Optimum Combining in Digital Mobile Radio with Cochannel Interference

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Abstract—This paper studies optimum signal combining for space diversity reception in cellular mobile radio systems. With optimum combining, the signals received by the antennas are weighted and combined to maximize the output signal-to-interference-plus-noise ratio. Thus, with cochannel interference, space diversity is used not only to combat Rayleigh fading of the desired signal (as with maximal ratio combining) but also to reduce the power of interfering signals at the receiver. We use analytical and computer simulation techniques to determine the performance of optimum combining when the received desired and interfering signals are subject to Rayleigh fading. Results show that optimum combining is significantly better than maximal ratio combining even when the number of interferers is greater than the number of antennas. Results for typical cellular mobile radio systems show that optimum combining increases the output signal-to-interference ratio at the receiver by several decibels. Thus, systems can require fewer base station antennas and/or achieve increased channel capacity through greater frequency reuse. We also describe techniques for implementing optimum combining with least mean square (LMS) adaptive arrays.

I. Introduction

SPACE diversity provides an attractive means for improving the performance of mobile radio systems. With space diversity, the signals from the receiving antennas can be combined to combat multipath fading of the desired signal and reduce the relative power of interfering signals.

Previous studies of mobile radio systems (e.g., [1]) have considered space diversity only for combating multipath fading of the desired signal. Interference at each receiving antenna is assumed to be independent in these studies. Under this condition, maximal ratio combining[1] [1, p. 316] produces the highest output signal-to-interference-plus-noise ratio (SINR) at the receiver. However, in most systems (in particular, cellular mobile radio systems [1]) the same interfering signals are present at each of the receiving antennas. Thus, the received signals can be combined to suppress these interfering signals in addition to combating desired signal fading and thereby achieve higher output SINR than maximal ratio combining.

The output SINR can be maximized by using adaptive array techniques at the receiver (e.g., [2]–[4]). We will not analyze the performance of the various adaptive array techniques in this paper, but only study the performance of the optimum combiner that maximizes the output SINR. Although the adaptive array (i.e., optimum combiner) has been studied extensively, it has not been previously analyzed with the fading conditions of digital mobile radio.

This paper studies the performance of the optimum combiner in digital mobile radio systems. We assume flat Rayleigh fading across the signal channel and independent fading between antennas. The average bit error rate (BER) of the optimum combiner is studied for coherent detection of phase shift keyed (PSK) signals. Analytical and computer simulation results show that optimum combining is significantly better than maximal ratio combining even when there are more interferers than receive antennas. Results for typical cellular mobile radio systems show that the optimum combiner can increase the output SINR several decibels more than maximal ratio combining.

In Section II we describe the optimum combiner. Section III studies the BER of the optimum combiner when the desired and interfering signals are subject to Rayleigh fading. We discuss analytical results for one interferer and Monte Carlo simulation results for multiple interferers. In Section IV we consider the optimum combiner performance in cellular mobile radio systems. Section V discusses the possible methods for implementing the optimum combiner in mobile radio with a least mean square (LMS) [3] adaptive array. A summary and conclusions are presented in Section VI.

II. Optimum Combiner

A. Description and Weight Equation

Fig. 1 shows a block diagram of an M element space diversity combiner. The signal received by the ith element $y_i(t)$ is split with a quadrature hybrid into an in-phase signal $x_{I_i}(t)$ and a quadrature signal $x_{Q_i}(t)$. These signals are then multiplied by a controllable weight $w_{I_i}(t)$ or $w_{Q_i}(t)$. The weighted signals are then summed to form the array output $s_o(t)$.

The space diversity combiner can be described mathematically using complex notation [5]. Let the weight vector $\mathbf{w}$ be given by

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[1] In maximal ratio combining, the received signals are weighted proportionately to their signal-voltage-to-noise-power ratios and combined in phase.

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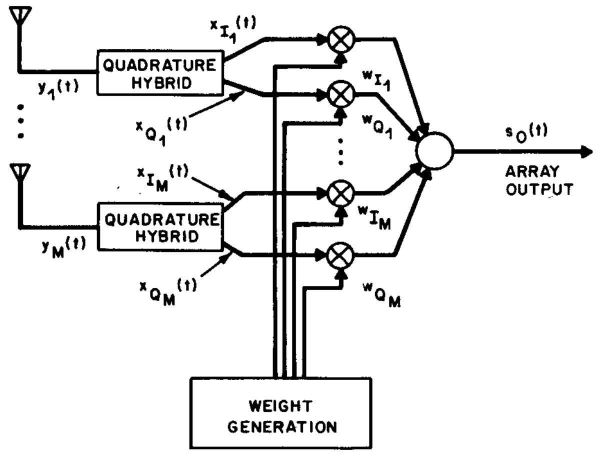

Fig. 1. Block diagram of an M element space diversity combiner.

$$\mathbf{w}=\begin{bmatrix} w_{I_1} \\ \vdots \\ w_{I_M} \end{bmatrix} - j \begin{bmatrix} w_{Q_1} \\ \vdots \\ w_{Q_M} \end{bmatrix} \tag{1}$$

and the received signal vector $\mathbf{x}$ be given by

$$\mathbf{x}=\begin{bmatrix} x_{I_1} \\ \vdots \\ x_{I_M} \end{bmatrix} + j \begin{bmatrix} x_{Q_1} \\ \vdots \\ x_{Q_M} \end{bmatrix}. \tag{2}$$

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

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

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

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

and

$$E\left[s_j^2(t)\right]=1 \qquad \text{for } 1\leqslant j\leqslant L. \tag{5}$$

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

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

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

The received interference-plus-noise correlation matrix is given by

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

where the superscripts * and T denote conjugate and transpose, respectively. Assuming the noise and interfering signals are uncorrelated, we can show that

$$\mathbf{R}_{nn}=\sigma^2\mathbf{I}+\sum_{j=1}^{L}E\left[\mathbf{u}_j^*\mathbf{u}_j^T\right] \tag{8}$$

where σ^2 is the noise power and $\mathbf{I}$ is the identity matrix. In (8) the expected value is taken over a period much less than the reciprocal of the fading rate (e.g., several bit intervals). Note that we have assumed that the fading rate is much less than the bit rate.

Finally, the equation for the weights that maximize the output SINR is (from [6])

$$\mathbf{w}=\alpha\mathbf{R}_{nn}^{-1}\mathbf{u}_d^* \tag{9}$$

where α is a constant,[2] and the superscript -1 denotes the inverse of the matrix.

B. Discussion

The mobile radio environment is quite different from the signal environment in which adaptive arrays (i.e., optimum combiners) are usually employed. In a typical adaptive array application in a nonfading environment, at the receiver there are only a few interfering signals, and their power is much greater than that of the desired signal. The adaptive array places nulls in the antenna pattern in the direction of these interferers, greatly suppressing these signals in the array output. In general, an M element array can null up to $M-2$ interfering signals and still optimize desired signal reception. The output SINR is, therefore, substantially increased by the array.

In mobile radio systems, on the other hand, at the receiver there can be several interfering signals whose power is close to that of the desired signal, and numerous interfering signals whose power is much less than that of the desired signal. Therefore, the number of interfering signals may be much greater than M, and the array may not be able to greatly suppress every interfering signal. Thus, the array output SINR may not be markedly increased by the array.

