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Calculation model for SFN reception and reference receiver characteristics of ISDB-T system

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REPORT ITU-R BT.2209

Calculation model for SFN reception and reference receiver characteristics of ISDB-T system

(2010)

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Summary

This Report provides detailed considerations on receiver characteristics under single frequency network (SFN) conditions for ISDB-T system. It introduces new technical parameters that dominate receiver performances, in addition to the conventional planning parameters. The new parameters are amplitude proportional noise (APN), FFT window setting margin and interpolation filter characteristics used for reference carrier recovery. Using these parameters, the overall receiver characteristics can be expressed by a single parameter called guard interval mask characteristics, which is useful in estimating whether or not the signal is received correctly.

Furthermore, the Report gives a reference receiver characteristic that would be applied in frequency planning and/or network design of ISDB-T based broadcasting systems. The calculation method for SFN reception and the reference receiver characteristics have been established in ARIB TR-B14, which is successfully applied in planning and designing of broadcasting networks in Japan.

It is shown that SFN does not work unconditionally, but works well only when the reception signals are kept under certain conditions in terms of reception voltages, DURs, delays between main and SFN signals and so on.

Chapter I – SFN with delays less than guard interval duration

First we will derive the condition that gives single frequency network (SFN) failure when a single SFN wave exists. Then, we will discuss about the conditions when multiple SFN waves exist. Also we will give some considerations on the receiver characteristics that are necessary to estimate the area of failure. We call SFN with delays less than guard interval duration as “inner-GI SFN” in short.

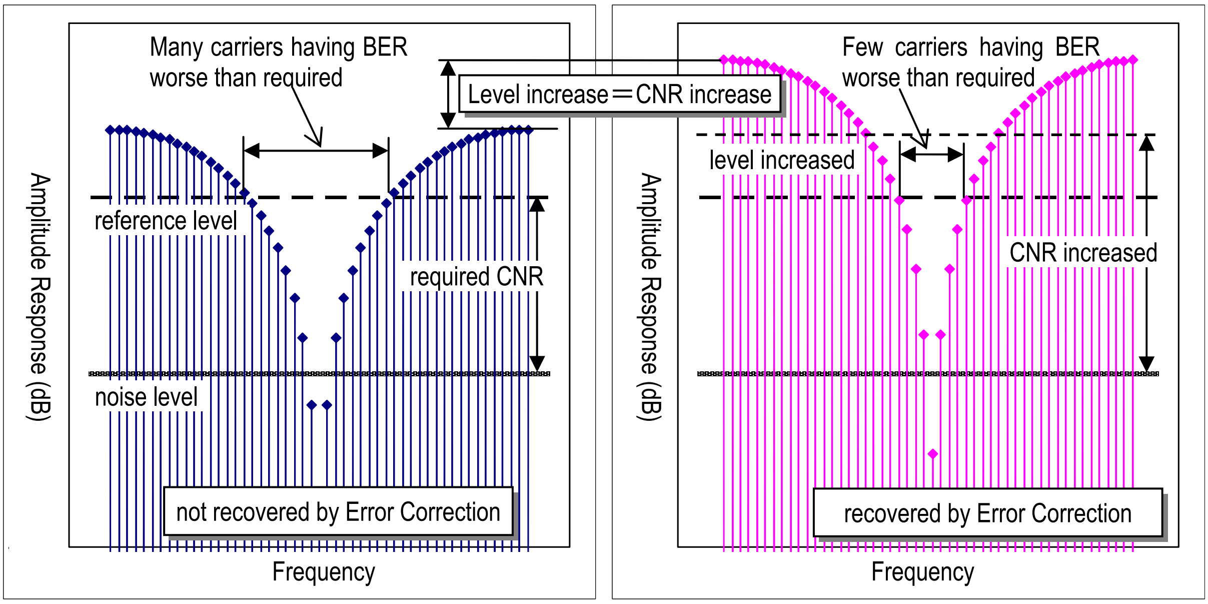
1 In the case of a single SFN wave

The received signal exhibits ripples in frequency response when the desired signal is received with SFN waves. In this case, the bit error rates (BER) of the OFDM carriers positioned at peaks in frequency response becomes better, because the input signal levels are high for those carriers. On the other hand, the BER of the carriers positioned at dips in frequency response become worse because the input levels are low. We can estimate the occurrence of SFN failure by calculating the BERs for every carrier and summing up them to check whether or not the total BER is worse than the required value.

Figure 1 shows an example of frequency response of received signal. Figure 1a) is the case where the desired signal is just at the level that gives the required carrier-to-noise ratios (CNR). In this example, we assume the total BER being worse than the required value, as there are many carriers of which BER is worse than the reference value. If we increase the levels of both desired and SFN signals, the number of carriers having worse BER is decreased, and then the signal is correctly received, as shown in Fig. 1b). The SFN failure takes place depending not only on the desired to undesired ratios (DUR) but also on the levels of the received signal itself.

FIGURE 1

Example of received signal



(a) in the case of reference signal level

(b) in the case of signal level increased

1.1 Mathematical equations used in this text

The relationship between BER and CNR are to be following:

for QPSK

$$BER = \frac{1}{2} Erfc\left(\sqrt{\frac{1}{2}\frac{C_p}{N_p}}\right) \quad \text{and} \quad \frac{C_a}{N_a} = \sqrt{\frac{C_p}{N_p}} = \sqrt{2} \cdot Erfc^{-1}(2 \cdot BER) \tag{1}$$

for 16-QAM

$$BER = \frac{3}{8} Erfc\left(\sqrt{\frac{1}{10}\frac{C_p}{N_p}}\right) \quad \text{and} \quad \frac{C_a}{N_a} = \sqrt{\frac{C_p}{N_p}} = \sqrt{10} \cdot Erfc^{-1}\left(\frac{8}{3} \cdot BER\right) \tag{2}$$

for 64-QAM

$$BER = \frac{7}{24} Erfc\left(\sqrt{\frac{1}{42}\frac{C_p}{N_p}}\right) \quad \text{and} \quad \frac{C_a}{N_a} = \sqrt{\frac{C_p}{N_p}} = \sqrt{42} \cdot Erfc^{-1}\left(\frac{24}{7} \cdot BER\right) \tag{3}$$

where:

C_p: average power of signal

C_a: r.m.s. amplitude of signal

N_p: noise power

N_a: r.m.s. amplitude of noise

and $Erfc(x) = \frac{2}{\sqrt{\pi}} \int_x^{\infty} \exp(-t^2)\, dt$

The relationship between the *Erfc* above and *Ndist* (normal distribution function) generally used in mathematics are as follows:

$$Ndist(x) = \frac{1}{\sqrt{2\pi}} \int_{-\infty}^{x} \exp\left(-t^2/2\right) \mathrm{d}t$$

$$\begin{aligned} Erfc(x) &= \frac{2}{\sqrt{\pi}} \int_{x}^{\infty} \exp\left(-t^2\right) \mathrm{d}t \\ &= \frac{2}{\sqrt{\pi}} \int_{\sqrt{2}\,x}^{\infty} \exp\left(-u^2/2\right) \frac{\mathrm{du}}{\sqrt{2}} = 2 \frac{1}{\sqrt{2\pi}} \int_{\sqrt{2}\,x}^{\infty} \exp\left(-u^2/2\right) \mathrm{du} \\ &= 2 \cdot Ndist\left(-\sqrt{2}\,x\right) \end{aligned} \quad (4)$$

where:

u: $\sqrt{2}\,t$

The relationship between the inverse functions of the above are:

by using:

$$\begin{aligned} P &= Erfc(x) = 2 \cdot Ndist\left(-\sqrt{2}x\right) \\ \frac{P}{2} &= Ndist\left(-\sqrt{2}\,x\right) \;\Rightarrow\; x = \frac{-1}{\sqrt{2}} Ndist^{-1}\left(\frac{P}{2}\right) \\ \therefore\; Erfc^{-1}(P) &= \frac{-1}{\sqrt{2}} Ndist^{-1}\left(\frac{P}{2}\right) \end{aligned} \quad (5)$$

1.2 Increase in the required CNR

In the case of a single SFN wave, the frequency response of the received signal is given by equation (6).

$$F_a(\omega) = \sqrt{1 + U_a^2 + 2U_a \cdot \cos(\omega \cdot \tau)} \quad (6)$$

where:

U_a: amplitude of SFN wave relative to desired wave

τ: delay of SFN signal.

