Signal Acquisition and Tracking with Adaptive Arrays in the Digital Mobile Radio System IS-54 with Flat Fading

Jack H. Winters, *Senior Member*

***Abstract*— This paper considers the dynamic performance of adaptive arrays in wireless communication systems. With an adaptive array, the signals received by multiple antennas are weighted and combined to suppress interference and combat desired signal fading. In these systems, the weight adaptation algorithm must acquire and track the weights even with rapid fading. Here, we consider the performance of the Least-Mean-Square (LMS) and Direct Matrix Inversion (DMI) algorithms in the North American digital mobile radio system IS-54. We show that implementation of these algorithms permits the use of coherent detection, which improves performance by 1 dB over differential detection. Results for two base station antennas with flat Rayleigh fading show that the LMS algorithm has large tracking loss for vehicle speeds above 20 mph, but the DMI algorithm can acquire and track the weights to combat desired signal fading at vehicle speeds up to 60 mph with less than 0.2 dB degradation from ideal performance with differential detection. Similarly, interference is also suppressed with performance gains over maximal ratio combining within 0.5 dB of the predicted ideal gain.**

I. Introduction

Antenna arrays with optimum combining reduce the effects of multipath fading of the desired signal and suppress interfering signals, thereby increasing both the performance and capacity of wireless systems. To be practical, though, the implemented combining algorithms must be able to rapidly acquire and track the desired and interfering signals.

Most previous theoretical and computer simulation studies of the increase in performance and capacity with optimum combining, e.g., [1]–[6], assumed ideal tracking of the desired and interfering signals. In the computer simulation study where block-by-block adaptation was considered [7], the data rate was at least 5 orders of magnitude greater than the fading rate. Although this is appropriate for the indoor radio system studied in [7] which used kbps data rates at 900 MHz, digital mobile radio systems have a much lower data-to-fading-rate ratio. For example, in the North American digital cellular system IS-54 [8] with a data rate of 24.3 ksymbols/s in the 800 MHz band, at 60 mph the data-to-fading ratio is only 300, while in the Western European GSM [8] it is around 2 000. In a previous experiment [4]–[6] that demonstrated the feasibility of optimum combining with a three-fold increase in capacity (suppression of two equal-power interferers with eight antennas), the Least-Mean-Square (LMS) algorithm tracked these signals with data-to-fading-rate ratios as low as 25. However, the tracking error loss could not be measured because of A/D quantization noise. Furthermore, this experimental system had many more antennas than interferers, which is not typical of most wireless systems.

Here we consider the dynamic performance of adaptive arrays in wireless communication systems. Specifically, we consider the performance of the LMS and Direct Matrix Inversion (DMI) algorithms in tracking the desired and interfering signals in the digital mobile radio system IS-54. We show that implementation of these algorithms permits the use of coherent detection, which improves performance by 1 dB over differential detection. Results for two base station antennas and flat Rayleigh fading show that the LMS algorithm has large tracking loss at speeds above 20 mph. However, the DMI algorithm can acquire and track the weights to combat desired signal fading at vehicle speeds up to 60 mph with less than 0.2 dB degradation from the ideal (perfect tracking) performance of optimum combining with differential detection. Similarly, interference is also suppressed with performance gains over maximal ratio combining within 0.5 dB of the predicted ideal gain.

In Section II, we determine the performance of optimum combining with ideal signal tracking. In Section III we study the performance of the LMS and DMI algorithms for acquisition and tracking of the signals in IS-54. A summary and conclusions are presented in Section IV.

II. Ideal Performance

A. Weight Equation

Fig. 1 shows a block diagram of an M antenna element adaptive array. The complex baseband signal received by the ith element in the kth symbol interval $x_i(k)$ is multiplied by a controllable complex weight $w_i(k)$. The weighted signals are then summed to form the array output $s_o(k)$. The output signal is subtracted from a reference signal $r(k)$ (described in Section III) to form an error signal $\epsilon(k)$. Weight generation circuitry determines the weights from the received signals and the error signal. In this paper, we are interested in determining the weights that minimize the mean-square error, i.e., $|\epsilon^2(k)|$.

Let the weight vector $\boldsymbol{w}$ be given by

$$\boldsymbol{w} = [w_1 w_2 \cdots w_M]^T \tag{1}$$

Manuscript received September 8, 1992; revised October 26, 1992.
The author is with AT&T Bell Laboratories, Holmdel, NJ 07733.
IEEE Log Number 9211016.

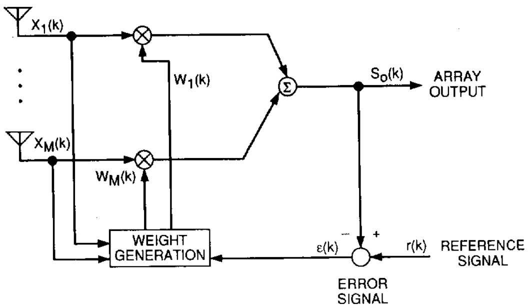


Fig. 1. Block diagram of an M element adaptive array.

where the superscript T denotes transpose, and the received signal vector $\boldsymbol{x}$ is given by

$$\boldsymbol{x} = [x_1 x_2 \cdots x_M]^T. \tag{2}$$

The received signal consists of desired signal, thermal noise, and interference and, therefore, can be expressed as

$$\boldsymbol{x} = \boldsymbol{x}_d + \boldsymbol{x}_n + \sum_{j=1}^{L} \boldsymbol{x}_j \tag{3}$$

where $\boldsymbol{x}_d$, $\boldsymbol{x}_n$, and $\boldsymbol{x}_j$ are the received desired signal, noise, and jth interfering signal vectors, respectively, and L is the number of interferers. Furthermore, let $s_d(k)$ and $s_j(k)$ be the desired and jth interfering signals, with

$$E[s_d^2(k)] = 1 \tag{4}$$

$$E[s_j^2(k)] = 1 \quad \text{for } 1 \leq j \leq L. \tag{5}$$

Then $\boldsymbol{x}$ can be expressed as

$$\boldsymbol{x} = \boldsymbol{u}_d s_d(k) + \boldsymbol{x}_n + \sum_{j=1}^{L} \boldsymbol{u}_j s_j(k) \tag{6}$$

where $\boldsymbol{u}_d$ and $\boldsymbol{u}_j$ are the desired and jth interfering signal propagation vectors, respectively.