To be useful in mobile radio systems, however, the adaptive array does not have to greatly suppress interfering signals or vastly increase the output SINR. Interfering signals need only to be reduced in power by a few decibels so that their power is below the sum of the power of other interferers. Furthermore, a substantial output SINR improvement is not required because a several decibel increase in output SINR can make possible large increases in the channel capacity of the system. Thus, although the signal environment in mobile radio systems is quite different from that of the typical adaptive array system, adaptive array techniques can still offer significant advantages.

One other major difference is the ability of the array to resolve closely spaced transmitters. In a nonfading environment the adaptive array cannot suppress an interfering

[2]Note that α does not affect the array output SINR (i.e., the array performance) and, therefore, we will not consider its value.

signal if the angular separation between the interfering and desired transmiters is too small. In this case, the desired signal phase difference between receive antennas is nearly the same as that for the interfering signal. Therefore, the array cannot both null one signal and enhance reception of the other. The use of additional antennas results in a small decrease in the required angular separation, but the problem remains.

In mobile radio, because of multipath, the signal phase at one antenna is independent of the signal phase at another antenna when the antenna separation is greater than half of a wavelength (several inches at 800 MHz) [1, p. 311].[3] Therefore, the adaptive array antenna pattern is meaningless. Similarly, at the receive antennas, the signal phases from two different transmit antennas are independent when the transmit antennas are more than half of a wavelength apart. Therefore, for all practical purposes, the received signal phases are independent of vehicle location. Thus, the resolution of interfering and desired signals does not depend on how closely the vehicles are located. Instead, for all locations, there is a small probability that the array cannot resolve the two signals. This occurs when the phase differences between the antennas are nearly the same for both the desired and interfering signals. With moving mobiles,[4] however, the period of time with unresolved signals is very brief, and the performance of the adaptive array can be averaged over the fading. Furthermore, since the signal phase differences between the antennas are independent, the probability of unresolved signals with M antennas is approximately equal to the probability of unresolved signals with two antennas raised to the $M-1$ power. Thus, each additional antenna greatly decreases the probability of unresolved signals, and this probability becomes negligible with only a few antennas.

In summary, in mobile radio the adaptive array cannot resolve desired and interfering signals a small percentage of the time, rather than over a given angular separation. Therefore, we need not be concerned with vehicle location for the resolving of the signals. In the following analysis we study the optimum combiner performance averaged over the Rayleigh fading.

III. Optimum Combiner Performance with Fading

This section studies the performance of the optimum combiner when the received signals are subject to fading. For the analysis we assume flat fading across the channel, with the fading independent between antennas. We assume that the received signal has an envelope with a Rayleigh distribution and a phase with a uniform distribution. We study the steady state performance of the optimum combiner, first determining the output SINR distribution and then the BER for coherent detection of PSK. All results are compared to those for maximal ratio combining.

In cellular mobile radio, the interference plus noise at the receiver consists primarily of cochannel interference. In a typical system, there are numerous cochannel interfering signals, each of which affects the performance of the optimum combiner. An exact analysis of the performance is, therefore, quite complicated, especially since, with fading, each of these signals has a random amplitude. Therefore, in the analysis in this paper, we consider only the strongest interferers individually. The remaining interfering signals are combined and considered as lumped interference that is uncorrelated between antennas. Since with Rayleigh fading the in-phase and quadrature components of each of the received interfering signals have a Gaussian distribution, the components of the sum of these signals also have a Gaussian distribution, and the sum can be considered as thermal noise. Thus, under this assumption the combiner cannot suppress the lumped interference, and we are therefore analyzing a worse situation since the actual combiner performance will be better. Therefore, although this analysis does not show the maximum improvement (over maximal ratio combining) with optimum combining, it does show most of it. Note that if we consider all the interference as lumped interference, the results are identical to those for maximal ratio combining.

The analysis involves several parameters which are defined as follows:

$$\Gamma = \frac{\text{mean received desired signal power per antenna}}{\begin{array}{c}\text{mean received noise plus interference power}\\ \text{per antenna}\end{array}} \tag{10}$$

$$\Gamma_d = \frac{\text{mean received desired signal power per antenna}}{\text{mean received noise power per antenna}} \tag{11}$$

$$\Gamma_j = \frac{\text{mean received } j\text{th interferer signal power per antenna}}{\text{mean received noise power per antenna}} \tag{12}$$

$$\gamma_R = \frac{\text{local mean desired signal power at the array output}}{\begin{array}{c}\text{mean noise plus interference power at the}\\ \text{array output}\end{array}} \tag{13}$$

and

$$\gamma = \frac{\text{local mean desired signal power at the array output}}{\begin{array}{c}\text{local mean noise plus interference power}\\ \text{at the array output}\end{array}} \tag{14}$$

In the above definitions, mean is the average over the Rayleigh fading, and local mean is the average over a period less than the reciprocal of the fading rate (e.g., several bit durations). It is useful to note that

[3] In our analysis we assume the antennas are spaced far enough apart so that the received signal phases are independent.

[4] If all mobiles are stationary, channel reassignment can be used to eliminate the problem.

$$\Gamma = \frac{\Gamma_d}{1 + \sum_{j=1}^{L} \Gamma_j}. \tag{15}$$

A. Analytical Results with One Interferer

We first consider γ_R with one interferer. γ_R can be determined from [6, weight equation (9)]

$$\gamma_R = \mathbf{u}_d^T \mathbf{R}_{nn}^{-1} \mathbf{u}_d^* \tag{16}$$

where from (8)

$$\mathbf{R}_{nn} = \sigma^2 \mathbf{I} + E\left[\mathbf{u}_1^* \mathbf{u}_1^T\right]. \tag{17}$$

We note that with our fading model the components of $\mathbf{u}_d$ and $\mathbf{u}_1$ are complex Gaussian random variables that vary at the fading rate. Thus, to determine γ_R, the expected value in (17) must be averaged over the Rayleigh fading. It can also be seen that γ_R will vary at the fading rate.

The probability density function of γ_R can be calculated to be given by [7]

$$p(\gamma_R) = \frac{e^{-\gamma_R/\Gamma_d}\left(\frac{\gamma_R}{\Gamma_d}\right)^{M-1}(1+M\Gamma_1)}{\Gamma_d(M-2)!} \cdot \int_0^1 e^{-((\gamma_R/\Gamma_d)M\Gamma_1)t}(1-t)^{M-2}\,dt. \tag{18}$$

Thus, the cumulative distribution function is given by

$$P(\gamma_R) = \int_0^{\gamma_R/\Gamma_d} \frac{e^{-x}x^{M-1}(1+M\Gamma_1)}{(M-2)!} \cdot \int_0^1 e^{-xM\Gamma_1 t}(1-t)^{M-2}\,dt\,dx. \tag{19}$$

In the above equations it is seen that γ_R can be normalized by Γ_d. Therefore, from (15) we can also normalize γ_R by Γ and compare the performance of optimum combining to that of maximal ratio combining for fixed average received SINR.

