Effects of Received Power Imbalance on the Diversity Gain of a Digital TV MRC Array Antenna

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SUMMARY This paper presents a basic investigation of the power imbalance problem with regard to maximum ratio combining (MRC) array antennas for digital TV broadcast reception. First, the relationship between the decrease in the diversity gain and reduction in the received power was investigated using two-element and four-element dipole array antennas by means of a Monte Carlo simulation. The relationship between the decrease in the diversity gain and the number of branches imposed to reduce the received power was also investigated. Then, a simple method of predicting the reduction in the diversity gain under imbalanced power conditions is given using the simulation results. The objective is to determine a criterion associated with the gain reduction that allows us to achieve the required system performance. Finally, the proposed method is confirmed by analysis using a model representing a typical portable digital broadcasting TV set held with both hands that simulates the power imbalance condition.

key words: maximum ratio combining (MRC), diversity antenna, power imbalance, digital TV, Monte Carlo simulation

1. Introduction

Digital broadcasting terrestrial services have spread all over the world, including Europe and the Americas. In Japan, the Terrestrial Integrated Services Digital Broadcasting (ISDB-T) system has been adopted. The ISDB-T system has a unique feature in that it offers not only high definition television (HDTV) services for fixed reception, commonly referred to as “full segmentation,” but also mobile multimedia services for mobile reception, called “one segmentation.” Initially, the ISDB-T system was started to primarily provide vehicular reception services [1], [2]. However, there is currently an increasing demand for mobile reception services, even for HDTV full segmentation reception on small handheld digital broadcasting receivers, including a notebook-sized portable LCD TV set and a cellular phone-type handset receiver, all using built-in antennas [3].

In order to achieve the performance required for HDTV reception, maximum ratio combining (MRC) diversity systems are potential candidate technologies because MRC can reduce the required signal-to-noise ratio (SNR) for broadcasting signals, as compared to a single antenna receiver system [4]. However, to achieve the desired MRC performance, some technical obstacles must be overcome because some diversity branches comprising an MRC array implemented in a tightly volume-constrained device experience a significant antenna gain reduction. In addition, in the case of a handset receiver, the electromagnetic (EM) interaction between the branch antenna elements and the user, particularly from the hand and the fingers of the user, can cause further severe gain reduction [5], [6]. In this situation, the difference in the antenna gain, leading to power imbalance of the received signals, is a major cause of degradation in MRC performance because the fading correlation between branches tends to be low owing to a wide angular power spread of incident waves in azimuth [7]. Hence, the diversity gain of an MRC array antenna decreases mainly owing to the power imbalance in the received signals. However, there has been little study on the relationship between the power imbalance and the diversity gain of an MRC diversity antenna.

This paper presents a basic investigation on the power imbalance problem with regard to an MRC array antenna for digital TV broadcast reception. First, the relationship between the decrease in the diversity gain and reduction in the received power was investigated using two-element and four-element dipole array antennas by means of a Monte Carlo simulation. The relationship between the decrease in the diversity gain and the number of branches imposed to reduce the received power was also investigated. Then, a simple method of predicting the reduction in the diversity gain under imbalanced power conditions is given using the simulation results. The objective is to determine a criterion associated with the gain reduction that allows us to achieve the required system performance. Finally, the proposed method is confirmed by analysis using a model representing a typical portable digital broadcasting TV set held with both hands that simulates the imbalanced power condition.

2. Analysis of an MRC Array under Imbalanced Power Conditions

Figure 1 shows the minimum mean square error (MMSE) adaptive array antenna model used for the simulation considering power imbalance. The MRC function is achieved by the MMSE adaptive array without interference signals. As shown in this figure, attenuators are connected at the input port of each element; thus, the received signals can be intentionally reduced by ΔG. With this configuration, we can simulate the received imbalanced power conditions for MRC antennas.

In practice, the human body and electrical devices in the immediate vicinity of an antenna cause the gain of the...
The half-wavelength dipole array antenna has a small correlation of 0.01 (see Sect. 3), and thus changes in the mutual coupling due to the reduction in the radiation efficiency result in small changes in the correlation.

Figure 2 shows the channel model for an MRC array antenna under a Rayleigh fading environment. This model was originally developed for simulating handset adaptive arrays with multiple interferences [8]. Although the author assumed a radio wave environment with the absence of interference signals for achieving the MRC function using the MMSE scheme, the simulation procedure in an interference environment is briefly described. This will be helpful for considering future radio wave propagation environments of digital TV systems, because appreciable interference may appear from dense neighboring broadcasting stations.