The received signal (desired-signal-plus-interference-plus-noise) correlation matrix is given by

$$\boldsymbol{R}_{xx} = E\left[\left(\boldsymbol{x}_d + \boldsymbol{x}_n + \sum_{j=1}^{L} \boldsymbol{x}_j\right)^* \left(\boldsymbol{x}_d + \boldsymbol{x}_n + \sum_{j=1}^{L} \boldsymbol{x}_j\right)^T\right] \tag{7}$$

where the superscript * denotes complex conjugate and the expectation is taken with respect to the signals $s_d(k)$ and $s_j(k)$. Assuming the desired signal, noise, and interfering signals are uncorrelated, the expectation is evaluated to yield

$$\boldsymbol{R}_{xx} = \boldsymbol{u}_d^* \boldsymbol{u}_d^T + \sigma^2 \boldsymbol{I} + \sum_{j=1}^{L} \boldsymbol{u}_j^* \boldsymbol{u}_j^T \tag{8}$$

where σ^2 is the noise power and $\boldsymbol{I}$ is the identity matrix. Note that $\boldsymbol{R}_{xx}$ varies with the fading and that we have assumed that the fading rate is much less than the bit rate.

We define the received desired signal to noise ratio ρ as

$$\rho = \frac{E[|u_{di}|^2]}{\sigma^2} \qquad i = 1, \cdots, M \tag{9}$$

the interference-to-noise ratio (INR) as

$$\text{INR} = \frac{E[|u_{ji}|^2]}{\sigma^2} \qquad i = 1 \text{ to } M,\ j = 1 \text{ to } L \tag{10}$$

and the signal-to-noise-plus-interference ratio (SINR) as

$$\text{SINR} = \frac{\rho}{1 + \text{INR} \cdot L} \tag{11}$$

where u_{di} and u_{ji} are the ith elements of $\boldsymbol{u}_d$ and $\boldsymbol{u}_j$, respectively, and the expected value now is with respect to the propagation vectors.

The equation for the weights that minimize the mean-square error (and maximize the output SINR) is [9]

$$\boldsymbol{w} = \boldsymbol{R}_{xx}^{-1} \boldsymbol{u}_d^* \tag{12}$$

where the superscript -1 denotes the inverse of the matrix. Note that scaling of the weights by a constant does not change the output SINR. In (12), we have assumed that $\boldsymbol{R}_{xx}$ is nonsingular so that $\boldsymbol{R}_{xx}^{-1}$ exists. If not, we can use pseudoinverse techniques [10] to solve for $\boldsymbol{w}$. These optimum-combining weights are the same as those in [5], as shown in Appendix A.

B. Optimum Combiner Performance

We determine the performance of ideal optimum combining in the digital mobile radio system IS-54 in the following manner. We first determine the bit error rate (BER) with the IS-54 modulation technique, $\pi/4$-shifted differential quadrature phase shift keying (DQPSK), for ideal maximal ratio combining. With maximal ratio combining, the received signals are combined to maximize the signal-to-noise ratio at the array output, which is the optimum combining algorithm when interference is not present. Analytical results are presented for both differential detection and coherent detection, since both cases are studied in Section III. We then determine the reduction in the receive SINR required for a given BER, with optimum combining as compared to maximal ratio combining when interference is present. This gain with optimum combining is determined using analytical results with one interferer and Monte Carlo simulation with $L \geq 2$. Although these gains are generated only for coherent detection of binary PSK (BPSK), these results are also applicable to both coherent and differential detection of DQPSK, but at different BER's. This is because, for given receive SINR, the output SINR with optimum combining is independent of the modulation and detection technique, as can be seen from the equations in Section II-A. Thus the gain with optimum combining and BPSK at a given receive SINR will be similar to that of DQPSK at the same receive SINR—only the corresponding BER will be different. Note that IS-54 specifies a maximum BER of 2×10^{-2}, and, therefore, our results are generated for BER's around 10^{-2}.

With coherent detection of DQPSK and maximal ratio combining, the average BER with flat Rayleigh fading is approximately given by

$$\overline{BER} = 2P_E - P_E^2 \tag{13}$$

where (from [11])

$$P_E = 2^{-M}\left(1 - \sqrt{\frac{\rho}{2+\rho}}\right)^M \cdot \sum_{k=0}^{M}\binom{M-1+k}{k}2^{-k}\left(1+\sqrt{\frac{\rho}{2+\rho}}\right)^k. \tag{14}$$

With differential detection of DQPSK and maximal ratio combining, the average BER with flat Rayleigh fading can be shown to be given by

$$\overline{BER} = \int_0^\infty p(\gamma)P_E(\gamma)\,d\gamma \tag{15}$$

where $P_E(\gamma)$ is given by (from [12])

$$P_E(\gamma) = e^{-\gamma}\left[\frac{1}{2}I_0(\gamma/\sqrt{2}) + \sum_{k=1}^{\infty}(\sqrt{2}-1)^k I_k(\gamma/\sqrt{2})\right] \tag{16}$$

where I_k is the kth order modified Bessel function of the first kind, and $p(\gamma)$ is the probability density of the signal-to-noise ratio after maximal ratio combining and is given by [13]

$$p(\gamma) = \frac{\gamma^{M-1}e^{-\gamma/\rho}}{\rho^M(M-1)!}. \tag{17}$$

Note that with differential detection of differential BPSK (DBPSK), the average BER is $1/2(1+\rho)^{-M}$.