Fig. 2 shows the cumulative distribution function of γ_R versus γ_R/Γ with optimum combining for several values of M and Γ_1. The $\Gamma_1 = 0$ distribution curve is also the distribution curve for maximal ratio combining. Fig. 2 shows that for fixed average received SINR the distribution function decreases as the interference power becomes a larger proportion of the total noise-plus-interference power. The decrease becomes even greater as M increases. Thus, optimum combining improves the receiver's performance the most when the interferer's power is large compared to the thermal noise power and there are several antennas.

As in [7] let us now consider the effect of a very high power interferer on optimum combining. Without interference (or with maximal ratio combining) $\Gamma_1 = 0$, and the cumulative distribution function is given by

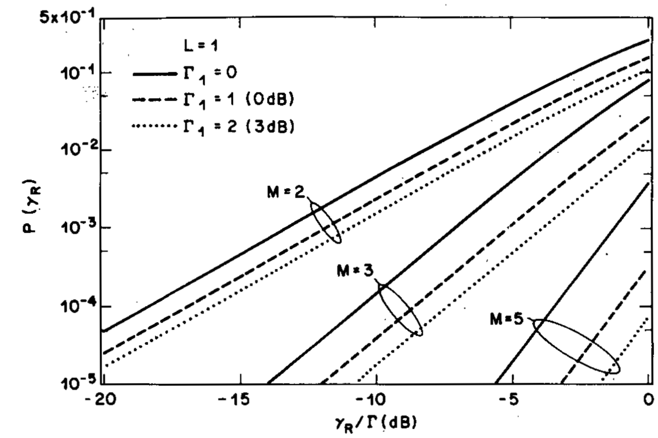

Fig. 2. The cumulative distribution function of γ_R versus γ_R/Γ for optimum combining when the desired and interfering signals are subject to fading. Results are shown for one interferer with several values of M and Γ_1. The distribution function for γ_R with fixed average received SINR is shown to decrease as the power of the interferer becomes a larger proportion of the total noise plus interference power. The decrease is even larger as M increases.

$$P(\gamma_R) = \int_0^{\gamma_R/\Gamma_d} \frac{e^{-x}x^{M-1}}{(M-1)!}\,dx \tag{20}$$

or

$$P(\gamma_R) = 1 - e^{-\gamma_R/\Gamma_d} \sum_{k=1}^{M} \frac{\left(\frac{\gamma_R}{\Gamma_d}\right)^{k-1}}{(k-1)!} \tag{21}$$

which agrees with [1, p. 319]. With a high power interferer $\Gamma_1 = \infty$, and the cumulative distribution function is given by

$$P(\gamma_R) = \int_0^{\gamma_R/\Gamma_d} \frac{e^{-x}x^{M-2}}{(M-2)!}\,dx \tag{22}$$

or

$$P(\gamma_R) = 1 - e^{-\gamma_R/\Gamma_d} \sum_{k=1}^{M-1} \frac{\left(\frac{\gamma_R}{\Gamma_d}\right)^{k-1}}{(k-1)!}. \tag{23}$$

Thus, optimum combining with an infinitely strong interferer gives the same results as maximal ratio combining without the interferer and with one less antenna. In other words, optimum combining with a strong interferer and an additional antenna will always do better than maximal ratio combining without the interferer.

For digital mobile radio this has the following impact. Let us consider a system where the required system performance can be achieved with maximal ratio combining at the base station receiver and adaptive retransmission (see Section V-B) for base-to-mobile transmission. Then, another mobile can be added per channel per cell by using optimum combining and adding one antenna at the base station. Thus, optimum combining provides a relatively simple means for growth in a system.

The BER for coherent detection of PSK is given by

$$\mathrm{BER} = \int_0^\infty p(\gamma_R) \frac{1}{2} \operatorname{erfc}\left(\sqrt{\Gamma \gamma_R}\right) d\gamma_R. \tag{24}$$

From (18) and (24) the BER for optimum combining with one interferer can be calculated to be given by [7]

$$\mathrm{BER} = \frac{(-1)^{M-1}(1+M\Gamma_1)}{2(M\Gamma_1)^{M-1}} \left\{ -\frac{M\Gamma_1}{1+M\Gamma_1} + \sqrt{\frac{\Gamma_d}{1+\Gamma_d}} - \frac{1}{1+M\Gamma_1}\sqrt{\frac{\Gamma_d}{1+M\Gamma_1+\Gamma_d}} - \sum_{k=1}^{M-2} (-M\Gamma_1)^k \cdot \left[1 - \sqrt{\frac{\Gamma_d}{1+\Gamma_d}}\left(1 + \sum_{i=1}^{k} \frac{(2i-1)!!}{i!(2+2\Gamma_d)^i}\right)\right]\right\} \tag{25}$$

where

$$(2i-1)!! = 1 \cdot 3 \cdot 5 \cdot \cdots \cdot (2i-1). \tag{26}$$

Similarly, for maximal ratio combining the BER can be determined from (21) and (24) as [8]

$$\mathrm{BER} = 2^{-M}\left(1 - \sqrt{\frac{\Gamma}{1+\Gamma}}\right)^M \sum_{k=0}^{M-1} \binom{M-1+k}{k} \cdot 2^{-k}\left(1 + \sqrt{\frac{\Gamma}{1+\Gamma}}\right)^k. \tag{27}$$

Fig. 3 shows the BER versus the average received SINR (Γ) for optimum combining with one interferer. Results are shown for several values of M and Γ_1. The results for $\Gamma_1 = 0$ are the same as those for maximal ratio combining.

Let us define the optimum combining improvement as the decrease (in decibels) in the average received SINR required for a given BER as compared to maximal ratio combining. Fig. 3 shows that the improvement is nearly independent of Γ for BER's less than 10^{-2}. Thus, for most systems of interest, the improvement is independent of Γ_d and depends only on Γ_1 and M. That is, the improvement depends on the interferer's power relative to the combined power of the other interferers and not the power of the desired signal.

In Fig. 4 the optimum combining improvement is plotted versus Γ_1 for one interferer and several values of M. The results are shown for a BER of 10^{-3}, but as discussed above, similar results can be obtained for other BER values (less than 10^{-2}). Results show that as Γ_1 and M increase, the improvement also increases.

Fig. 4 also shows the maximum improvement that can be achieved if the interferer is completely nulled in the array output. The difference between the maximum and the actual improvement for a given M ($M > 2$) is shown to be constant for large Γ_1. From (23) it can be seen that this difference is the increase in the required received SINR with the loss of one antenna with maximal ratio combining. For example, the required received SINR with maximal ratio combining is 2.3 dB for $M = 5$ and 4.0 dB for $M = 4$. Thus, for $M = 5$ the optimum combining improvement is 1.7 dB less than maximum for large Γ_1.