Figure 2 shows examples of increase in the required CNR when an SFN wave exists. The horizontal axis of the graphs denotes the input power of the SFN wave relative to the desired signal, that is, the inverse figures of DUR. The vertical axis denotes the increase in the required CNR. The actual values of CNR necessary for correct reception are obtained by adding these increases to the reference CNR such as 20 dB for 64-QAM-FEC3/4, 22.5 dB for 64-QAM-FEC7/8, and so on. The curves in Fig. 2 are obtained by calculating the CNR that gives BER of 8.5×10^{-3} for FEC3/4 or 1.1×10^{-3} for FEC7/8 against a number of SFN waves of which amplitude and delay are given randomly. The relationships given by these curves are called "CNR Increase Functions" in this document.

Since CNR Increase Function cannot be expressed by simple mathematical formula, we introduce approximated function for it, as below:

$$\begin{aligned} CN_{up}(U\mathrm{dB}) &= \alpha \cdot \exp\left[-\left|\beta \cdot U\mathrm{dB}\right|^{\gamma}\right] && (U\mathrm{dB} \le 0) \\ &= \alpha \cdot \exp\left[-\left|\beta \cdot U\mathrm{dB}\right|^{\gamma}\right] - U\mathrm{dB} + A \cdot U\mathrm{dB} && (U\mathrm{dB} > 0) \end{aligned} \quad (7)$$

where:

*U*dB: denotes the power of SFN wave expressed (dB)

$CN_{up}(U\text{dB})$: denotes the increase in the required CNR expressed (dB)

The values of the coefficients, α, β and γ are given in Table 1. The value for *A* in equation (7) should be zero in usual cases.

FIGURE 2

Example of increase in the required CNR

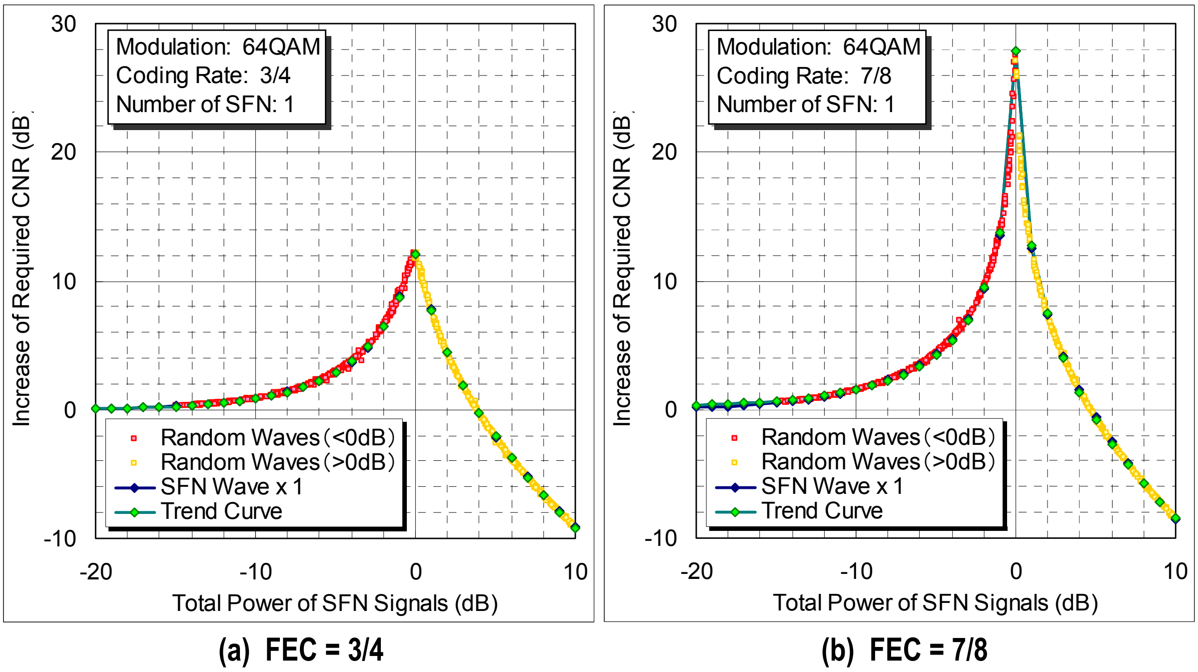


TABLE 1

Coefficient for CNR increase function

Coefficient	Modulation	FEC:7/8	FEC:5/6	FEC:3/4	FEC:2/3	FEC:1/2
α	64-QAM	27.749	20.257	12.090	8.1386	3.8797
	16-QAM	29.800	22.163	13.874	9.8342	5.4002
	QPSK	32.255	24.378	15.953	11.827	7.2391
β	64-QAM	0.5592	0.4117	0.2953	0.2527	0.2074
	16-QAM	0.6074	0.4453	0.3171	0.2702	0.2251
	QPSK	0.6702	0.4876	0.3450	0.2922	0.2437
γ	64-QAM	1.0662	0.7253	0.9096	1.0341	1.2100
	16-QAM	0.5954	0.6936	0.8616	0.9776	1.1378
	QPSK	0.5710	0.6608	0.8115	0.9172	1.0662

The value of α corresponds to the maximum increase in the required CNR, which takes place at DUR = 0 dB. A large difference is found in the maximum CNR increase between FEC 3/4 7/8, comparing to 2-3 dB in the case without SFN waves. This difference must be noted in the design of broadcasting networks, in other words, FEC of 7/8 could not be applied in the actual world where SFN is more or less used.

1.3 Noise characteristics in receivers

The noise that affects the reception performance can be summarized in two categories; one is the noise of which amplitude is independent of the input signal, such as thermal noise, and the other is the noise of which amplitude increases/decreases according to the input signal level. The former is called "fixed noise" and the latter "amplitude proportional noise" (APN) in this text.

Fixed noise consists of thermal noise, man-made noise, noise figure of receivers, and so on. An example of link budget reported by the National Council on Information and Telecommunication in Japan applies thermal noise of 300 K, man-made noise of 700 K, and noise figure of 3 dB. The value of fixed noise is estimated to be 8.5 dB(μV) (@75Ω) in this case.

Amplitude proportional noise is mainly determined by the receiver characteristics, such as the quantization noise of A/D conversion, clipping noise, phase jitter of local oscillator.

The words "fixed" and "amplitude proportion" come from the features of noise when expressed equivalently at the receiver input terminal.

1.3.1 Clipping noise

Since the amplitude distribution of OFDM signal exhibits a normal distribution, an enormous dynamic range is required for the receiver signal processing to treat without distortions. Normal receivers clip the signals exceeding a certain level, and generate clipping noise in some extent.

Figure 3 explains the affect of clipping. The clipped waveform is equivalent to the input waveform with adding the signal components that exceed the clipping level in the opposite polarity. The component that exceeds the clipping level has a pulse waveform, as shown in the Figure.

The probability that the OFDM signal takes a level of x is written by $Gauss(x/S_{rms})$, where S_{rms} denotes the *r.m.s.* amplitude of the signal, and $Gauss(*)$ the probability density function of a normal distribution. The amplitude of the pulse component that exceeds the clipping level is written by $(x - CL)$, where CL denotes the clipping level.