Figure 2 shows a radio propagation model comprising one desired and two interference signals. A handset is assumed to be surrounded by \( N \) scatterers arranged uniformly in azimuth. Using this model, a Monte Carlo simulation has been carried out. Assuming that the \( n \)th path has a transfer function with vertical and horizontal components, the transfer function of the \( n \)th path at the \( k \)th antenna is given by

\[
\begin{align*}
h_{Vkn} &= \sqrt{\frac{\text{XPR}}{1 + \text{XPR}}} h_{Vn} \Delta G_k E_{Vk}(\theta_n, \phi_n) \exp(j \varphi_{Vn}) \\
h_{Hkn} &= \sqrt{\frac{1}{1 + \text{XPR}}} h_{Vn} \Delta G_k E_{Hk}(\theta_n, \phi_n) \exp(j \varphi_{Hn})
\end{align*}
\]  

where \( E_{Vk}(\theta_n, \phi_n) \) and \( E_{Hk}(\theta_n, \phi_n) \) are the complex electric field directivity of the \( k \)th antenna element for the \( \theta \) and \( \phi \) components, which are calculated by the method of moments, respectively. \( \Delta G_k \) is the reduction in the antenna gain, as shown in Fig. 1. \( h_{Vn} \) represents the equivalent amplitude of the incident waves and can be set to an arbitrary value; thus, it is assumed to have unity amplitude. XPR is the cross-polarization power ratio. For the vertical and horizontal polarization components, their phases, \( \varphi_{Vn} \) and \( \varphi_{Hn} \), are independent of each other and uniformly distributed from 0 to \( 2\pi \).

For each path, the two polarization components are combined as a complex sum of vertical and horizontal components, as shown in Fig. 3:

\[ h_{kn} = h_{Vkn} + h_{Hkn} \]  

The resultant channel response at the \( k \)th antenna is expressed as the sum of \( N \) paths in Fig. 2 using the following equation:

\[ h_k = \sum_{n=1}^{N} h_{kn} \exp\left\{ j \frac{2 \pi d}{\lambda} \cos(\phi_n - \phi_m) \right\} \]  

where \( \lambda \) is the wavelength in free space, and \( d \) is the distance traveled by a handset that moves toward the azimuth direction of \( \phi_m \). The scheme represented by Eqs. (1) and (4) is applied repeatedly to generate both the desired and the interference channel responses. The \( l \)th sample of the received signal at the \( k \)th antenna \( u_{kl} \) is composed of the desired signal \( s_{dl} \), the interference signal \( s_{ipl} \), and noise \( n_{kl} \):

\[ u_{kl} = h_{dk} s_{dl} + \sum_{p} h_{ipk} s_{ipl} + n_{kl} \]  

where \( h_{dk} \) and \( h_{ipk} \) are the channel responses of the desired and the \( p \)th interference signals for the \( k \)th antenna, each of which is calculated from Eq. (4). Next, \( u_{kl} \) is multiplied by weight \( w_{kl}^* \), as shown in Fig. 1, and summed in order to obtain the output \( y_{l} \) as \( y_{l} = w_{l}^* u \), where \( ^* \) denotes the Hermitian conjugate operator. By applying the MMSE condition to output \( y_{l} \), the optimum weight vector \( w_{o} \) is expressed
as \( \mathbf{w}_o = R^{-1}\mathbf{r}_d \), where \( R = E[\mathbf{u}_d\mathbf{u}_d^H] \) is the autocorrelation matrix of the received signal, and \( \mathbf{r}_d = E[\mathbf{u}_d\mathbf{d}_r^H] \) is the correlation vector between the received and the reference signals.

In a practical system, the reference signals comprising predetermined or known symbol trains, commonly referred to as a training sequence, are transmitted prior to the transmission of the actual data sequence. In the simulation conducted here, the reference signal is assumed to be equal to the desired signal.