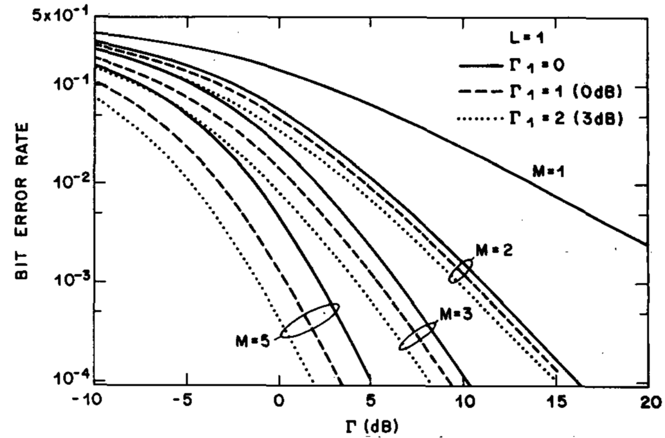

Fig. 3. The average BER versus the average received SINR for optimum combining with one interferer. Results are shown for several values of M and Γ_1. The improvement with optimum combining (in decibels) is shown to be nearly independent of the average received SINR for BER's less than 10^{-2}.

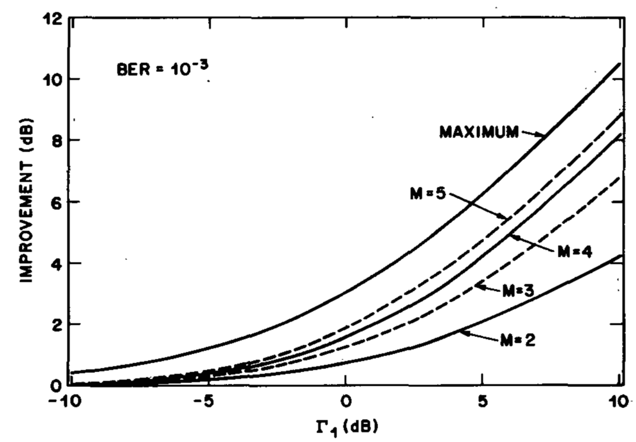

Fig. 4. The improvement of optimum combining over maximal ratio combining versus Γ_1 with one interferer for several values of M and a 10^{-3} BER. Results show that as Γ_1 and M increase, the improvement becomes significant.

B. Simulation Results with Multiple Interferers

With two or more interferers, it is extremely difficult to determine analytically the optimum combiner performance. Therefore, in this section we use Monte Carlo simulation to determine the performance of optimum combining. For the simulation we consider γ rather than γ_R as in the previous section.

For a given bit duration, the array output SINR is given by

$$\gamma = \frac{P_d}{P_{i+n}} \tag{28}$$

where P_d is the power of the desired signal and P_{i+n} is the power of the interference plus noise. The desired signal power is given by

$$P_d = \frac{1}{2}|\mathbf{w}^\dagger \mathbf{u}_d^*|^2 \tag{29}$$

where the superscript † denotes complex conjugate transpose. The power of the interference plus noise is given by

$$P_{i+n} = \frac{1}{2}\left|\mathbf{w}^\dagger\left[\mathbf{u}_n + \sum_{j=1}^{L}\mathbf{u}_j\right]^*\right|^2 \tag{30}$$

where $\mathbf{u}_n$ is the vector of the thermal noise vectors at the receiver. The components of $\mathbf{u}_n$ are independent complex Gaussian random variables with a variance corresponding to the noise power. The array output SINR is then given by

$$\gamma = \frac{|\mathbf{w}^\dagger \mathbf{u}_d^*|^2}{\left|\mathbf{w}^\dagger\left(\mathbf{u}_n + \sum_{j=1}^{L}\mathbf{u}_j\right)^*\right|^2}. \tag{31}$$

The weight vector for the optimum combiner is given in (9). As can be seen from (8), $\mathbf{R}_{nn}$ is Hermitian, and therefore, from (9)

$$\mathbf{w}^\dagger = \alpha \mathbf{u}_d^T \mathbf{R}_{nn}^{-1}. \tag{32}$$

Thus, the SINR can be expressed as

$$\gamma = \frac{|\mathbf{u}_d^T \mathbf{R}_{nn}^{-1} \mathbf{u}_d^*|^2}{\left|\mathbf{u}_d^T \mathbf{R}_{nn}^{-1}\left(\mathbf{u}_n + \sum_{j=1}^{L}\mathbf{u}_j\right)^*\right|^2}. \tag{33}$$

With Rayleigh fading, the components of $\mathbf{u}_d$, $\mathbf{u}_n$, and $\mathbf{u}_j$ are complex Gaussian random variables with zero mean and variance Γ_d, σ^2, and Γ_j, respectively. Therefore, through Monte Carlo simulation, the probability distribution of γ can be determined.

We now discuss the distribution of γ with maximal ratio combining so that a comparison to optimum combining can be made. For maximal ratio combining, the weights are given by

$$\mathbf{w} = \mathbf{u}_d^* \tag{34}$$

or

$$\mathbf{w}^\dagger = \mathbf{u}_d^T. \tag{35}$$

Therefore, from (31) the SINR is given by

$$\gamma = \frac{|\mathbf{u}_d^T \mathbf{u}_d^*|^2}{\left|\mathbf{u}_d^T\left(\mathbf{u}_n + \sum_{j=1}^{L}\mathbf{u}_j\right)^*\right|^2}. \tag{36}$$

Since the components of $\mathbf{u}_n + \sum_{j=1}^{L}\mathbf{u}_j$ are the sum of independent complex Gaussian random variables, the components are also independent complex Gaussian random variables. Thus, the probability density function of γ can be determined analytically to be given by [1, p. 367]

$$p(\gamma) = \frac{M\left(\frac{\gamma}{\Gamma}\right)^{M-1}}{\Gamma\left(1+\frac{\gamma}{\Gamma}\right)^{M+1}} \tag{37}$$

and the cumulative distribution function is given by

$$P(\gamma) = \left(\frac{\frac{\gamma}{\Gamma}}{1+\frac{\gamma}{\Gamma}}\right)^M. \tag{38}$$

The cumulative distribution of γ is plotted versus γ/Γ in Fig. 5. Simulation results with 100 000 samples are shown for optimum combining with two interferers that have 3 dB higher power than the noise, and analytical results are shown for maximal ratio combining. With 100 000 samples there are small deviations in the simulation results only for very small values of the distribution function. Fig. 5 shows that optimum combining significantly decreases the value of the distribution function as compared to maximal ratio combining. This decrease becomes even greater as M increases.

The BER can be determined from the cumulative distribution function by the equation

$$\text{BER} = \frac{1}{\pi}\int_0^1 \cos^{-1}(\sqrt{\gamma})\, p(\gamma)\, d\gamma \tag{39}$$

or

$$\text{BER} = \frac{1}{\pi}\int_0^1 \cos^{-1}(\sqrt{\gamma})\left(\frac{dP(\gamma)}{d\gamma}\right) d\gamma. \tag{40}$$

Thus, the BER for optimum combining was determined from the simulation results using the above equation. Since the cumulative distribution function can be determined for γ normalized by Γ, from one simulation run we can determine the BER over a wide range of Γ's. Similarly, the BER for maximal ratio combining is seen from (37) and (39) to be given by

$$\text{BER} = \frac{M}{\pi}\int_0^{1/\Gamma} \cos^{-1}(\sqrt{\Gamma x})\frac{x^{M-1}}{(1+x)^{M+1}}\, dx \tag{41}$$

which is numerically equivalent to (27).

For one interferer, the BER results obtained using the above equations (with a simulation using 100 000 samples) agree with the analytical results shown in Fig. 3.