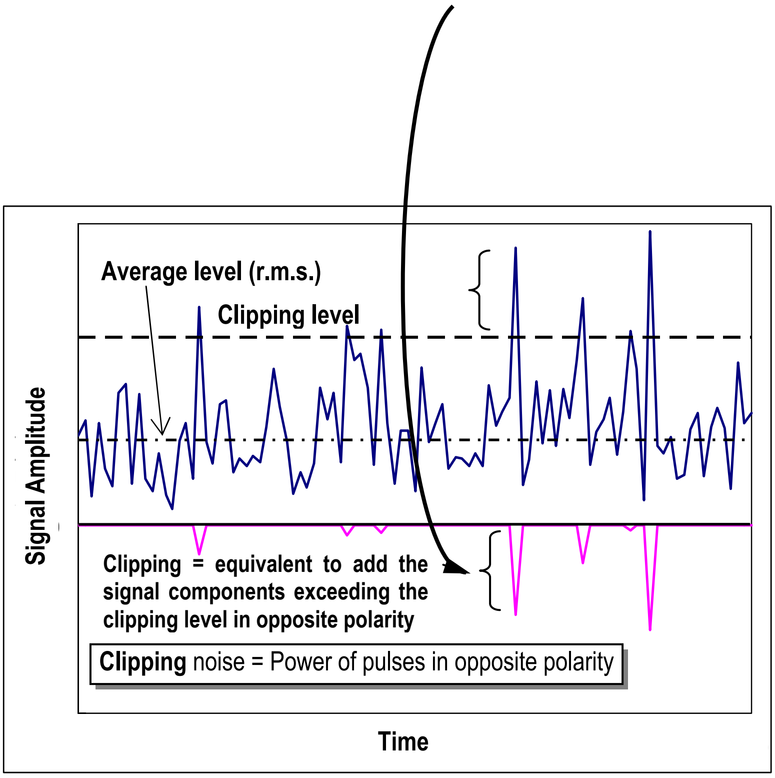
Using the above, we can estimate the power of the pulses or clipping noise as below:

$$NP_{clip} = \int_{CL/S_{rms}}^{\infty} (x - CL)^2 \cdot Gauss\ (x/S_{rms})\, \mathrm{d}x \tag{8}$$

The spectrum of clipping noise can be regarded as flat noise, because the intervals of these pulses are considerably large and isolated pulse has flat spectrum.

FIGURE 3

Clipping noise



1.3.2 Quantization noise of A/D conversion

It is assumed for normal receivers that the clipping level is set equal to the full scale of the A/D converter. Then, the quantization noise is given by the well-known equation (9).

$$\begin{aligned} NP_{adc} &= LSB^2/12 \\ &= \left(CL\cdot 2^{-N}\right)^2/12 = CL^2\cdot 2^{-2N}/12 \end{aligned} \tag{9}$$

where:

LSB: denotes the amplitude of the least significant bit

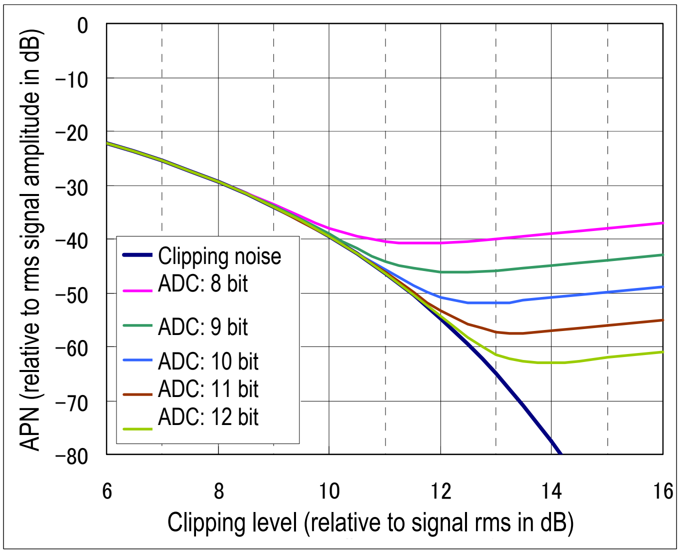
N: denotes the bit-length of A/D converter.

The noise power given by equations (8) and (9) is in proportion to square of clipping level, that is, the amplitude of the noise is in proportion to the clipping level. As the receivers usually adjust the signal amplitude by AGC to a certain level corresponding to the clipping level, we can regard these noises to be in proportion to the input signal level.

Figure 4 shows examples of amplitude proportional noise. It is seen from the figure that the clipping level should be set at approximately 4 times (+12 dB) of the *r.m.s.* signal amplitude and that the optimum clipping level depends on the bit-length of A/D conversion.

FIGURE 4

Amplitude proportional noise



1.3.3 Other amplitude proportional noises

There are a lot of amplitude proportional noise sources other than described above, such as phase jitter of PLL, computational errors, and so on. In addition, transmitted signals include APN such as inter-modulation components generated in the power amplifiers, noises included in the reception signal (in the case of relay stations with broadcasting wave relay), etc. We can treat these transmitter noises to be equivalently generated in the receiver, and we will express the APN including all noise sources in a single value relative to the *r.m.s.* signal amplitude.

1.4 Conditions that gives SFN failure

Now we will express the factors that are used in the digital reception estimation. They are:

$$\begin{aligned} CN &= CN_0 \times CN_{up}\left(U_{rms}/D_{rms}\right) \\ NP &= NP_{fix} + NP_{amp}\left(S_{rms}\right) \\ S_{rms} &= \sqrt{D_{rms}^2 + U_{rms}^2} \end{aligned} \tag{10}$$

where:

CN:	required CNR when SFN wave exists
CN_0:	required CNR without SFN waves
$CN_{up}(*)$:	CNR increase function defined by equation (7)
U_{rms}:	*r.m.s.* amplitude of SFN wave
D_{rms}:	*r.m.s.* amplitude of desired wave
S_{rms}:	*r.m.s.* amplitude of the input wave composed of the desired and SFN waves
NP:	sum of fixed noise and amplitude proportional noise
NP_{fix}:	noise power of fixed noise
$NP_{amp}(*)$:	noise power of amplitude proportional noise expressed as a function.

When the power of received signal is higher by CN times than the noise power calculated by equation (10), the digital signal is correctly received, as the CNR of the received signal being higher than the required one. On the contrary, when the signal is lower than the CN times of noise power, the digital signal cannot be correctly received, as the CNR of the received signal being lower than the required one. Hence, the condition that gives SFN failure can be obtained by resolving the following equation:

$$\begin{aligned} D_{rms}^2 &= CN \times NP \\ &= CN_0 \cdot CN_{up}\left(U_{rms}/D_{rms}\right) \times \left\{ NP_{fix} + NP_{amp}\left(\sqrt{D_{rms}^2 + U_{rms}^2}\right)\right\} \end{aligned} \tag{11}$$

Figure 5 shows examples of SFN failure conditions calculated by equation (11) in the case of APN of –40 dB. The area surrounded by the pink curve represents the conditions that give SFN failure. In the case of FEC 3/4, SFN failure does not take place even at the worst case of DUR = 0 dB when the desired signal is higher than a certain level, while in the case of FEC 7/8, it occurs around DUR = 0 dB regardless of the desired signal level.

FIGURE 5

Example of SFN failure conditions (64-QAM, APN = –40 dB)