Fig. 2 shows the average bit error rate versus ρ (SINR with INR $= -\infty$ dB) for DQPSK and $M = 1, 2$ with maximal ratio combining. Results are also shown for DBPSK, which requires 3 dB lower ρ for the same BER with coherent detection. Note that for $M = 1$, with both DQPSK and DBPSK, differential detection requires a 0.4 dB higher ρ for a given BER than coherent detection. For $M = 2$, differential detection of DBPSK requires a 0.7 dB higher ρ than coherent detection, while differential detection of DQPSK requires a 1.0 dB higher ρ than coherent detection. Differential detection of DQPSK requires a 11.2 dB SINR for a 10^{-2} BER.

Now, let us consider the BER with ideal optimum combining. The BER with optimum combining and flat Rayleigh fading in the presence of noise only is given by the results above for maximal ratio combining. With one interferer that also experiences flat Rayleigh fading, the BER for coherent detection of BPSK is given by [1, eq. (25)]. For multiple interferers with flat Rayleigh fading, this BER can be determined by Monte Carlo simulation as described in [1].

Fig. 3 shows the gain in dB of ideal optimum combining over maximal ratio combining with two antennas for one to six interferers versus the interference-to-noise ratio (INR). This gain was determined from the reduction in the required SINR for a 10^{-3} BER at the receiver with coherent detection of BPSK. The gain occurs because optimum combining is suppressing interference in addition to increasing desired signal-to-noise ratio. A 10^{-3} BER was chosen because the

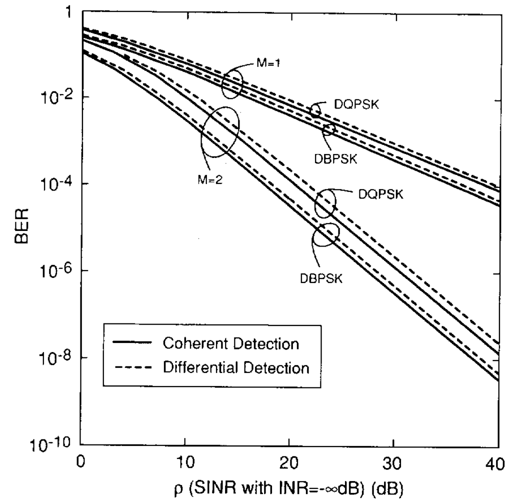


Fig. 2. Average BER versus E_b/N_0 for coherent and differential detection of DQPSK and DBPSK with $M = 1$ and 2 with flat Rayleigh fading and maximal ratio combining.

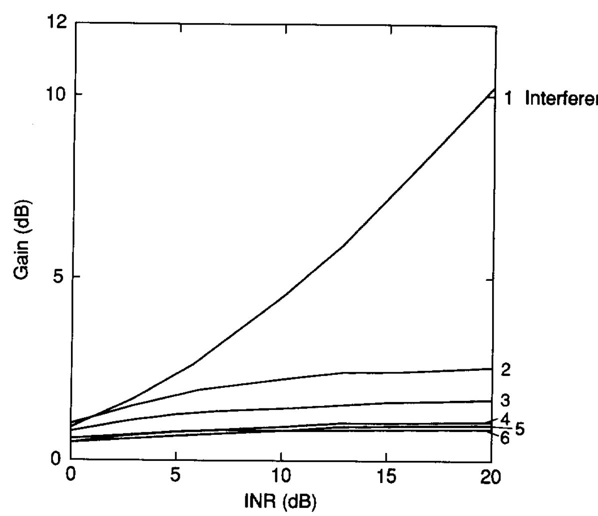


Fig. 3. Gain in dB of ideal optimum combining over maximal ratio combining with two antennas for one to six interferers versus INR at a 10^{-3} BER for coherent detection of BPSK.

11.1 dB SINR required for a 10^{-3} BER with maximal ratio combining and coherent detection of BPSK [11], [14] is close to the 11.2 dB SINR required for a 10^{-2} BER with maximal ratio combining and differential detection of DQPSK. Thus Fig. 3 also shows the gain for a 10^{-2} BER with differential detection of DQPSK. As shown in [1], the gain does not vary significantly for BER's between 10^{-2} and 10^{-3}.

With two antennas, optimum combining can completely suppress one interferer. Thus the maximum gain with optimum combining and one interferer is $10\log_{10}(10^{\mathrm{INR}/10}+1)$ dB, which is approximately INR for large INR. However, this gain can only be achieved without desired signal fading. With fading, as shown in [4]–[6], the complete suppression of one interferer results in the loss of one order of diversity against multipath fading, which corresponds to a 12.9 dB increase in the SINR required at a 10^{-3} BER [11], [14]. Thus to achieve gain, optimum combining must trade off a partial loss in diversity improvement for partial interference suppression. The resulting gain is approximately half (in dB) the maximum gain possible without desired signal fading. With more than one interferer and two receive antennas, the gain is seen to be much lower. However, the gain is almost 1 dB even with six interferers.

III. Performance of LMS and DMI in IS-54

In the digital mobile radio system IS-54, the frequency reuse factor (number of channel frequency sets) is 7. However, as shown in [6], it may be possible to reduce the frequency reuse factor to 4 (nearly doubling the system capacity) through the use of optimum combining of the signals from the two existing receive base station antennas. However, for this result in [6], we assumed ideal optimum combining, i.e., perfect tracking of the desired and interfering signals by the combining algorithm at the base station. Below, we consider the dynamic performance of optimum combining in IS-54.