For two interferers, the BER results are shown in Fig. 6. The simulation used 100 000 samples per data point. Fig. 6 shows that there is a marked improvement with optimum combining as the number of antennas increases. For example, for a BER of 10^{-3} and M equal to 5, optimum combining requires 4.2 dB less SINR than maximal ratio combining. Thus, in this case, optimum combining with five antennas (which requires -1.9 dB for a 10^{-3} BER) is

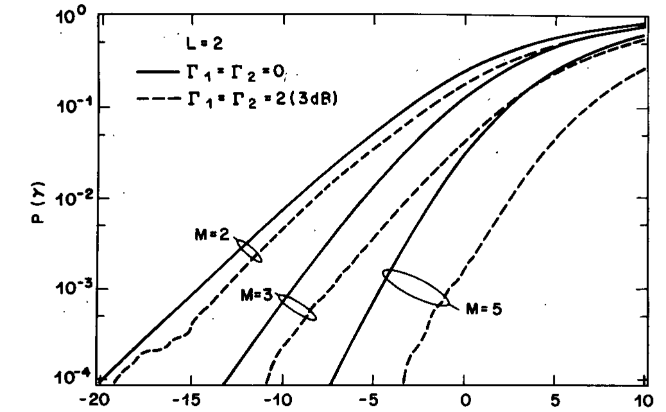

Fig. 5. The cumulative distribution function of γ versus γ/Γ for optimum combining with two interferers that have 3 dB higher power than the noise. Analytical results for maximal ratio combining ($\Gamma_1 = \Gamma_1 = 0$) are also shown. Optimum combining is seen to significantly decrease the distribution function.

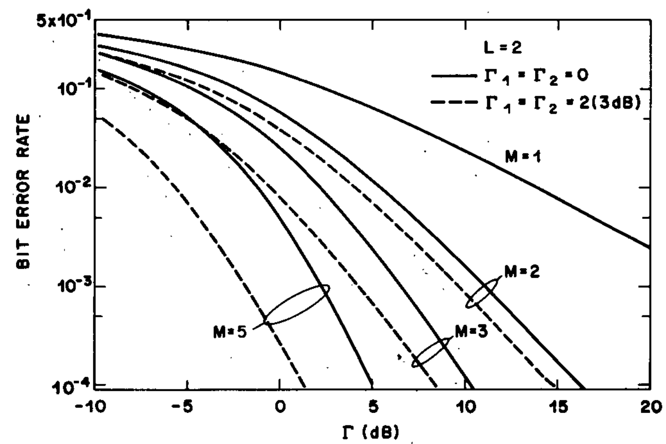

Fig. 6. The average BER versus the average received SINR for optimum combining with two interferers. Results are shown for optimum combining with two equal power interferers ($\Gamma_1 = \Gamma_2 = 2$) and for maximal ratio combining ($\Gamma_1 = \Gamma_2 = 0$). There is a marked improvement with optimum combining as the number of antennas increases.

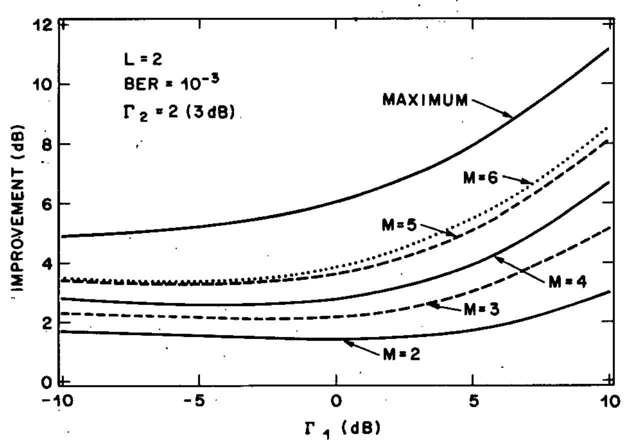

Fig. 7. The improvement of optimum combining over maximal ratio combining versus the signal-to-noise ratio of one interferer when there is also a second interferer with a 3 dB signal-to-noise ratio. The improvement is within about 2 dB of the maximum improvement with six or more antennas.

better than maximal ratio combining with nine antennas (which, from (27), requires -1.7 dB for a 10^{-3} BER).

Fig. 7 shows the optimum combiner improvement over maximal ratio combining versus the signal-to-noise ratio of one interferer when there is also a second interferer with a 3 dB signal-to-noise ratio. The simulation used 100 000 samples per data point. The results are shown for a 10^{-3} BER, but as seen in Fig. 6, these results are similar to the results for other BER's less than 10^{-2}. Fig. 7 also shows the maximum improvement possible if both interfering signals are completely nulled in the receiver output (i.e., the difference between the maximal ratio combiner performance with and without interference). The improvement is within about 2 dB of the maximum with six or more antennas.

Fig. 8 shows the improvement versus the number of antennas with one to six equal power ($\Gamma_j = 3$ dB) interferers. Again, 100 000 samples per data point were used. The improvement is shown to be between 1–6 dB as M varies from 2 to 8. Thus, optimum combining has some improvement over maximal ratio combining even with a few antennas, and the improvement greatly increases with the number of antennas.

Although the results of Fig. 8 are for equal power interferers with a particular value of Γ_j, they demonstrate the following characteristics of optimum combining that apply to other interference cases as well. First, when the number of antennas is much greater than the number of interferers, the improvement is limited. That is, in this case there is little improvement (relative to maximal ratio combining) with additional antennas. This can also be seen from Figs. 4 and 7. Second, except for the above case, the increase in the improvement (in decibels) with each additional antenna is approximately constant (about 0.6 dB for $\Gamma_j = 3$ dB). Finally, the most interesting characteristic is that there is a large improvement even when the number of interferers is greater than the number of antennas. This implies that in analyzing systems we must consider many interferers individually even if there are only a few antennas. For example, consider the case of five antennas with six interferers, each with Γ_j equal to 3 dB. From Fig. 8, the improvement is 2.7 dB. However, if only five interferers are considered individually, and the power of the sixth one is combined with the thermal noise, Γ_j is -1.8 dB and the improvement is only 1.6 dB. Thus, we must consider individually as many interferers as possible to determine accurately the actual optimum combining improvement.

IV. Performance in Typical Systems

This section studies the performance of optimum combining in typical cellular mobile radio systems. Using the techniques of Section III, we study optimum combining when the signals are subject to Rayleigh fading.[5] Optimum combining is studied only at the base station receiver because multiple antennas and the associated signal processing for optimum combining are less costly to implement at the base station than on numerous mobiles. (Adaptive retransmission with time division [1], [9] can be used to

[5] In an actual mobile radio system, the signals are also subject to shadow fading [9] which greatly complicates analysis. We therefore only consider Rayleigh fading so that system comparisons can easily be made.