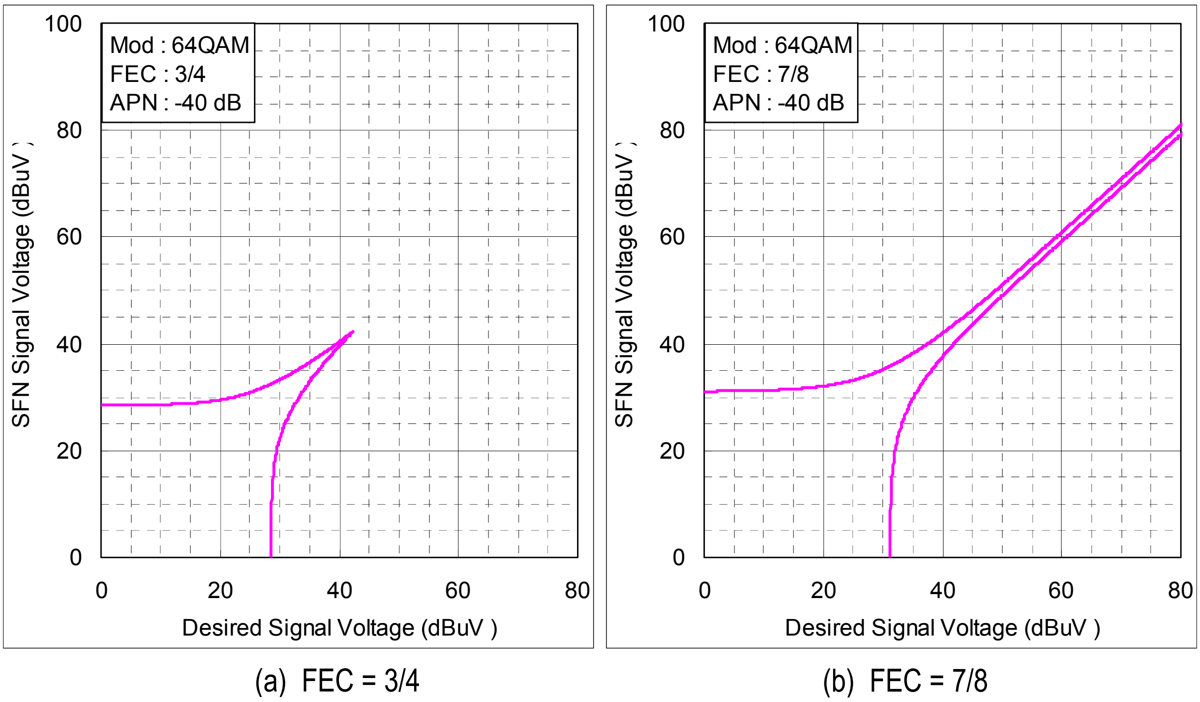
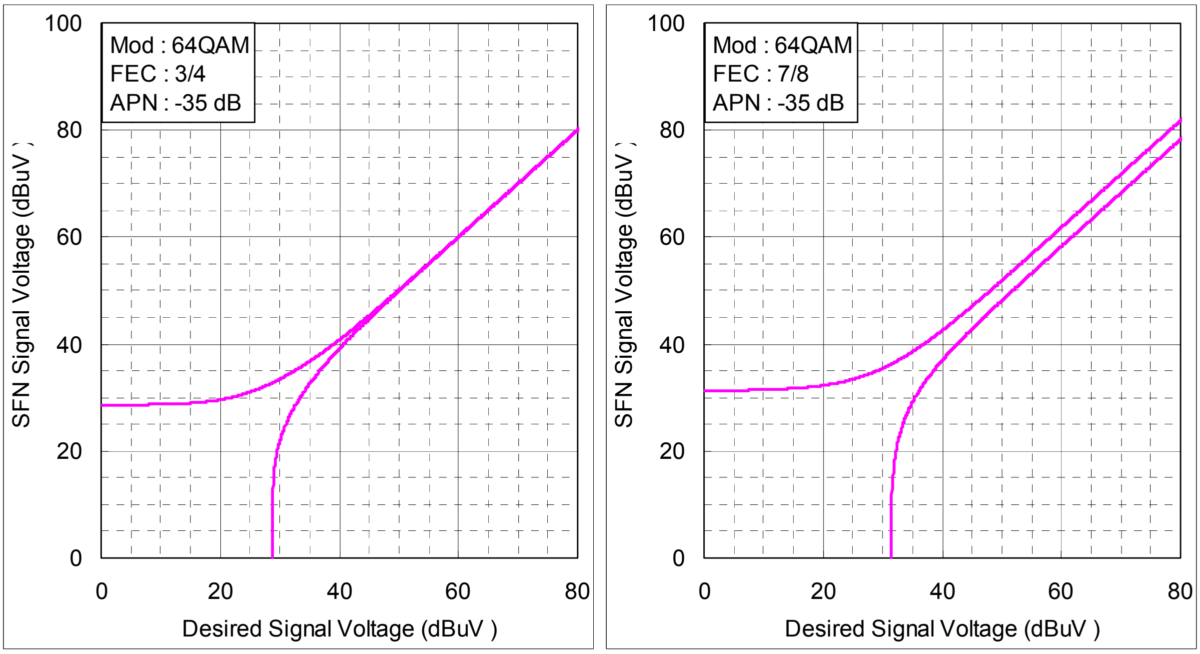


Figure 6 shows the examples of SFN failure in the case of APN of –35 dB. In this case, the failure takes place just on the condition of DUR = 0 dB even with FEC 3/4. The range of DUR that generates SFN failure becomes wider with FEC 7/8. The SFN failure conditions thus depend on the FEC applied as well as on the APN, which is one of the receiver characteristics to be specified.

FIGURE 6

Example of SFN failure conditions (64-QAM, APN = –35 dB)



(a) FEC = 3/4 (b) FEC = 7/8

1.5 SFN non-failure conditions

It is attractive that there exist the conditions without SFN failure, as shown in Fig. 5. We will give considerations on which conditions SFN failure does not occur. We will take into account only the APN, since the conditions without SFN failure are limited in the high signal voltage range in which fixed noise can be neglected. The maximum increase in the required CNR takes place at DUR = 0 dB, as shown in Fig. 2, and the value is equal to α in Table 1.

For example, the required CNR is calculated to be 32.1 dB by adding the CNR increase of 12.1 dB to the reference CNR of 20 dB in the case of 64-QAM-FEC3/4, and it is 50.2 dB (increase of 27.7 dB + reference of 22.5 dB) in the case of 64-QAM-FEC7/8.

The input signal level is 3 dB higher compare to the desired signal when DUR = 0 dB, and hence, the APN is also larger by 3 dB. Then we will obtain the conditions for SFN non-failure when the desired signal is higher than this increased CNR.

$$NP_{amp} \leq -\left(CN_0 + CN_{up}(\max) + 3\right) = -\left(CN_0 + \alpha + 3\right) \qquad \text{dB} \qquad (12)$$

It is derived by resolving Equation (12) that SFN non-failure conditions require the APN being less than –35 dB for 64-QAM-FEC3/4 and –53 dB for 64-QAM- FEC7/8.

It is seen in Fig. 4 that the APN of –35 dB can be realized with an 8-bit A/D converter, and that 11-bit or more is required to obtain APN of –53 dB. It may be difficult to realize APN to be less than –50 dB when taking into account the other noise sources such as jitter of PLL, etc. Thus it depends on the receiver performance whether or not the SFN failure takes place and we have to specify the reference receiver characteristics to be used in the design of broadcasting networks.

1.6 Location variation

If we can predict the accurate values of field strength of the desired and SFN waves, we can estimate whether or not the SFN failure takes place at the location concerned, by applying equation (12) and/or Fig. 5. However, we introduce statistic method, because it is impossible to predict the accurate field strengths.

Here we assume the mean and standard deviation of the field strength averaged in the area concerned to be μ_x and σ_x respectively for the desired wave, and those for the SFN wave to be μ_y and σ_y. Further assuming the correlation between the desired wave and the SFN wave to be ρ, then we can express the probability density of field strengths as follows:

$$F(E_x, E_y) = \frac{1}{2\pi\sqrt{A}} \exp\left[-\left\{a_{11}(E_x - \mu_x)^2 + 2a_{12}(E_x - \mu_x)(E_y - \mu_y) + a_{22}(E_y - \mu_y)^2\right\}\right]$$
$$\text{where,}\quad A = \sigma_x^2 \sigma_y^2 (1 - \rho^2), \quad a_{11} = \sigma_y^2 / A, \quad a_{12} = -\rho\,\sigma_x \sigma_y / A, \quad a_{22} = \sigma_x^2 / A \tag{13}$$

We can calculate the probability of SFN failure occurrence at a particular location by integrating equation (13) over the conditions that give SFN failure. Then the probability of SFN failure occurrence is written by equation (14).

$$P(\mu_x, \mu_y) = \iint\limits_{\text{SFN failure conditions}} F(E_x, E_y)\,\mathrm{d}E_x\,\mathrm{d}E_y \tag{14}$$

Figures 7 and 8 show the calculation results of equation (14) for 64-QAM-FEC3/4 and for 64-QAM-FEC7/8, respectively. The probability of the failure for 64-QAM-FEC3/4 is almost zero in the region of received voltage being higher than 45 dB(µV), while it remains at 10-20% even in high reception voltage regions for 64-QAM-FEC7/8.