A. Weight Generation

The weights can be calculated by a number of techniques. Here, we will consider two techniques: the Least Mean Square (LMS) and the Direct Matrix Inversion (DMI) algorithm [9]. For digital implementation of the LMS algorithm, the weight update equation is given by

$$\boldsymbol{w}(k+1) = \boldsymbol{w}(k) + \mu \boldsymbol{x}^*(k)\epsilon(k) \tag{18}$$

where μ is a constant adjustment factor, $x(k)$ is the received signal vector in the kth bit interval, and the error is given by

$$\epsilon(k) = r(k) - s_0(k) \tag{19}$$

where

$$s_0(k) = \boldsymbol{w}^T \boldsymbol{x}(k). \tag{20}$$

With DMI, the weights are given by [9]

$$\boldsymbol{w} = \hat{\boldsymbol{R}}_{xx}^{-1} \boldsymbol{r}_{xd} \tag{21}$$

where the estimated receive signal correlation matrix is given by

$$\hat{\boldsymbol{R}}_{xx} = 1/K \sum_{j=1}^{K} \boldsymbol{x}^*(j)\boldsymbol{x}^T(j) \tag{22}$$

where K is the number of samples used, and the estimated reference signal correlation vector is given by

$$\hat{\boldsymbol{r}}_{xd} = 1/K \sum_{j=1}^{K} \boldsymbol{x}^*(j) r(j). \tag{23}$$

Note that, as before, we have assumed that $\hat{\boldsymbol{R}}_{xx}$ is nonsingular. If not, pseudoinverse techniques can be used [10].

The LMS algorithm is the least computationally-complex weight adaptation algorithm. However, the rate of convergence to the optimum weights depends on the eigenvalues of $\boldsymbol{R}_{xx}$, i.e., on the power of the desired and interfering signals [9]. Thus weaker interference will be acquired and tracked at a slower rate than the desired signal, and the desired signal will be tracked at a slower rate during a fade (when accurate tracking is most important).

The DMI algorithm is the most computationally-complex algorithm because it involves matrix inversion. However, DMI has the fastest convergence, and the rate of convergence is independent of the eigenvalues of $\boldsymbol{R}_{xx}$, i.e., signal power levels. One issue with the DMI algorithm is its modification for tracking time-varying signals. Here we consider calculating the weights at each symbol interval using one of two data weighting functions: 1) a sliding window (fixed K in (22) and (23)) and 2) an exponential forgetting function on $\hat{\boldsymbol{R}}_{xx}$ and $\hat{\boldsymbol{r}}_{xd}$, namely,

$$\hat{\boldsymbol{R}}_{xx}(k+1) = \beta \hat{\boldsymbol{R}}_{xx}(k) + \boldsymbol{x}^*(k)\boldsymbol{x}^T(k) \tag{24}$$

$$\hat{\boldsymbol{r}}_{xx}(k+1) = \beta \hat{\boldsymbol{r}}_{xd}(k) + \boldsymbol{x}^*(k) r(k) \tag{25}$$

where β is the forgetting factor.

For $M = 2$, the DMI algorithm has about the same computational complexity as the LMS algorithm. In particular, weight calculation from the inversion of the 2×2 correlation matrix (21) does not even require division by the determinant, since this is only a weight scale factor that does not affect the output SINR. For larger M, since the complexity of matrix inversion grows with M^3 (versus M for LMS), DMI becomes very computation intensive. However, the matrix inversion can be avoided by using recursive techniques based on least-square estimation or Kalman filtering methods [9], which greatly reduce complexity (to the order of M^2) but have performance that is similar to DMI [9]. Similarly, pseudoinverse techniques [10] can be used if $\hat{\boldsymbol{R}}_{xx}^{-1}$ does not exist. Therefore, our performance results for DMI should also apply to these recursive techniques.

Next, consider reference signal generation. Since this signal is used by the adaptive array to distinguish between the desired and interfering signals, it must be correlated with the desired signal and uncorrelated with any interference. Now, the digital mobile radio system IS-54 [8] uses time division multiple access (TDMA), with three user signals in each channel and each user transmitting two blocks of 162 symbols in each frame. For mobile to base transmission, each block includes a 14-symbol synchronization sequence starting at the 15th symbol. This sequence is common to all users in a given time slot (block), but is different for each of the six time slots per frame. Since base stations operate asynchronously, signals from other cells have a high probability of having different timing (since there are 972 symbols per frame) and being uncorrelated with the sequence in the desired signal. Thus as proposed in [6], for weight acquisition we will use the known 14-symbol synchronization sequence as the reference signal. DMI is used to determine the initial weights using this sequence, since accurate initial weights are required. Note that the weights must be reacquired for each block, because with a 24.3 ksymbols/s data rate and fading rates as high as 81 Hz, the fading may change completely between blocks received from a given user. After weight acquisition, the output signal consists mainly of the desired signal, and (during proper operation) the data is detected with a bit error rate that is not more than 10^{-2} to 10^{-1}. Thus we can use the detected data as the reference signal, using either the LMS or DMI algorithm for tracking.[1] In our simulation results shown below, we did not consider the effect of data errors on the reference signal; i.e., the reference signal symbols were the same as the transmitted symbols.

Note that since the modulation technique is DQPSK, the error of interest is only the relative phase between adjacent symbols, rather than the error vector $r(k) - s_o(k)$ in (19). Indeed, the LMS algorithm can use the phase error of each symbol, i.e., $/r(k) - /s_o(k)$, where $/y$ is the phase of y, as the error signal.[2] This results in no amplitude control of $s_o(k)$, but the amplitude is not used for DQPSK detection anyway. However, we found better tracking with the error vector (19) and, therefore, used (19) for our results shown below. Note that with the DMI algorithm we do not have the option of using the phase error—we must use the error vector (19).