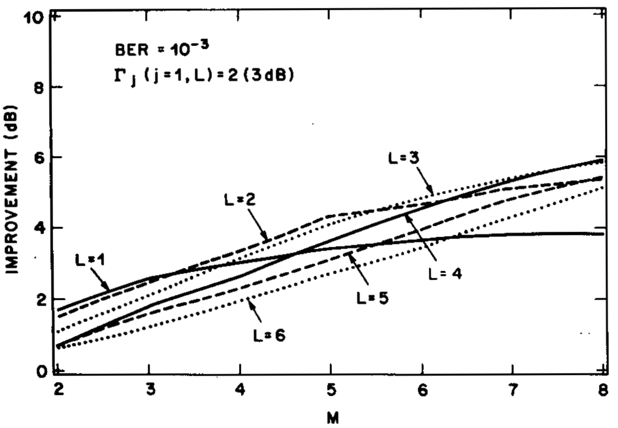

Fig. 8. The improvement of optimum combining versus the number of antennas with one to six equal power interferers. The improvement is between 1–6 dB as M varies from 2 to 8.

improve reception at the mobile with multiple base station antennas only (see Section V-B).) As before, all results for optimum combining are compared to maximal ratio combining.

Analysis of optimum combining with numerous interferers requires a substantial amount of computer time. It is therefore nearly impossible to determine the average performance of the adaptive array in the typical cellular system with random mobile locations. Therefore, in this section we consider a worst case scenario only, i.e., the mobile transmitting the desired signal is at the point in the cell farthest from the base station, and the interfering mobiles in the surrounding cells are as close as possible to the base station of the desired mobile. Furthermore, in the analysis we consider only the six strongest interferers individually. The power of the other interferers is combined and considered as thermal noise.

The systems studied involve two different cell geometries with hexagonal cells. In one geometry the base stations are located at the cell center, and in the other geometry the base stations are at the three alternate corners of the cell and are equipped with sectoral horns. In the latter geometry, each of the base station's three antennas has a 120° beamwidth and serves the three adjoining cells. We also consider both frequency reuse in every cell and the use of three channel sets. Furthermore, because in the typical system the signal strength falls with the inverse of the distance raised to between the third and fourth power, we also consider these two extremes.[6]

The performance of optimum combining and maximal ratio combining in typical mobile radio systems is shown in Table I. For each of the systems described above, Table I lists the number of antennas required to achieve a 10^{-3} BER and the average output SINR margin. We also show the margin with an additional antenna.

The results show that with three-corner base station geometry and frequency reuse in every cell, optimum combining more than halves the required number of antennas. Furthermore, the increase in margin with an additional

[6]The calculation of the power of the signals in these cellular systems will not be described here. The method is similar to that described in [10].

TABLE I
COMPARISON OF OPTIMUM AND MAXIMAL RATIO COMBINING IN TYPICAL MOBILE RADIO SYSTEMS—THE NUMBER OF ANTENNAS REQUIRED AND THE SINR MARGIN FOR A 10^{-3} BER

Case			Maximal Ratio Combining		Optimum Combining	
Base Station Geometry	Frequency Reuse	Decay Exponential	Number of Antennas	SINR Margins[a] (dB)	Number of antennas	SINR Margin[a] (dB)
3-corner	1	3	14	0.5	6	0.0
			15	1.1	7	1.7
		4	12	0.1	5	1.9
			13	0.7	6	4.5
	3	3	3	0.3	3	2.6
			4	2.9	4	5.8
		4	2	1.0	2	3.0
			3	5.5	3	8.6
Centrally Located	1	3	55	0.0	17	0.4
			69	1.0	18	0.9
		4	55	0.0	11	0.5
			69	1.0	12	1.1
	3	3	7	0.3	5	0.8
			8	1.2	6	2.4
		4	5	1.6	4	2.7
			6	2.9	5	5.3

[a]Margins are accurate to within a few tenths of a decibel and were determined from simulation results using 100 000 samples.

antenna is much greater. With the same geometry and three channel sets, even though only a few antennas are required with maximal ratio combining, optimum combining increases the margin by 2–3 dB. With centrally located base stations and frequency reuse in every cell, optimum combining substantially reduces the number of antennas. As few as 11 antennas are required with optimum combining as compared to more than 50 with maximal ratio combining. Finally, with three channel sets, optimum combining requires one less antenna and has higher margins.

Thus, the improvement with optimum combining is the largest in systems where a large number of antennas is required because of low received SINR. However, even with high SINR and few antennas, the improvement is 2 dB or more. Therefore, the results for typical cellular systems agree with those of Section III (i.e., Fig. 8).

In an actual system we would expect the optimum combining improvement to be even greater than that shown in Table I because of the following three reasons. First, all the channels in all the cells may not always be occupied. Thus, the total interference power will be less, and the power of the strongest interferers (when transmitting) relative to the power of the sum of the other interferers Γ_j will be higher. As shown in Section III, as Γ_j increases, the optimum combining improvement increases. Second, with random mobile locations rather than the worst case, the total interference power will be lower. Thus, Γ_j for the strongest interferers (those closest to the desired mobile's base station) will be higher, and therefore, so will the improvement. Third, for the results in Table I only the six strongest interferers were considered individually, and thus the results are somewhat pessimistic.

Finally, we note that in actual systems the fading can be non-Rayleigh with direct paths existing between an interfering mobile and a base station (i.e., the fading might not be independent at each antenna). Under these conditions, the performance of maximal ratio combining can be significantly degraded while optimum combining can still achieve the maximum output SINR.

V. Implementation

In this section we discuss the implementation of optimum combining in mobile radio. We consider the use of an LMS [3] adaptive array at the base station receiver and adaptive retransmission with time division for base-to-mobile transmission. For the LMS adaptive array, we discuss the dynamic range, reference signal generation, and modulation technique.

A. The LMS Adaptive Array

1) Description: Of the various adaptive array techniques [2]–[4] that can be used in mobile radio, the LMS technique appears to be the most practical one for mobile radio because it is not too complex to implement and it does not require that the desired signal phase difference between antennas be known *a priori* at the receiver.

Fig. 9 shows a block diagram of an M element LMS adaptive array. It is similar to the optimum combiner of Fig. 1 except for the addition of a reference signal $r(t)$ and an error signal $e(t)$. As shown in Fig. 9, the array output is subtracted from a reference signal (described below) $r(t)$ to form the error signal $e(t)$. The element weights are generated from the error signal and the $x_{I_i}(t)$ and $x_{Q_i}(t)$ signals by using the LMS algorithm which minimizes the power of the error signal.

The reference signal is used by the array to distinguish between the desired and interfering signals at the receiver. It must be correlated with the desired signal and uncorrelated with any interference. Under these conditions the minimization of the power of the error signal suppresses interfering signals and enhances the desired signal in the array output. Generation of the reference signal in digital mobile radio systems is described in Section V-A3).

We now consider the weight equation for the LMS adaptive array in a mobile radio system. In the typical system the bit rate is 32 kbits/s, and the carrier frequency is about 840 MHz. With the signal bandwidth 1.5 times the bit rate, the relative bandwidth of the mobile radio channel is only 0.006 percent, and we can consider the signal as narrow band. For narrow-band signals, the weight equation for the LMS array is given by [6, eq. (9)], i.e., the LMS adaptive array maximizes the output SINR. However, these are the steady state weights, and in mobile radio the signal environment is continuously changing. Therefore, we must consider the transient performance of the array. That is, because the weights are constantly changing, the performance will be degraded somewhat from that of the optimum combiner. (Analysis of the transient performance is not considered in this paper.) Also, we must consider the dynamic range of the LMS adaptive array.