FIGURE 7

Example of probability of SFN failure occurrence (64-QAM-FEC3/4)

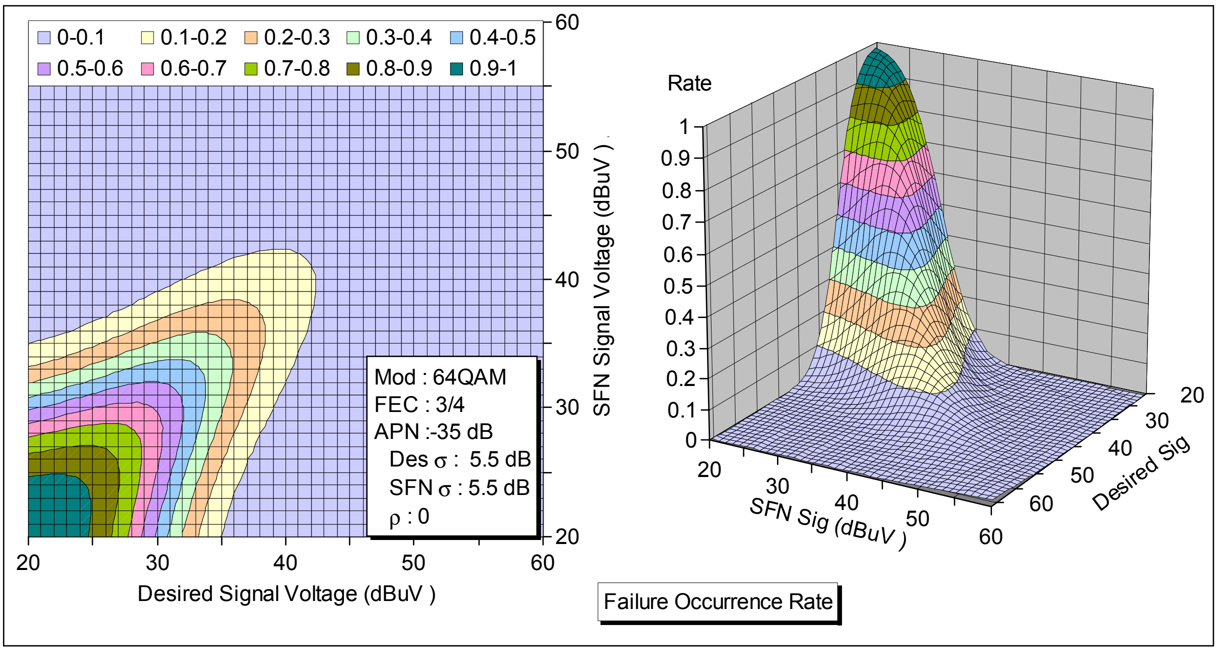
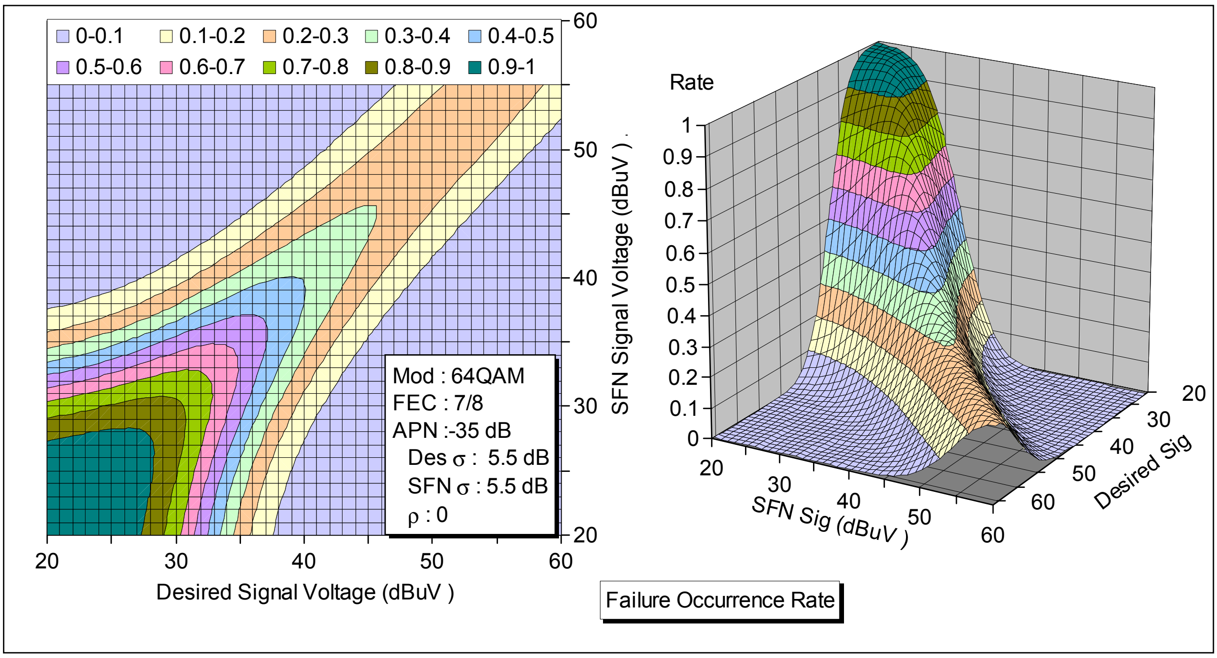


FIGURE 8

Example of probability of SFN failure occurrence (64-QAM-FEC7/8)



We apply equation (14) as the basic algorithm of "digital reception simulator", which is used in the channel plan calculation, especially for small-scale transmission stations.

1.7 Location correlation

The field strength distributions of the desired wave and of the SFN wave can be assumed to exhibit the same statistics each other, although they come from different directions. Recommendation ITU-R P.1546 describes that the standard deviation of field strengths for digital broadcasting waves to be 5.5 dB. Considering that the field strength distribution is caused mainly by the clutters near the reception point, such as terrain, trees, buildings, etc., field strength seems to have correlation between the desired and SFN waves. For example, the field strength would be lower at a hollow in the ground regardless of wave directions, or it would be higher at a top of hill. The correlation would be stronger with closer the directions.

Figure 9 shows changes in the probability of SFN failure occurrence by the correlation ρ. We don't know at this moment which values of correlation are to be applied, and further study is required.

2 In the case of multiple SFN waves

The ARIB document gives some considerations on the cases where multiple SFN waves exist. It discusses how to deal with multiple SFN waves. The CNR increase function, for example, varies largely depending on the number of waves, amplitude and phase of waves and so on. A computer simulation is carried out, in which a number of SFN waves with random amplitudes and random delay/phase are generated, and it is shown that in most case, the CNR increases are less than the values obtained in the previous section. It is also shown that 90% of cases are covered with additional 2 dB of the CNR increase function defined in the previous section.

Considerations are also given on the actual reception conditions, and the section concludes that we can apply the results described in the previous section in statistic sense. The section is not directly related to the reference receiver characteristics and is omitted.

2.1 How to specify multiple SFN waves

2.2 Failure regions with two SFN waves

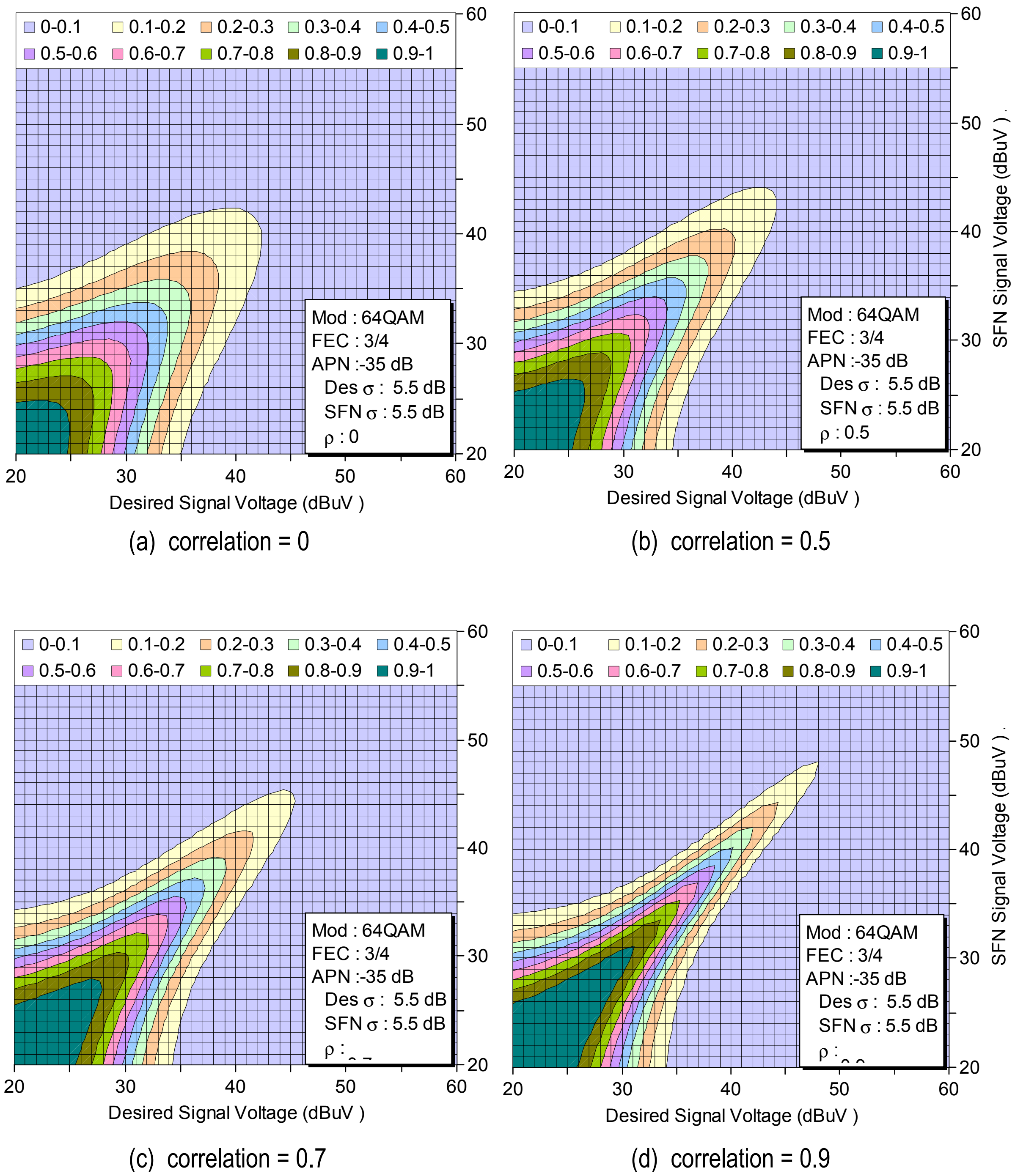
2.3 Failure regions with multiple SFN waves

