名稱或內容性質簡述
科教處計畫加填項目	測驗工具(含質性與量性)	0	
	課程/模組	0	
	電腦及網路系統或工具	0	
	教材	0	
	舉辦之活動/競賽	0	
	研討會/工作坊	0	
	電子報、網站	0	
	計畫成果推廣之參與（閱聽）人數	0	

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請就研究內容與原計畫相符程度、達成預期目標情況、研究成果之學術或應用價值（簡要敘述成果所代表之意義、價值、影響或進一步發展之可能性）、是否適合在學術期刊發表或申請專利、主要發現或其他有關價值等，作一綜合評估。

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2. 研究成果在學術期刊發表或申請專利等情形： 論文：□已發表 □未發表之文稿 ■撰寫中 □無 專利：□已獲得 □申請中 ■無 技轉：□已技轉 □洽談中 ■無 其他：（以 100 字為限）
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