B. Results

To determine the performance of the acquisition and tracking algorithms in IS-54, we used IS-54 computer simulation programs written by S. R. Huszar and N. Seshadri. We modified the transmitter, fading simulator, and receiver programs for flat Rayleigh fading with one interferer and added our optimum combining algorithms with both coherent and differential detection. Specifically, the transmitted desired signal consisted of blocks of 162 symbols with $\pi/4$-shifted DQPSK modulation. The symbols in each block were randomly generated 2-bit symbols for symbols 1–14 and 29–162, and a synchronization sequence for symbols 15–28. This signal, sampled at 8 times the symbol rate, was filtered by a square root cosine rolloff filter with a rolloff factor of 0.35. For the interfering signal, randomly generated symbols, independent of the desired signal symbols, were used for the data, and a synchronization sequence that is orthogonal to that to the desired signal was used for symbols 15–28. The relative timing of the interfering and desired signals was adjustable in increments of 1/8 of the symbol duration. Independent, flat Rayleigh fading for each signal at the two receive antennas was generated by multiplying each signal by a complex Gaussian random number, which varied at the fading rate [13]. The received signals were then weighted, combined, and filtered by a square root cosine rolloff filter, followed by coherent or differential detection.

Let us first consider the performance with DMI for acquisition and LMS for tracking with differential detection without interference. Fig. 4 shows the BER versus SINR for vehicle speeds of 0, 20, and 60 mph at 900 MHz, corresponding to fading rates of 0, 27, and 81 Hz. Computer simulation results are shown for the BER over 178 blocks ($\approx 28\,000$ symbols, which should be adequate for BER $> 10^{-3}$), along with theoretical results for maximal ratio combining (15). At 0 mph, the fading channel was constant over each block, but independent between blocks. Also, LMS tracking was not used at 0 mph, and thus the results show the accuracy of weight acquisition by DMI. DMI is seen to have less than 1-dB implementation loss for BER's between 10^{-3} and 10^{-1}. At 20 and 60 mph, the tracking performance of the LMS algorithm is poor. For SINR below 14 dB, the LMS algorithm tracks so poorly that the best BER is obtained with $\mu = 0$, i.e., if LMS tracking is not used. This lack of tracking causes little degradation at 20 mph, but a several dB loss in performance at 60 mph. For SINR above 14 dB, the LMS algorithm improves performance, with the best μ equal to 0.08. At 20 mph, the performance with the LMS algorithm is about the same as at 0 mph. However, at 60 mph there is a 4.2-dB implementation loss at 10^{-2} BER. Thus the LMS algorithm is not satisfactory for optimum combining in IS-54.[3]

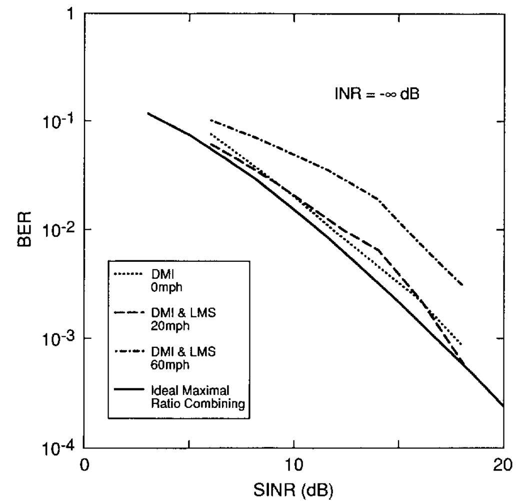


Fig. 4. BER versus SINR for vehicle speeds of 0, 20, and 60 mph with DMI for acquisition and LMS for tracking.

Next, consider DMI for both acquisition and tracking with differential detection without interference. Fig. 5 shows the average BER versus SINR with DMI and vehicle speeds of 0 and 60 mph. For these results, we used DMI with a 14-symbol sliding window ($K = 14$ in (22) and (23)), which gave us the best results for a 10^{-2} BER at 60 mph. At this BER, DMI has a negligible increase in implementation loss at 60 mph as compared to 0 mph.

Although differential detection is typically used in mobile radio because of phase tracking problems, we can also use coherent detection with optimum combining. This is because

[1] We do tracking in each block (starting from the synchronization sequence) in the forward direction for symbols 29 to 162, and in the reverse direction for symbols 14 to 1.

[2] Since DQPSK also has constant amplitude, the constant modulus algorithm can also be used to generate an error signal, i.e., $\epsilon(k) = s_o(k) - s_o(k)/|s_o(k)|$, for the LMS algorithm, as shown in [15]. A reference signal is, therefore, not needed, but this means that the receiver can acquire and track an interfering signal rather than the desired signal, and, therefore, the algorithm cannot be used for optimum combining when interference is present.

[3] In [15] it was shown that the LMS algorithm was satisfactory for diversity combining and equalization using the constant modulus algorithm for error signal generation in GSM, with data-to-fading-rate ratios as low as 1700. However, as mentioned before, this technique cannot distinguish between the desired and interfering signals.

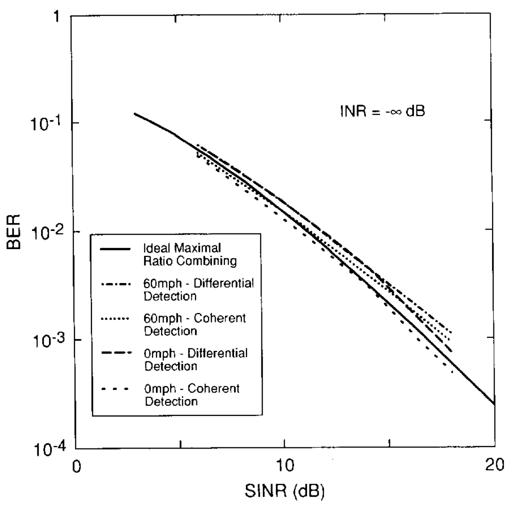


Fig. 5. BER versus SINR for vehicle speeds of 0 and 60 mph with DMI for acquisition and tracking.