2) Dynamic Range: One limitation of the LMS adaptive array technique is the dynamic range over which it can operate. In an LMS adaptive array, the speed of response to the weights is proportional to the strength of the signals at the array input. For the array to operate properly, the weights must change fast enough to track the fading of the desired and interfering signals. However, the weights must also change much more slowly than the data rate so that the data modulation is not altered. It has been shown [11] that for PSK signals the maximum rate of change in the weights without significant data distortion is about 0.2 times the data rate. For the typical mobile radio system, the maximum fading rate is about 70 Hz (for a carrier frequency of 840 MHz and a vehicle speed of 55 mi/h), and the code rate is 32 kbits/s. Thus, the permissible range in signal power at the array input is given by

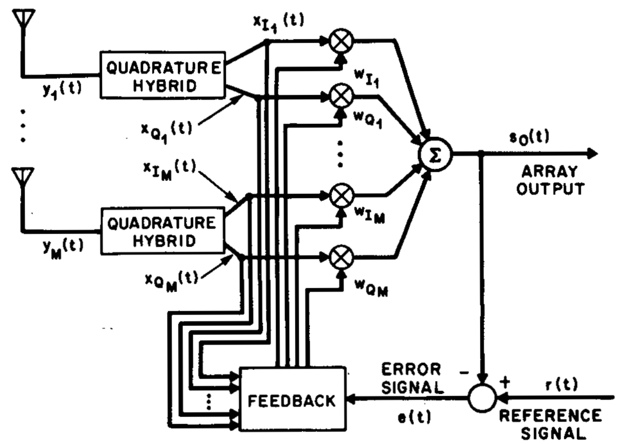

Fig. 9. Block diagram of an M element LMS adaptive array.

$$\text{Dynamic Range} = \frac{0.2 \times 32 \times 10^3}{70} \doteq 20 \text{ dB}. \tag{42}$$

The received signals in a mobile radio system vary by more than 20 dB, however, and therefore automatic transmitter power control (which could add significantly to the cost of the mobile radio) is required to control the power of the strongest signals at the receiver. With this power reasonably fixed, the dynamic range determines the power ratio of the strongest to the weakest received signal that the array can track. A 20 dB dynamic range is certainly not large, but it is more than adequate for mobile radio for the reasons described below.

In the mobile radio systems studied in this paper (see Section IV), the average received SINR at each antenna is relatively small. This is because an adaptive array is not needed when the received SINR is large. For example, for maximal ratio combining with two antennas, an average received SINR at each antenna of 11 dB [1] is required for coherent detection of PSK with a 10^{-3} BER. For optimum combining the required SINR is less with two antennas and, of course, even lower with more antennas. Thus, the received SINR is much less than 20 dB for all cases of interest. (It is typically between -5 and 5 dB.)

A small received SINR affects array operation as follows. First, if the power of an interfering signal is more than 20 dB below the desired signal's power at an antenna, the array need not track the interfering signal at that antenna because it has a negligible effect on the output SINR. Second, if the power of an interfering signal is more than 20 dB higher than the desired signal's power at an antenna, the array need not track the desired signal at that antenna because the resulting weight for the antenna will be almost zero. Thus, because the received SINR is small in the systems where the LMS adaptive array is practical, a 20 dB dynamic range is adequate. Note that if the received SINR is large (e.g., greater than 20 dB, as in a lightly loaded system), the LMS adaptive array will have the same performance as maximal ratio combining.

3) Reference Signal Generation and Modulation Technique: The LMS adaptive array must be able to distinguish between the desired signal and any interfering signals. This is accomplished through the use of a reference signal as discussed in Section V-A1). The reference signal must be correlated with the desired signal and uncorrelated with any interference.

A reference signal generation technique that allows for signal discrimination is described in [12] and involves the use of pseudonoise codes with spread-spectrum techniques. To generate the spread-spectrum signal the pseudonoise code symbols, generated from a maximal length feedback shift register, are mixed with lower speed voice (data) bits, and the resulting bits are used to generate a PSK signal. The code modulation frequency is an integer multiple of the voice bit rate, and this multiple is defined as the spreading ratio k.

The reference signal is generated from the biphase spread-spectrum signal using the loop shown in Fig. 10. The array output is first mixed with a locally generated signal modulated by the pseudonoise code. When the codes of the locally generated signal and the desired signal in the array output are synchronized, the desired signal's spectrum is collapsed to the data bandwidth. The mixer output is then passed through a filter with this bandwidth. The biphase desired signal is therefore unchanged by the filter. The filter output is then hard limited so that the reference signal will have constant amplitude. The hard-limiter output is mixed with the locally generated signal to produce a biphase reference signal. The reference signal is therefore an amplitude scaled replica of the desired signal. Any interference signal without the proper code has its waveform drastically altered by the reference loop. When the coded locally generated signal is mixed with the interference, the interference spectrum is spread by the code bandwidth. The bandpass filter further changes the interference component out of the mixer. As a result, the interference at the array output is uncorrelated with the reference signal. Thus, with spread spectrum, a reference signal is continuously generated that is correlated with the desired signal and uncorrelated with any interference. Furthermore, since pseudonoise codes are used, every mobile can be distinguished by a unique code.

Unfortunately, spread spectrum increases the biphase signal bandwidth by a factor of k and therefore increases both the total cochannel interference power and the number of interferers in cellular mobile radio. For example, with frequency reuse in every cell, the cochannel interference power and the number of interferers from surrounding cells are increased by factors of k and $2k-1$, respectively. This increase in interference power is canceled by the processing gain of spread spectrum, but the increased number of interferers degrades the performance of the LMS adaptive array. Furthermore, $2(k-1)$ cochannel interferers are now present within the desired mobile's cell. Thus, even with a small spreading ratio (e.g., 5 or less) the performance of the LMS adaptive array with the biphase spread-spectrum signal can be worse than that of maximal ratio combining, making the LMS system impractical.

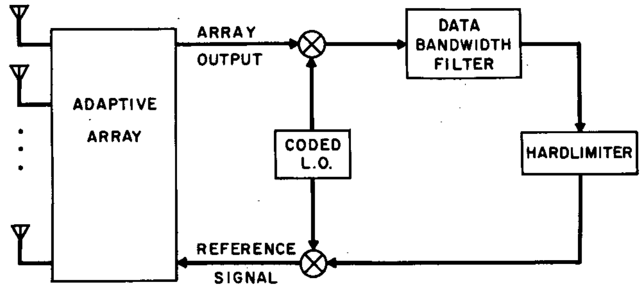

Fig. 10. Reference signal generation loop with the adaptive array. When the desired signal is a biphase spread-spectrum signal, the reference signal is correlated with it but not with any interference.

The bandwidth increase with spread spectrum and its associated problems can be overcome in the following way. The biphase spread-spectrum signal is combined with an orthogonal biphase signal modulated by the voice bits only (see [13]). The data modulation rate of the orthogonal biphase signal is the same as the code modulation rate of the biphase spread-spectrum signal. The resulting four-phase signal therefore has a bandwidth determined by the data rate only, i.e., the bandwidth is not increased by the spreading ratio. Furthermore, a reference signal for the four-phase signal can be generated from its biphase spread-spectrum signal component using the loop described earlier. As shown in [14], the performance of the LMS adaptive array with the four-phase signal is close to that with the biphase signal. Therefore, with this system, we can generate a reference signal without any increase in interference power or the number of interferers and achieve an improvement with an LMS adaptive array close to that for optimum combining which is shown in Sections III and IV.