The resultant SINR at the array output is given by

\[
SINR_o = \frac{\mathbf{w}_o^H R_d \mathbf{w}_o}{\mathbf{w}_o^H (R_s + \sigma^2 I) \mathbf{w}_o}
\]

where \( \sigma^2 \) is the noise variance at each of the receivers. \( R_d = E[\mathbf{u}_d\mathbf{u}_d^H] \) and \( R_s = E[\mathbf{u}_d\mathbf{u}_d^H] \) are the autocorrelation matrices of the desired and interference signals, respectively. The BER for the coherent detection of phase shift keying signals can be calculated as

\[
P_e = \alpha \text{erfc} \sqrt{\beta \text{SINR}_o}
\]

assuming that the interference signals are considered to be equivalent to Gaussian noise. In Eq. (7), \( \text{erfc} \) denotes the complementary error function. Here, we use \( \alpha = 1/2 \) and \( \beta = 1/2 \) for QPSK signals, which are used in the one-segmentation ISDB-T system, and \( \alpha = 7/24 \) and \( \beta = 1/42 \) for 64QAM signals, which are adopted in the full-segmentation ISDB-T system. The average BER, \( P_{ave} \), over the entire traveling distance is calculated by summing every instantaneous BER, \( P_{ei} \), with the following equation:

\[
P_{ave} = \frac{1}{S} \sum_{i=1}^{S} P_{ei},
\]

where \( S \) signifies the number of samples.

### 3. Basic Study Using Dipole Arrays

We considered array antennas comprising two and four vertically oriented half-wavelength dipole antennas with half-wavelength spacing in free space. Figure 4 shows the configuration of the four-element dipole array antenna. The origin is placed at element #1, and each element is aligned to the \( z \)-axis in an arrangement toward the \( y \)-axis. Because the analysis was conducted at a frequency of 600 MHz, the array spacing was set to 250 mm.

Table 1 lists the analysis conditions. The SIR was set to infinity, meaning that there are no interference signals for achieving the MRC function using the MMSE scheme. XPR was set to infinity, indicating that only vertically polarized waves exist. The number of scatterers was assumed to be 15.

In the Monte Carlo simulation, the dipole array is moved by a traveling distance of 50\( \lambda \), with the distance between two successive points in the simulation being \( \Delta d = 0.01\lambda \). The average BER over the entire traveling distance is calculated using Eq. (8).

Figure 5 shows the amplitude and phase radiation patterns of the four-element dipole array antenna in the \( x-y \) plane, equivalent to the \( \theta \)-polarized component, \( E_{Vx} \), in

![Figure 4 Configuration of the four-element dipole array antenna.](image)

**Table 1 Analytical conditions.**

<table>
<thead>
<tr>
<th>Frequency</th>
<th>600 MHz</th>
</tr>
</thead>
<tbody>
<tr>
<td>SIR</td>
<td>Infinity(No Interference)</td>
</tr>
<tr>
<td>XPR</td>
<td>Infinity(VERTICAL POL.)</td>
</tr>
<tr>
<td>Number of scatterers(K_s)</td>
<td>15</td>
</tr>
<tr>
<td>Initial phase for scatterers</td>
<td>Random</td>
</tr>
<tr>
<td>Traveling distance</td>
<td>50 wavelengths</td>
</tr>
<tr>
<td>Number of sample (S)</td>
<td>5000</td>
</tr>
<tr>
<td>Sampling interval (( \Delta d ))</td>
<td>0.01 wavelengths</td>
</tr>
<tr>
<td>Moving direction (( \phi_v ))</td>
<td>5°</td>
</tr>
<tr>
<td>Modulation</td>
<td>QPSK, 64QAM</td>
</tr>
<tr>
<td>Antenna element</td>
<td>Half-wavelength dipole</td>
</tr>
<tr>
<td>Method of EM analysis</td>
<td>Method of moments</td>
</tr>
</tbody>
</table>

Eq. (1). The lateral axis in the graph of phase characteristics is the azimuth angle \( \phi \) with respect to the \( x \)-axis in Fig. 2(b).

The analysis was conducted using the method of moments.

Some appreciable distortion in the pattern can be seen in Fig. 5 due to EM couplings between the elements for the four-element dipole array antenna, even though each individual dipole antenna has an omnidirectional radiation pattern in the horizontal plane. Figure 5 also shows that the combination of elements #1 and #4 and elements #2 and #3 have symmetrical radiation patterns with respect to the \( x \)-axis.

As seen in Fig. 5, the phase characteristics of all the elements show sinusoidal variations. It should be noted that the maximum phase variation for element #4 is found to be 525.4° and -555.2° (1080.6° peak-to-peak value), which agrees with the phase variation corresponding to an array spacing of 1.5 wavelengths between elements #1 and #4.