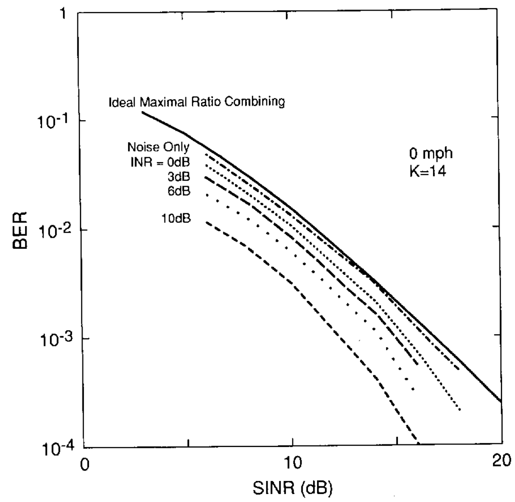


Fig. 6. BER versus SINR with one interferer for a vehicle speed of 0 mph with DMI for acquisition and tracking.

optimum combining requires coherent combining of the received signals, which means that the weights must track the received signal phase, and the array output signal phase should match the phase of the coherent reference signal. Thus coherent detection of the array output is possible, which, as shown in Section II, decreases the required SINR for a 10^{-2} BER by 1.0 dB with ideal phase tracking.[4] With the LMS algorithm, however, tracking is so poor that coherent detection is worse than differential detection. On the other hand, with the DMI algorithm, there is improvement with coherent detection. Fig. 5 shows that coherent detection decreases the required SINR for a 10^{-2} BER by 1 dB, resulting in performance that is 0.3 dB better than the theoretical performance of differential detection (but 0.7 dB worse than ideal coherent detection). At 60 mph, the performance degrades by an additional 0.5 dB; i.e., the performance is 0.2 dB worse than ideal differential detection (and 1.2 dB worse than ideal coherent detection). Thus the use of coherent rather than differential detection cancels most of the implementation loss of DMI at 60 mph.

Finally, consider the dynamic performance of optimum combining for interference suppression. For the results shown below, the symbol timing for the desired and interfering signals was the same. Our results showed that this was the worst case since there was a slight improvement in performance with timing offset between the two signals (see below).

With the LMS algorithm, even at 20 mph the performance does not improve with the INR, showing that the algorithm is not accurately tracking the interferer.

However, with DMI, the performance improvement with INR agrees with ideal tracking results. Fig. 6 shows the average BER versus SINR at 0 mph with one interferer with INR $= -\infty$, 0, 3, 6, and 10 dB. DMI with a 14-symbol sliding window and coherent detection was used as before. The required SINR for a 10^{-2} BER is 10.2, 9.5, 8.6, and 6.5 dB for INR = 0, 3, 6, and 10 dB, respectively, which is within 0.5 dB of the predicted gain shown in Fig. 3.

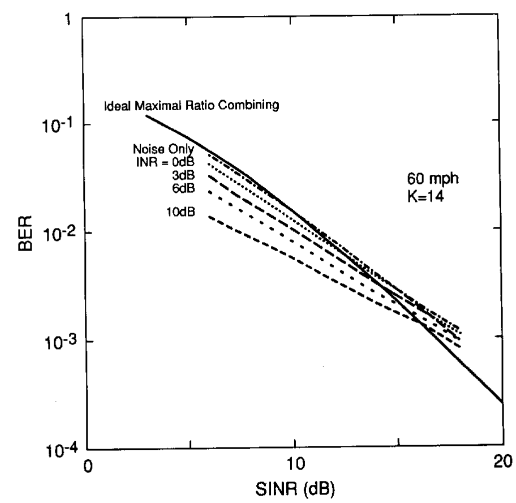


Fig. 7. BER versus SINR with one interferer for a vehicle speed of 60 mph with DMI for acquisition and tracking, and $K = 14$.

Fig. 7 shows the average BER versus SINR at 60 mph with one interferer. Again, a 14-symbol sliding window was used since this gave the best results at a 10^{-2} BER. At a 10^{-2} BER these results show a gain with INR that is within 0.5 dB of the gain shown in Fig. 3. The implementation loss increases the SINR, though, resulting in poor performance at a 10^{-3} BER with $K = 14$. However, note that the optimum window size for a given BER is determined by a tradeoff of two effects. As the window size decreases, the weights have more error due to the averaging of fewer samples, but less error caused by channel variation over the window. Our results showed that as SINR increases, the performance is improved by decreasing K.

[4] Note that this is significant in comparison to the 3.6 dB gain with optimum combining in IS-54 with 2 receive antennas [6]. Also, it is almost half of the 2.5 dB gain needed for a frequency reuse factor of 3 rather than 4 (and an additional 33% capacity increase).

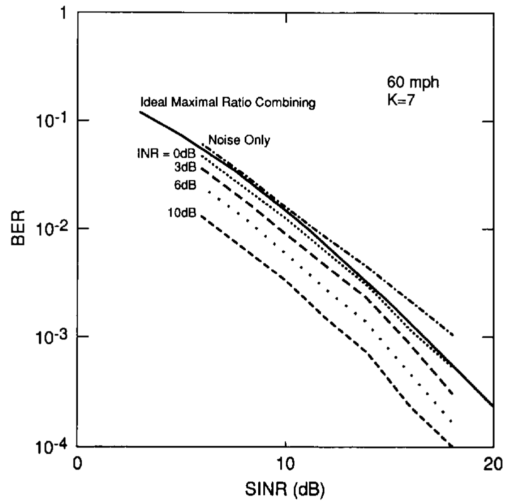


Fig. 8. BER versus SINR with one interferer for a vehicle speed of 60 mph with DMI for acquisition and tracking, and $K = 7$.