2.4 Consideration on multi-path conditions

2.5 "Appropriate function" to be applied in the simulation.

FIGURE 9

Changes in probability of SFN failure by location correlation



Chapter II – SFN with delays exceeding guard interval duration

When the delay of SFN wave exceeds the guard interval duration, the orthogonal features among OFDM carriers are lost, resulting in a large increase in the required DUR. The Japanese channel plan applies DUR = 28 dB including margins for several factors. It may be reasonable for the basic channel plan to include margins, because the basic plan should guarantee stable networks operation. If there were enough room in the spectrum, we could apply the plan values for each of transmission stations. But there is not enough spectrum in the real world, hence we have to establish the appropriate values, which are required especially for small-scale transmission stations.

3 In the case of SFNs with delays exceeding the guard interval duration

First we will analyze the characteristics of OFDM signals under the existence of SFN waves having delays exceeding guard interval duration. Then we will discuss about the receiver characteristics to be specified. We call the SFN with delays exceeding guard interval duration as "outer-GI SFN" in short.

3.1 SFN wave with a large delay exceeding guard interval duration

Since SFN signal with a very large delay has no correlation to the desired signal, we can treat it as random noise. The required DUR is calculated by the following equation:

using the relations:

$$CN_0 = \frac{C_{rms}^2}{N_{fix}^2 + U_{rms}^2} \quad \text{and} \quad DU_0 = \frac{C_{rms}^2}{U_{rms}^2}$$
$$DU_0 = \frac{CN_0}{1 - CN_0 \cdot \left(N_{fix}^2 / C_{rms}^2\right)} \tag{15}$$

where:

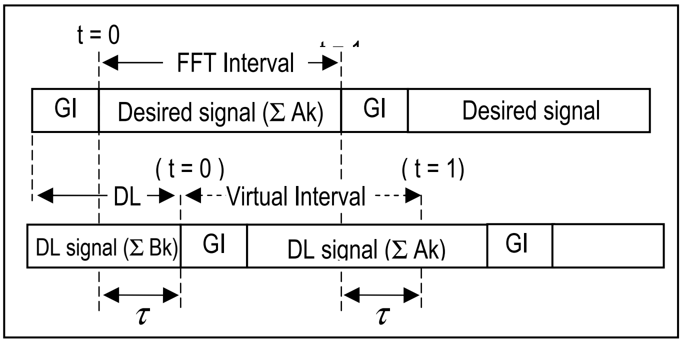
- C_{rms} and U_{rms}: denote the *r.m.s.* amplitude of the desired wave and of the SFN wave respectively
- N_{fix}: denotes the fixed noise of receiver
- CN_0: the ratio of desired signal to unnecessary components (the same values as the reference CNR)
- DU_0: the required DUR.

3.2 SFN with relatively small delays (but larger than guard interval duration)

We will analyze the demodulated signals when an outer-GI SFN wave exists. We assume the amplitude of delayed signal to be U and the portion of the delay exceeding GI to be τ (normalized by GI), as shown in Fig. 10.

FIGURE 10

OFDM demodulation



The demodulated signals are obtained by multiplying each of OFDM carriers to the input signal and then integrating each product over the FFT interval. Assuming the original OFDM signal to be expressed by equation (16), then the demodulated component for the main signal (X) and that for delayed signal (Y) are written as below:

$$S_o = \sum A_k \exp(j2\pi kt) \tag{16}$$

$$X = \int_0^1 \sum A_k \exp(j2\pi kt) \times \exp(-j2\pi k_o t)\,\mathrm{d}t = \int_0^1 \sum A_k \exp(j2\pi(k - k_o)t)\,\mathrm{d}t = A_{ko} \tag{17}$$

$$Y = U\left[\int_0^\tau \sum B_k \exp(j2\pi kt) \times \exp(-j2\pi k_o t)\,\mathrm{d}t + \int_\tau^1 \sum A_k \exp(j2\pi kt) \times \exp(-j2\pi k_o t)\,\mathrm{d}t\right] \tag{18}$$

Taking into account the guard interval, the integral of the 2nd term in equation (18) has the same result as integrating over "virtual interval" instead of "FFT interval" shown in the Fig. 10.

$$\begin{aligned} \textit{2nd term} &= U\left[\int_0^1 \sum A_k \exp(j2\pi(k - k_o)t)\,\mathrm{d}t - \int_{1-\tau}^1 \sum A_k \exp(j2\pi(k - k_o)t)\,\mathrm{d}t\right] \\ &= UA_{ko} - U\int_{-\tau}^0 \sum A_k \exp(j2\pi(k - k_o)t)\,\mathrm{d}t \end{aligned} \tag{19}$$

$$Y = UA_{ko} + U\int_0^\tau \sum B_k \exp(j2\pi(k - k_o)t)\,\mathrm{d}t - U\int_{-\tau}^0 \sum A_k \exp(j2\pi(k - k_o)t)\,\mathrm{d}t \tag{20}$$

The 1st term in equation (20) represents the signal component, the 2nd term inter-symbol interference, and the 3rd term inter-carrier interference. Thus, the delayed signal, of which delay exceeds the guard interval, includes the signal component, and we call it as "effective inner-GI component". This component interferes with the main signal and produces the ripples in frequency response, resulting in the required CNR increase. This can be treated in the same way as the inner-GI SFN wave with an exception of interfering duration, which is $(1 - \tau)$ instead of FFT Interval.

The inter-symbol interference, i.e., the 2nd term in equation (20) has no correlation with the signal component, it can be treated as random noise of which power increases in proportion to τ.

The expectation of 3rd term increases in proportion to τ when τ is close to zero, although the value itself varies according to the signal components A_k. Since the 3rd term equals to signal component when $\tau = 1$, the expectation of interference component is in proportion to $(1 - \tau)$ when τ is close to 1. Thus, the inter-carrier interference, i.e., the 3rd term in equation (20) can be regarded as random noise of which power increases in proportion to $\tau(1 - \tau)$. We call these components that have no correlation to the signal as "effective outer-GI component", and we can treat them in the same way as random noise.

Taking into account the above features of the 2nd and 3rd terms in equation (20), we can obtain the required DUR under outer-GI SFN environments by resolving equation (21).

$$Required\,CNR = CN_0 + CN_{up}\left(U\mathrm{dB} + 10 \cdot \log\,(1-\tau)\right) \qquad \mathrm{dB}$$

$$Noise\,Component = \tau U_{rms}^2 + \tau(1-\tau)U_{rms}^2 + N_{amp}^2 \qquad \text{(in real number expression)}$$

$$-10 \cdot \log\left(\tau U_{rms}^2 + \tau(1-\tau)U_{rms}^2 + N_{amp}^2\right) = CN_0 + CN_{up}\left(U\mathrm{dB} + 10 \cdot \log(1-\tau)\right) \qquad (21)$$

where:

U_{rms}^2: denotes power of delayed wave (dB)

N_{amp}^2: power of APN (dB)

CN_0: reference CNR (dB)

$CN_{up}(*)$: CNR Increase Function (dB)

UdB: power of delayed wave (dB).