Table 2 shows the radiation efficiencies and mean effective gains (MEGs) [9] of the four-element dipole array antenna in free space at 600 MHz. Although the amplitude radiation patterns exhibit some distortion as described in Fig. 5, Table 2 shows that the decrease in the radiation efficiency is very small. Table 2 also shows that the MEG of each element is approximately 0.5 dB smaller than that of a single dipole antenna in the radio wave propagation environment under the conditions listed in Table 1. The fading correlation, defined as the square of the absolute value of complex correlation coefficient, between the elements is calculated to be smaller than 0.01. The fading correlation used here is equivalent to the envelope correlation described in [10].
Amplitude and phase radiation patterns of the four-element dipole array antenna.

Table 2  Radiation efficiencies and mean effective gains in free space of the four-element dipole array antenna at 600 MHz.

<table>
<thead>
<tr>
<th>Element#1</th>
<th>Element#2</th>
<th>Element#3</th>
<th>Element#4</th>
</tr>
</thead>
<tbody>
<tr>
<td>Radiation efficiency (dB)</td>
<td>-0.18</td>
<td>-0.29</td>
<td>-0.29</td>
</tr>
<tr>
<td>MEG (dBi)</td>
<td>1.71</td>
<td>1.52</td>
<td>1.52</td>
</tr>
</tbody>
</table>

Figure 6 shows the average BER characteristics as a function of the average input SNR with the reduction in antenna gain $\Delta G = \Delta G_k$ as parameters. For a two-branch MRC array antenna, the antenna gain of branch #1 is reduced, and in the case of a four-branch MRC array antenna, the gains of branches #1 and #2 are reduced with the same values. As a reference case, the BER for a single dipole is also shown in this figure. In the two-branch MRC array antenna, only two elements (#1 and #2) shown in Fig. 4 exist.

Now, the diversity gain is defined as the ratio of the average SNR of a single dipole antenna to that of the antenna under consideration when BER is achieved at a prescribed value, $P_{ave} = P_o$, by the following equation:

$$G_{div} = \frac{\text{SNR}_{Single\ Dipole}}{\text{SNR}_{P_{ave}=P_o}}.$$  (9)

For the QPSK modulation shown in Fig. 6(a), when $P_o = 10^{-3}$ and $\Delta G = 0$ dB, the diversity gains for the two-branch and four-branch MRC are found to be $G_{div} = 13.2$ dB and $20.5$ dB, respectively. In contrast, in the 64QAM shown in Fig. 6(b), the two corresponding values are $G_{div} = 10.6$ dB and $17.3$ dB. Furthermore, as can be seen from Fig. 6, the BER characteristics are degraded with increasing $\Delta G$.

Consequently, as a limiting case of sufficiently large $\Delta G$, the BER of the two-branch MRC array antenna converges to the BER of a single dipole, whereas the BER of the four-branch MRC array antenna converges to the BER of the two-branch MRC array antenna with $\Delta G = 0$ dB.

To understand the reduction in the diversity gain due to the received power imbalance shown in Fig. 6 in more detail, the diversity gain was calculated as a function of the reduction in the received power when different numbers of elements were imposed on the reduction in the received power.

Figure 7 shows the relationship between the diversity gain $G_{div}$ of the four-dipole MRC array antenna and the antenna gain reduction $\Delta G$ for QPSK (Figs. 7(a), (b), (c)) and 64QAM (Figs. 7(d), (e), (f)) modulation, in which (a, d), (b, e), and (c, f) show the diversity gain when $\Delta G$ is reduced by 0–30 dB for branches (#1), (#1 and #2), and (#1, #2, and #3), respectively. It was confirmed by other simulations that other combinations, for example (#1 and #4), give similar results.

In the figures, the three lines indicate the cases when
$P_o$ equals $10^{-2}$, $10^{-3}$, and $10^{-4}$. It is shown in [6] that, when using a real human, the gain reduction in a helical antenna mounted on a small metal housing by the fingers of the user was found to be around 8 to 9 dB. It can be understood from this fact that the most probable power difference is less than 10 dB for practical applications. Figure 7 shows that the diversity gain decreases in proportion to $\Delta G$ in the region where $\Delta G$ is less than 10 dB.

Figure 7 also shows that in each case, the diversity gain converges to a certain constant value for $\Delta G$ greater than 20 dB. The three symbols plotted at $\Delta G = 30$ dB in Figs. 7(a)–(f) show the results of calculations when a 3, 2, or 1-branch antenna is used. In each figure, it can be seen that the three curves converge to the corresponding symbols.