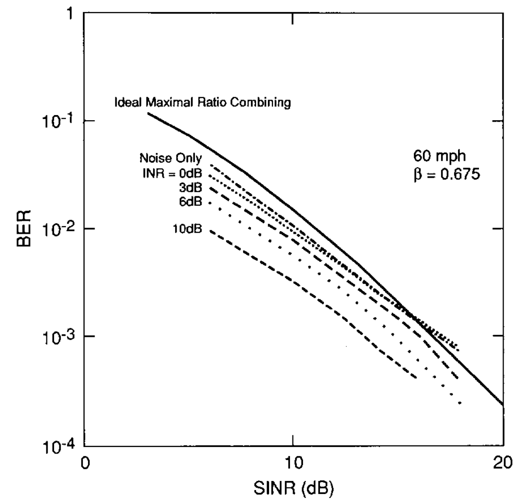


Fig. 9. BER versus SINR with one interferer for a vehicle speed of 60 mph with DMI for acquisition and tracking, and exponential weighting with $\beta = 0.675$.

Fig. 8 shows the performance with $K = 7$, which gave the best results at a 10^{-3} BER. At this BER, with interference, the improvement of optimum combining is seen to be close to that with $K = 14$ at a 10^{-2} BER (Fig. 7). Furthermore, with $K = 7$ at a 10^{-2} BER, the improvement with interference is similar to that with $K = 14$. However, with noise only, the BER for a given SINR is higher with $K = 7$ than with $K = 14$, because fewer samples are averaged to determine the weights.

Fig. 9 shows the performance of DMI with exponential weighting for $\beta = 0.675$. This β gave the best results for BER $= 10^{-2}$ and 10^{-3}. With noise only, the BER is seen to be lower than with either $K = 14$ or 7, and at a 10^{-2} BER the performance is close to that of ideal maximal ratio combining with coherent detection (i.e., 1.0 dB lower SINR than the curve shown for ideal maximal ratio combining with differential detection). With interference at a 10^{-2} BER, the gain with optimum combining is close to the predicted ideal gain; i.e., the performance is slightly better than DMI with a sliding window and $K = 14$. However, at a 10^{-3} BER with interference, the performance is slightly worse than that shown in Fig. 8 with $K = 7$. Thus either the sliding window or the exponential weighting technique can be used to generate accurately the optimum combining weights, even at 60 mph.

Finally, Fig. 10 shows the effect of timing offset between the desired and interfering signals. Results were generated for a 10 dB SINR at 60 mph with $K = 14$, as in Fig. 7. These results show that the BER varies with timing offset by less than 12% (< 0.4 dB improvement in SINR at a 10^{-2} BER), with the best performance when the interfering and desired signals are offset by approximately half the symbol duration.

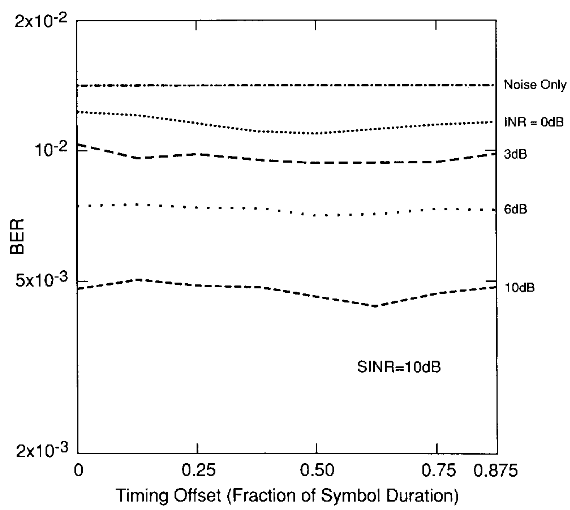


Fig. 10. Effect of timing offset on the BER for a vehicle speed of 60 mph with DMI for acquisition and tracking, $K = 14$, and SINR = 10 dB.

IV. Summary and Conclusions

In this paper, we have studied the dynamic performance of adaptive arrays in wireless communication systems. Specifically, we studied the performance of the LMS and DMI weight adaptation algorithms in IS-54 with data to fading rates as low as 300. We showed that implementation of optimum combining allows the use of coherent detection, which improves performance by over 1 dB as compared to differential detection. Although the performance of the LMS algorithm was not satisfactory, results showed that the DMI algorithm acquired the weights in the synchronization sequence interval and tracked the desired signal for vehicle speeds up to 60 mph with less than 0.2 dB degradation from the ideal performance with differential detection at a 10^{-2} BER. Similarly, an interfering signal was also suppressed with performance gains over maximal ratio combining within 0.5 dB of the predicted ideal gain. Thus our results indicate that we can obtain close to the ideal performance improvement of optimum combining even in rapidly fading environments.

ACKNOWLEDGMENT

We gratefully acknowledge useful discussions with G. D. Golden, J. Salz, and N. Seshadri.

APPENDIX A

To relate the weight equation (12) to that of [5, Eq. (11)], we need to consider three differences between the analysis given here and in [5]. First, in [4]–[6], we considered the generation of $N = L + 1$ separate outputs at the receiver, each with minimum mean-square error, while here we consider only the output of the desired signal. Using the notation of [4]–[6], the channel matrix $\boldsymbol{C}$ that relates the transmitted signal vector (including the L interferers) to the received signal vector $\boldsymbol{x}$ at a given time is given in our notation by

$$\boldsymbol{C} = [\boldsymbol{u}_d \boldsymbol{u}_1 \cdots \boldsymbol{u}_L]. \qquad \text{(A-1)}$$