The solution of equation (21) at $\tau = 0$ corresponds to the required DUR for inner-GI SFN environment. When the equation has no solution at $\tau = 0$, the SFN failure does not takes place.

Equation (22) is a concrete example of equation (21), and Fig. 11 shows the corresponding graphs for the equation. Figure 12 shows the relationship between delay and required DUR, which is called "bathtub curve".

$$\begin{aligned} &Required\ CNR = CN_0 + \alpha \cdot \exp\left(-\left|\beta \cdot (UdB + 10 \cdot \log(1-\tau))\right|^{\gamma}\right) \\ &Noise\,Component = \left(2\tau - \tau^2\right)10^{UdB/10} + 10^{NampdB/10} \end{aligned} \qquad (22)$$

where, α, β and γ are shown in Table 1 in § 1.2.

3.3 Aliasing affects of scattered pilot signals

In the previous section, we have analyzed the interference components in the OFDM demodulation process, in which we assumed implicitly that the rippled frequency response was ideally compensated by some means. This compensation process is equivalent to regenerate each of OFDM reference carriers from the received scattered pilot signal (SP hereinafter). As the SP signals are sent in every 3 OFDM carriers, the minimum frequency spacing in the ripples measured with the SP signals is limited, that is, the maximum delay that can be observed is limited. If there is a delayed wave exceeding this maximum value, correct observation cannot be performed, just as aliasing found in general sampling process.

FIGURE 11

Required CNR and interference component under outer-GI environment

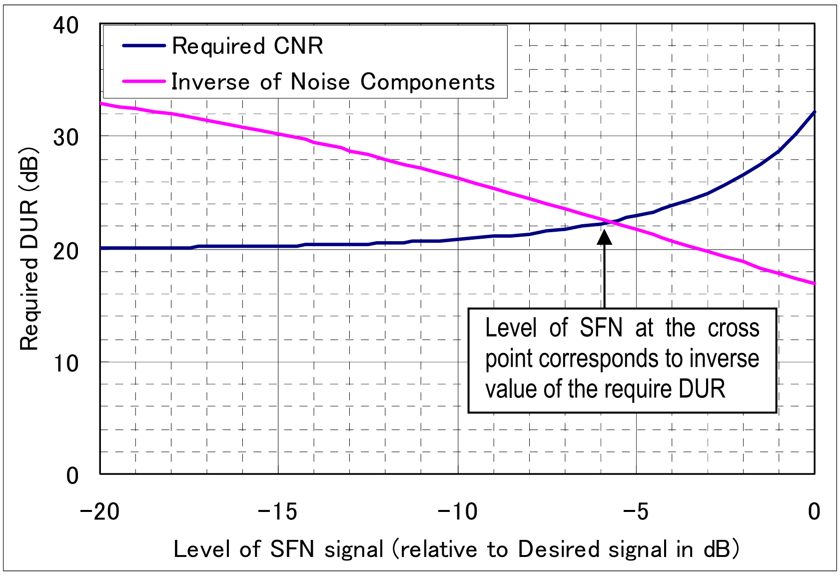


FIGURE 12

Bathtub curve (calculated for APN of –35 dB)

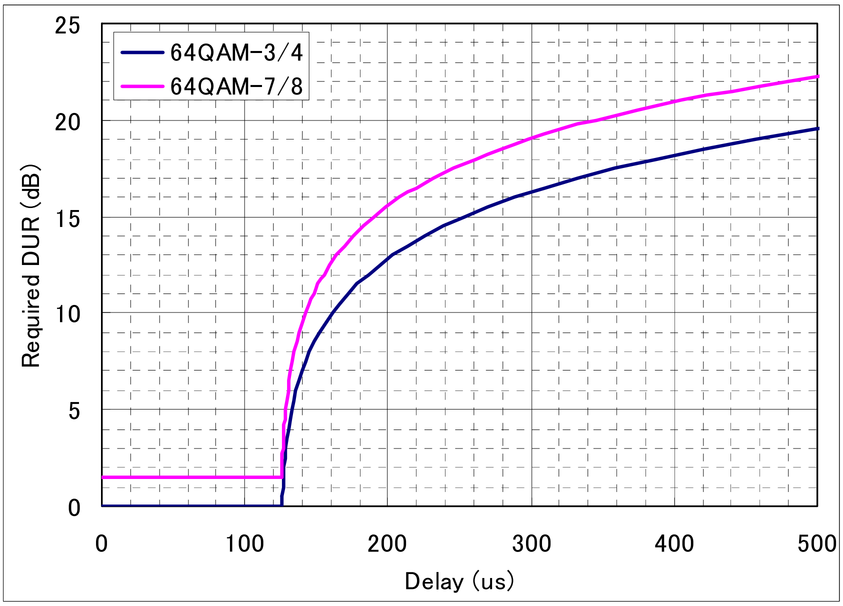
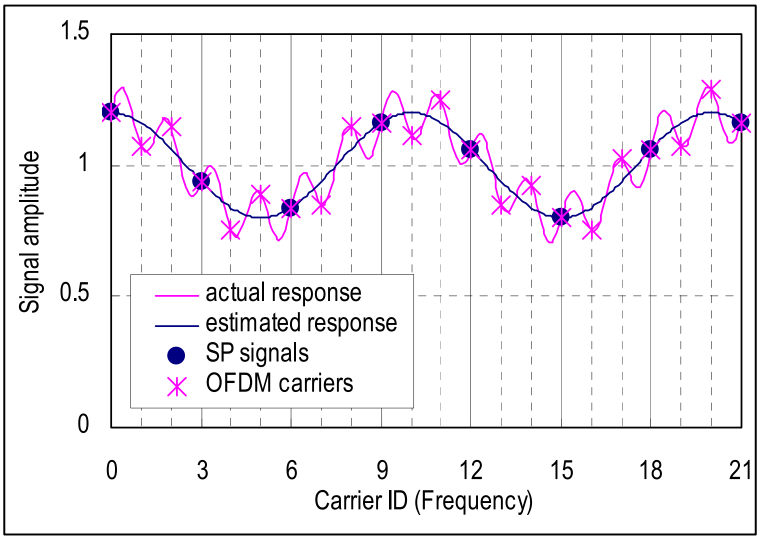


Figure 13 explains the above circumstance of SP signal. The figure shows the frequency response when delayed waves with a long delay and a short delay co-exist. The frequency response has fine ripples due to the long delay wave, and the OFDM carriers take the values as shown by * marks in the figure. On the other hand, we will estimate the ripples to be the rough curve shown in the figure, because we can observe the response sampled only at the SP signals. Therefore, the carrier references to be used in the demodulation include errors in phase and amplitude.

FIGURE 13

Frequency response estimated from SP signals



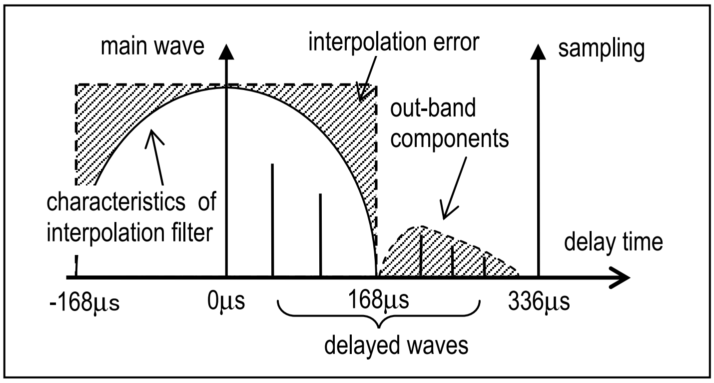
The average error can be estimated in the same way as the general analysis of distortions included in the signal recovered from the sampled values, although the error of each carrier varies depending on the delay and amplitude of SFN waves, characteristics of the interpolation filter, and so on. The average error (A_e^2) can be written as follows:

$$A_e^2 = \sum outband\ components + \sum interpolation\ error\ components \qquad (23)$$

Note that in analyzing a sampling system, it is usual way that the original function is the time-domain function and the conversion function is Fourier spectrum, while in equation (23), the original function is Fourier spectrum and the conversion function is the time-domain one. Figure 14 shows the out-band components and interpolation errors used in equation (23). The figure is the case for the FFT interval of 1 008 µs, and the value 336 µs corresponds to 1/3 of the FFT interval.