We now describe the modulation technique in detail by describing three possible ways to modulate the four-phase signal. The simplest technique is for the voice bits to modulate only the orthogonal biphase signal. The biphase spread-spectrum signal then contains the code plus data bits for transferring information from the mobile to the base station. With this first technique, the signal bandwidth corresponds to the voice bit rate r (e.g., 32 kbits/s). However, the energy-per-bit-to-noise (interference) density ratio E_b/N_o is half that of a biphase signal. Thus, the improvement with an LMS adaptive array is 3 dB less than that shown in Sections III and IV. A data channel is also available, however, with an r/k data rate. Furthermore, since the E_b/N_o for the data bits is k times that for the voice bits (because of the spread spectrum), the BER for the data bits is very low.

If a data channel is not required, then voice bits can replace the data bits. With this second technique, the voice bits are split into two channels, one modulating the biphase spread-spectrum signal and the other modulating the orthogonal biphase signal. The bit rate for the latter channel is k times that for the biphase spread-spectrum signal. The signal bandwidth is reduced by $k/(k+1)$ as compared to the first technique. However, the E_b/N_o of the voice bits on the biphase spread-spectrum signal is k times that on the orthogonal biphase signal. Through appropriate coding techniques, this difference can be used to improve the overall BER.

We can equalize the BER for both channels by decreasing the power of the biphase spread-spectrum signal by $1/k$. With this third technique the E_b/N_o for the voice bits is just $k/(k+1)$ times that for a biphase signal. For example, with k equal to 5, the improvement with an LMS adaptive array is 0.8 dB less than that shown in Sections III and IV. Table II summarizes the above results for the three modulation techniques.

A block diagram of the four-phase signal generation circuitry for the three modulation techniques is shown in Fig. 11. The code symbols of duration Δ are mixed with either voice or data bits of duration $k\Delta$. The resulting symbols modulate a local oscillator to generate a biphase spread-spectrum signal. As shown in the lower portion of Fig. 11, voice bits, also of duration Δ, modulate the local oscillator signal shifted by 90° to generate the orthogonal biphase signal. This signal is then combined with the biphase spread-spectrum signal to obtain the four-phase signal. By adjusting the biphase spread-spectrum signal level with β and modulating this signal with either voice or data bits, we can generate any of the three four-phase signals listed in Table II.

TABLE II
FOUR-PHASE SIGNAL PARAMETERS FOR THREE MODULATION TECHNIQUES IN AN LMS ADAPTIVE ARRAY SYSTEM

Technique No.	Relative Biphase Signal Powers	Spread-Spectrum Biphase Signal: Information Bits	Spread-Spectrum Biphase Signal: Bit Rate[a]	Spread-Spectrum Biphase Signal: E_b/N_o[b]	Orthogonal Biphase Signal: Information Bits	Orthogonal Biphase Signal: Bit Rate	Orthogonal Biphase Signal: E_b/N_o[b]
1	1:1	Data	r/k	$k/2$	Voice	r	0.5
2	1:1	Voice	$r/(k+1)$	$k/2$	Voice	$(\frac{k}{k+1})r$	0.5
3	$1/k$:1	Voice	$r/(k+1)$	$k/(k+1)$	Voice	$(\frac{k}{k+1})r$	$k/(k+1)$

[a]The code modulation rate is k times the bit rate.
[b]Relative to biphase signals.

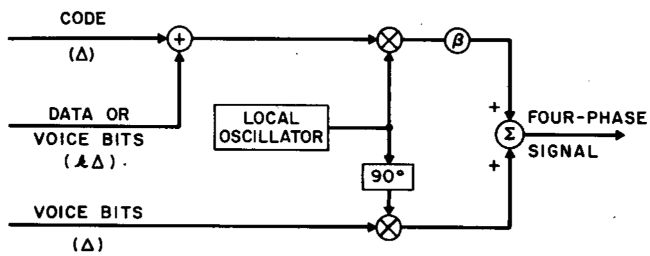

Fig. 11. Block diagram of the four-phase signal generation circuitry for the LMS adaptive array. A biphase spread-spectrum signal, modulated by code symbols plus data or voice bits, is combined with an orthogonal biphase signal, modulated by voice bits, to generate the four-phase signal.

B. Base-to-Mobile Transmission

As we have shown, the LMS technique can significantly improve signal reception at the base station. This improvement is, of course, also desired at the mobile. However, since there are many more mobiles than base stations, it is economically desirable to add the complexity of the LMS technique (particularly multiple antennas) only to the base stations.

Adaptive retransmission with time division [1], [9] can be used to improve reception at the mobile with multiple base station antennas only. With adaptive retransmission, the base station transmits at the same frequency as it receives, using the complex conjugate of the receiving weights. With time division, a single channel is time shared by both directions of transmission. Thus, with the LMS technique, during mobile-to-base transmission the antenna element weights are adjusted to maximize the signal-to-noise ratio at the receiver output. During base-to-mobile transmission, the complex conjugate of the receiving weights are used so that the signals from the base station antennas combine to enhance reception of the signal at the desired mobile and to suppress this signal at other mobiles. Therefore, by keeping the time intervals for transmitting and receiving much shorter than the fading rate (e.g., transmitting in 10 bit blocks), we can achieve the advantages of the LMS technique at both the mobile and the base station.

With adaptive retransmission using the LMS technique, each base station transmits in a way that maximizes the power of the signal received by the desired mobile relative to the total power of the signal received by all other mobiles. Thus, at the mobiles, interfering base station signals are suppressed and the improvement in the performance with the LMS technique as compared to maximal ratio combining should be similar to that at the base stations. The actual improvement for a given mobile, however, depends on the interference environment of every base station. Because of the complexity of the analysis, we will not study this improvement in detail. It should be noted, though, that for base-to-mobile transmission, spread spectrum on the signal is not required because a reference signal is not generated at the mobile. Therefore, without the degradation with the modulation scheme in the mobile-to-base transmission (see Section V-A-3), the BER at the mobile may be lower than that at the base station.

VI. SUMMARY AND CONCLUSIONS

In this paper we have studied optimum combining for digital mobile radio systems. The combining technique is optimum in that it maximizes the output SINR at the receiver even with cochannel interference. We determined the BER performance of optimum combining in a Rayleigh fading environment and compared the performance to that of maximal ratio combining. Results showed that with cochannel interference there is some improvement over maximal ratio combining with only a few receiving antennas, but there is significant improvement with several antennas. With optimum combining, the typical cellular system was seen to have greater margins and require fewer antennas than with maximal ratio combining. Finally, we described how optimum combining can be implemented in mobile radio with LMS adaptive arrays. Thus, we have shown that optimum combining is a practical means for increasing the channel capacity and performance of digital mobile radio systems.

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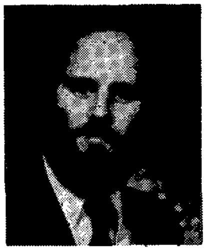

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