Thus the weight matrix $\boldsymbol{W}$ for the optimum linear combiner that generates N output signals is given by (from (12))

$$\boldsymbol{W} = \alpha \boldsymbol{R}_{xx}^{-1} \boldsymbol{C}^* \qquad \text{(A-2)}$$

with the vector $\boldsymbol{s}$ at the output of the combiner given by

$$\boldsymbol{s} = \boldsymbol{W}^T \boldsymbol{x}. \qquad \text{(A-3)}$$

Note that the weight vector $\boldsymbol{w}$ of (12) is just the first column of $\boldsymbol{W}$. Now, we can show that

$$\boldsymbol{R}_{xx} = \sigma^2 \boldsymbol{I} + \boldsymbol{C}\boldsymbol{C}^\dagger \qquad \text{(A-4)}$$

and from (A-2),

$$\boldsymbol{W} = \alpha[\sigma^2 \boldsymbol{I} + \boldsymbol{C}\boldsymbol{C}^\dagger]^{-1} \boldsymbol{C}^*. \qquad \text{(A-5)}$$

A second difference is that in [4]–[6] we considered the zero-forcing weights, which can be obtained from (A-5) in the limit, $\sigma^2 \to 0$, i.e.,

$$\boldsymbol{W} = \alpha[\boldsymbol{C}\boldsymbol{C}^\dagger]^{-1} \boldsymbol{C}^*. \qquad \text{(A-6)}$$

Note that $[\boldsymbol{C}\boldsymbol{C}^\dagger]^{-1}$ exists only when $N = M$. Otherwise, the inverse becomes the pseudoinverse.

Finally, the weight matrix of [5], which we will denote as $\boldsymbol{W}_{[5]}$, was defined as the transpose of the weight matrix given here, i.e.,

$$\boldsymbol{s} = \boldsymbol{W}_{[5]} \boldsymbol{x} \qquad \text{(A-7)}$$

and is given in [5] as

$$\boldsymbol{W}_{[5]} = \lim_{\sigma^2 \to 0} [\sigma^2 \boldsymbol{I} + \boldsymbol{C}^\dagger \boldsymbol{C}]^{-1} \boldsymbol{C}^\dagger. \qquad \text{(A-8)}$$

Although (A-6) and (A-8) look similar, note that $\boldsymbol{C}\boldsymbol{C}^\dagger$ (A-6) is an $M \times M$ matrix, while $\boldsymbol{C}^\dagger \boldsymbol{C}$ (A-8) is an $N \times N$ matrix. However, the weights can be shown to be equal (with a scalar multiple) in the limit $\sigma^2 \to 0$. The change in the weight equation was done to put it the form for DMI (21).

REFERENCES

[1] J. H. Winters, "Optimum combining in digital mobile radio with cochannel interference," *IEEE J. Select. Areas Commun.*, vol. SAC-2, July 1984.

[2] J. H. Winters, "Optimum combining for indoor radio systems with multiple users," *IEEE Trans. Commun.*, vol. COM-35, Nov. 1987.

[3] J. H. Winters, "On the capacity of radio communication systems with diversity in a Rayleigh fading environment," *IEEE J. Select. Areas Commun.*, vol. SAC-5, June 1987.

[4] J. H. Winters, J. Salz, and R. D. Gitlin, "The capacity of wireless communication systems can be substantially increased by the use of antenna diversity," in *Proc. 1st Int. Conf. Universal Personal Commun.*, Sept. 1992, pp. 28–32.

[5] J. H. Winters, J. Salz, and R. D. Gitlin, "The capacity increase of wireless communication systems with antenna diversity," in *Proc. 1992 Conf. Inform. Sciences Syst.*, vol. II, Mar. 18–20, 1992, pp. 853–858.

[6] J. H. Winters, J. Salz, and R. D. Gitlin, "Adaptive antennas for digital mobile radio," *Proc. IEEE Adaptive Antenna Syst. Symposium*, Melville, NY, Nov. 1992, pp. 81–87.

[7] S. A. Hanna, M. El-Tanany, and S. A. Mahmoud, "An adaptive combiner for co-channel interference reduction in multi-user indoor radio systems," in *Proc. IEEE Veh. Technol. Conf.*, St. Louis, MO, May 19–22, 1991, pp. 222–227.

[8] D. J. Goodman, "Trends in cellular and cordless communications," *IEEE Commun. Mag.*, vol. 29, pp. 31–40, June 1991.

[9] R. A. Monzingo and T. W. Miller, *Introduction to Adaptive Arrays.* New York: Wiley, 1980.

[10] A. Dembo and J. Salz, "On the least squares tap adjustment algorithm in adaptive digital echo cancellers," *IEEE Trans. Commun.*, vol. 38, pp. 622–628, May 1990.

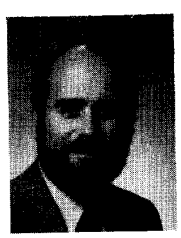
[11] P. Bello and B. D. Nelin, "Predetection diversity combining with selectively fading channels," *IRE Trans. Commun. Syst.*, vol. CS-10, p. 32, Mar. 1962.

[12] P. G. Proakis, *Digital Communications.* New York: McGraw-Hill, 1983, p. 175.

[13] W. C. Jakes, Jr., *et al.*, *Microwave Mobile Communications.* New York: Wiley, 1974.

[14] J. H. Winters, "Switched diversity with feedback for DPSK mobile radio systems," *IEEE Trans. Veh. Technol.*, vol. VT-32, pp. 134–150, Feb. 1983.

[15] T. Ohgane, "Characteristics of CMA adaptive array for selective fading compensation in digital land mobile radio communications," *Electron. Commun. Japan*, Part 1, vol. 74, no. 9, pp. 43–53, 1991.



Jack H. Winters (S'77–M'81–SM'88) received his B.S.E.E. degree from the University of Cincinnati, Cincinnati, OH, in 1977 and M.S. and Ph.D. degrees in electrical engineering from Ohio State University, Columbus, Ohio, in 1978 and 1981, respectively.

Since 1981, he has been with AT&T Bell Laboratories, where he is in the Wireless Communications Research Department. He has studied signal processing techniques for increasing the capacity and reducing signal distortion in fiber optic, mobile radio, and indoor radio systems and is currently studying adaptive arrays and equalization for mobile radio.