Table 3 shows the average gradient of the plots in Fig. 7 for QPSK and 64QAM when $\Delta G < 10$ dB. We see in Table 3(b) for 64QAM with three-element gain reduction (#1, #2, #3) that the gradient is equal to 6.45, 7.08, and 7.83 dB per 10 dB reduction in $\Delta G$ as $P_o$ is reduced, indicating that the received power imbalance has a greater influence on the

![Fig. 7](image.png)

**Fig. 7** Relationship between the diversity gain $G_{div}$ and the antenna gain reduction $\Delta G$ for the four-dipole MRC array antenna.

<table>
<thead>
<tr>
<th>BER($P_o$)</th>
<th>#1</th>
<th>#1, #2</th>
<th>#1, #2, #3</th>
</tr>
</thead>
<tbody>
<tr>
<td>$10^{-2}$</td>
<td>2.28 dB</td>
<td>4.11 dB</td>
<td>6.20 dB</td>
</tr>
<tr>
<td>$10^{-3}$</td>
<td>2.32 dB</td>
<td>4.53 dB</td>
<td>6.42 dB</td>
</tr>
<tr>
<td>$10^{-4}$</td>
<td>2.36 dB</td>
<td>4.62 dB</td>
<td>6.60 dB</td>
</tr>
</tbody>
</table>

**Table 3** Gradient of the average diversity gain per 10 dB reduction in $\Delta G$ shown in Fig. 7 in the region $\Delta G < 10$ dB.

(a) QPSK

<table>
<thead>
<tr>
<th>BER($P_o$)</th>
<th>#1</th>
<th>#1, #2</th>
<th>#1, #2, #3</th>
</tr>
</thead>
<tbody>
<tr>
<td>$10^{-2}$</td>
<td>2.36 dB</td>
<td>4.43 dB</td>
<td>6.45 dB</td>
</tr>
<tr>
<td>$10^{-3}$</td>
<td>2.40 dB</td>
<td>5.19 dB</td>
<td>7.08 dB</td>
</tr>
<tr>
<td>$10^{-4}$</td>
<td>2.45 dB</td>
<td>5.66 dB</td>
<td>7.83 dB</td>
</tr>
</tbody>
</table>

(b) 64QAM
diversity gain for smaller $P_o$. Table 3 also shows that in the region where $P_o$ is small (i.e., $P_o = 10^{-3}$), the average gradient increases with the number of elements with reduced received power. Furthermore, 64QAM has a larger gradient than QPSK when comparing the two modulation schemes, indicating that the full-segmentation ISDB-T system has a greater influence of power imbalance on the diversity gain than the one-segmentation ISDB-T system.

Using these results, the reduction in the diversity gain can be estimated in a simple manner by knowing the number of elements with reduced received power and the value of gain reduction when $\Delta G < 10$ dB. For a 64QAM-MRC array, when the gain of two elements is reduced by $\Delta G = 6$ dB, the reduction in the diversity gain is estimated to be $\Delta G_{div} = 5.19 \times 6/10 = 3.11$ dB at $P_o = 10^{-3}$.

As seen in Fig. 6, the average BER is a function of the average SNR. This means that when we consider an MRC array antenna mounted on a portable TV set in which the array configuration, radiation patterns, and MEG are different from those of the dipole array, as will be mentioned in Sect. 4, the BER characteristics will be shifted from the curves of the dipole MRC array plotted in Fig. 6, if the correlations are sufficiently small. Because the diversity gain can be obtained from the difference between the two curves of the average SNR of a reference antenna and that of the antenna under consideration, as in Eq. (9), the diversity characteristics calculated from the dipole array illustrated in Fig. 7 can be applied to a handheld MRC antenna. These considerations mean that a method of predicting the reduction in the diversity gain can also be used for an MRC array for a portable TV set. This will be further discussed in Sect. 4.

### 4. Consideration of Imbalanced Power Conditions Using a Portable TV Set

In this section, a simple method for predicting the reduction in the diversity gain under imbalanced power conditions is confirmed by analysis using a model representing a typical portable digital broadcasting TV set held with both hands that simulates the imbalanced power condition.

Figure 8 shows the analytical model for a portable TV set held with both hands, comprising two monopole antennas and two inverted L antennas (ILA). The dimensions of the monopoles and ILAs are determined so that a return loss of more than 10 dB can be achieved with the absence of both hands. The four antennas are mounted on a metal ground plane with the dimensions of 15 cm $\times$ 8 cm, comprising a four-branch MRC diversity antenna.

The analysis is conducted with the model being placed in the horizontal configuration, representing a practical use condition when a user watches TV. The hand is modeled by a simple parallelepiped holding a portable TV set with a palm thickness of 15 mm. The electrical properties of the model are chosen such that the relative permittivity is 57 and conductivity is 0.87 S/m, which are the average values of the human muscle at 600 MHz [11]. The location of the hands can be varied with respect to the TV set in order to obtain knowledge about how the proximity of the hands to the ILAs creates the imbalanced power condition and consequently the reduction in the diversity gain.

As for the MRC array model for a portable TV set, the attenuators shown in Fig. 1 are not connected at the input port of each element, i.e., $\Delta G = 0$ dB. Hence, the effects of EM coupling are fully involved in the analysis conducted by the method of moments described in this section.

Table 4 lists the fundamental characteristics of the portable TV set without hands. A return loss of more than 10 dB is obtained for each element without a matching circuit. The radiation efficiency is found to be approximately $-4$ dB owing to poor isolation characteristics between the monopole and the ILAs, nearly 3 dB for the worst case shown in Table 4.

Figure 9 shows the VSWR characteristics as a function of the hand location at 600 MHz. The measured results us-
ing a phantom hand are also plotted for comparison [12].

There are some discrepancies between the calculated and measured results. A possible cause of these discrepancies is the effects of leakage current flowing on the outer surface of a coaxial cable used in the experiment. However, it can be seen from Fig. 9 that the fundamental characteristics describing the effect of the hand agree with each other. For instance, the VSWR of the ILAs attains maximum values at \( h_1 = h_2 = 5 \text{ cm} \), and the monopoles are insensitive to changes in the hand location for both cases. This confirms the validity of the analysis conducted in this paper.

As can be seen from the figure, the VSWR of the monopoles changes moderately, whereas the VSWR of the ILAs increases significantly when the hand location, \( h_1 = h_2 \), is \( 5 \text{ cm} \). This is attributed to the fact that a location of \( h_1 = h_2 = 5 \text{ cm} \) means that the two ILAs are almost covered with both hands, as seen in Fig. 8. This results in a strong EM coupling between the hand and the ILA. An abrupt change in the VSWR for the ILA can also be understood from the fact that the ILA has a smaller frequency bandwidth than the monopoles.

Figure 10 shows the MEG and fading correlation as a function of the hand location at 600 MHz. The analysis was conducted with the analytical model shown in Fig. 8 in the horizontal configuration. XPR was assumed to be \( -6 \text{ dB} \).

It can be seen from Fig. 10(a) that the MEG of the ILA decreases significantly when the hand location, \( h_1 = h_2 \), is between 0 and 5 cm. This is close to the location where the VSWR is degraded, as described in Fig. 9. The difference between the MEG of the ILA and that of the monopole indicates a power imbalance between the two antennas. Hence, we see in Fig. 10(a) that a power imbalance of 7.4 dB is produced when the hand location, \( h_1 = h_2 \), is 2.5 cm.

Figure 10(b) shows that the fading correlation increases and reaches approximately 0.5 when the hand location is between 0 and 5 cm. A possible cause of this phenomenon is EM coupling due to the hands enhancing the current induced on the ground plane, resulting in similar radiation patterns between the monopoles and the ILAs when the hands are present, compared to the case when the hands are absent.

Figure 11 shows the required average SNR as a function of the hand location for the 64QAM when BER equals \( 10^{-2} \), \( 10^{-3} \), and \( 10^{-4} \). The three symbols, plotted at \( h_1 = h_2 = 15 \text{ cm} \) in Fig. 11, show the results of the calculations with the absence of both hands. We find that the required average SNR increases at around \( h_1 = h_2 = 2.5 \text{ cm} \) in Fig. 11 owing to the power imbalance condition shown in Fig. 10(a).
It is concluded from these considerations that although the difference in the antenna gain attributed to the effects of the hands is a major cause of the degradation in the MRC performance, the fading correlation is also responsible for the determination of the MRC performance. Hence, to improve the estimation accuracy for the reduction in the diversity gain, the inclusion of the fading correlation in the proposed method is essential. This will be further explored in our future work.

5. Conclusion

This paper presented a basic investigation on the power imbalance problem with regard to an MRC antenna for digital TV broadcast reception. The quantitative relationship between the decrease in the diversity gain and reduction in the received power was given. Using the results obtained from the analyses, a simple method of estimating the reduction in the diversity gain has been proposed. Furthermore, the proposed method is confirmed by analysis using a model of a typical portable digital broadcasting TV set held with both hands that simulates an imbalanced power condition.

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References


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