FIGURE 14

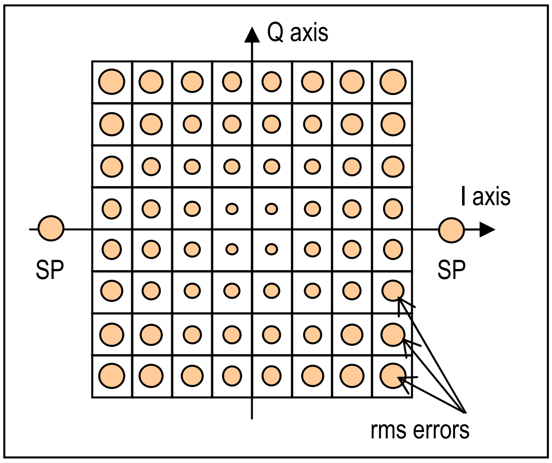
Delay profile, out-band components and interpolation errors



The affect of carrier recovery errors differs by signal points on the constellation, as shown in Fig. 15. It is large on the signal points at outer side of the constellation. To simplify the calculation, we assume here that the *r.m.s.* error at every signal point has the same error at the SP point. This simplification results that the calculated BER tends to become worse than that of actual one.

FIGURE 15

Affects of carrier recovery errors



3.4 Calculation method for the required DUR

When there are many kinds of interference factors, the required DUR can be written in general form as below:

when:

$$\frac{1}{CN_0} = W_1\left(U^2\right) + W_2\left(U^2\right) + \bullet\bullet\bullet$$

$$Required\ DUR = -10\log\left(U^2\right) \quad \text{dB} \tag{24}$$

where:

CN_0: denotes the required CNR that is determined by modulation and FEC (such as 20 dB for 64-QAM-FEC3/4)

U^2: denotes the power of delayed signal

$W_1(*)$, $W_2(*)$,etc.: represent the weighting functions which convert the delayed signal into the equivalent noise power against each interference factor.

The major factors that affect on the correct reception are inter-symbol interference (2nd term in equation (20)), inter-carrier interference (3rd term), and the carrier recovery error which we are discussing in this section.

We apply equation (24) as below:

$$\left[CN_0 \cdot CN_{up}\left(U_{rms}^2\right)\right]^{-1} = N_{amp}^2 + \tau \cdot U_{rms}^2 + \tau\,(1-\tau) \cdot U_{rms}^2 + (1-\tau) \cdot A_e^2 \tag{25}$$

The 2nd, 3rd and 4th terms in the right side of equation (25) correspond to inter-symbol interference, inter-carrier interference, and carrier recovery errors, respectively. Note that we neglect the fixed noise assuming the received signal to be enough high. Resolving equation (25) for U^2, the required DUR is obtained by the inverse number of U^2.

The hardware characteristics that affect on the receiver performance are N_{amp}^2 and A_e^2 in equation (25), and the other terms are independent of hardware characteristics. Therefore, we have to specify these two for the reference receiver to be applied in network design.

The actual values of amplitude proportional noise were at –35 dB for the worst receivers available on the market. So we have adopted this –35 dB as the specification of APN.

Before we specify the interpolation filter characteristics, we will take an overview on the filter characteristics. When assuming an ideal LPF as the interpolation filter, equation (25) is written as below:

$$\begin{aligned}
\left[CN_0 \cdot CN_{up}\left(U_{rms}^2\right)\right]^{-1} &= N_{amp}^2 + \tau U_{rms}^2 + \tau(1-\tau)U_{rms}^2 && (|\text{delay}| < 168\mu\text{s}) \\
\left[CN_0 \cdot CN_{up}\left(U_{rms}^2\right)\right]^{-1} &= N_{amp}^2 + \tau U_{rms}^2 + \tau(1-\tau)U_{rms}^2 + (1-\tau)U_{rms}^2 && (|\text{delay}| \geq 168\mu\text{s})
\end{aligned} \tag{26}$$

In this case, the OFDM carriers can be recovered perfectly and no recovery errors take place when the delay is within the Nyquist band. When the delay is out of Nyquist band, A_e^2 equals to the power of the delayed wave itself.

In the case where the interpolation filter is not an ideal LPF, the power of interpolation errors (A_e^2) is calculated by the below:

$$A_e^2 = \sum_{|DL_j|<168\mu s} \left[\left(1 - LPF(DL_j)\right) \cdot U_j\right]^2 + \sum_{|DL_k| \geq 168\mu s} U_k^2 \tag{27}$$

where:

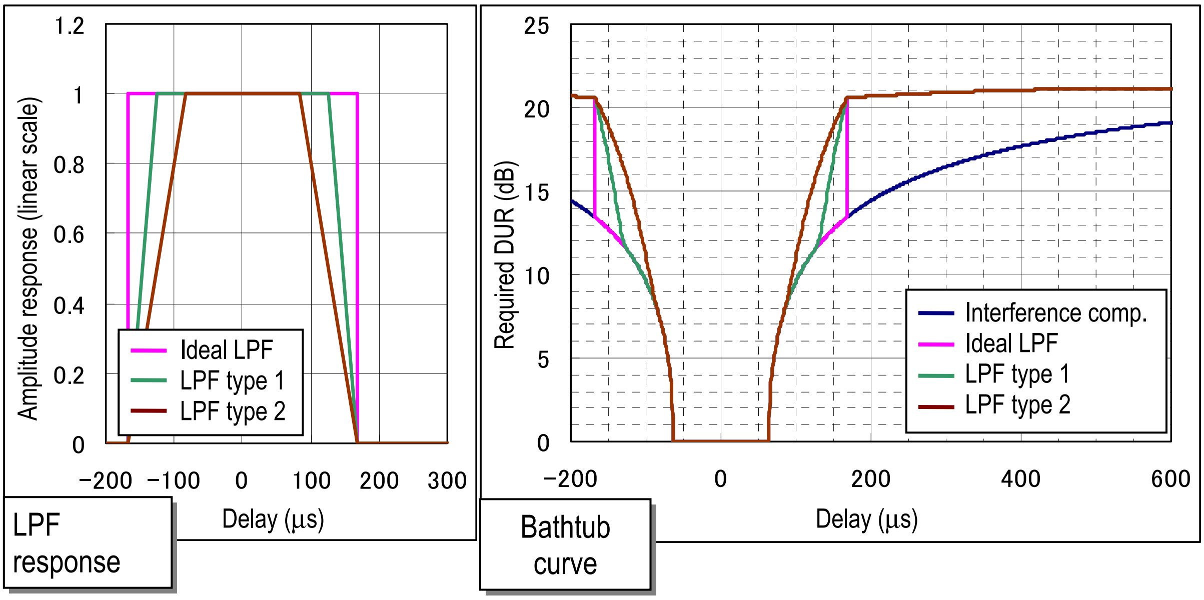
U_j: represents the amplitude of delayed waves within the Nyquist band ($|DL| < 168\mu s$)

U_k: the amplitude of delayed waves outside the Nyquist band ($|DL| > 168\mu s$).

Figure 16 shows examples of calculation results for some interpolation filters.

FIGURE 16

Interpolation filter and corresponding DUR characteristics (by equation (27))



3.4.1 Re-consideration on aliasing affects of scattered pilot signals

We have assumed the aliasing affects of SP signals to be in proportion to the average error power A_e^2. However, the required DUR of actual receivers is higher by 1 dB than that calculated by equation (25). This suggests that the carrier recovery errors are affected by other factors in addition to the average error power.

The recovery error for each OFDM carrier corresponds to the error voltage of that carrier but not to the *r.m.s.* error voltage. The error voltage differs from carrier to carrier, and hence the BER differs carrier by carrier. In these cases where the error voltage differs by carrier, the total BER averaged over all carriers becomes worse compared to the case where every carrier takes the same error voltage. This is because the BER degradation of the carriers with error voltages larger than the average is more severe compared to the BER improvement of the carriers with smaller error voltages. Therefore we need to add a new term in equation (25) to express the effects due to error voltages being different from carrier to carrier.

However, it is quite difficult to derive theoretically such effects. Considering that the effects may differ depending on the delay and/or phase of the input waves, and that the amount of effect is only 1-2 dB, we introduce a new coefficient to be consistent with the measured results, as below:

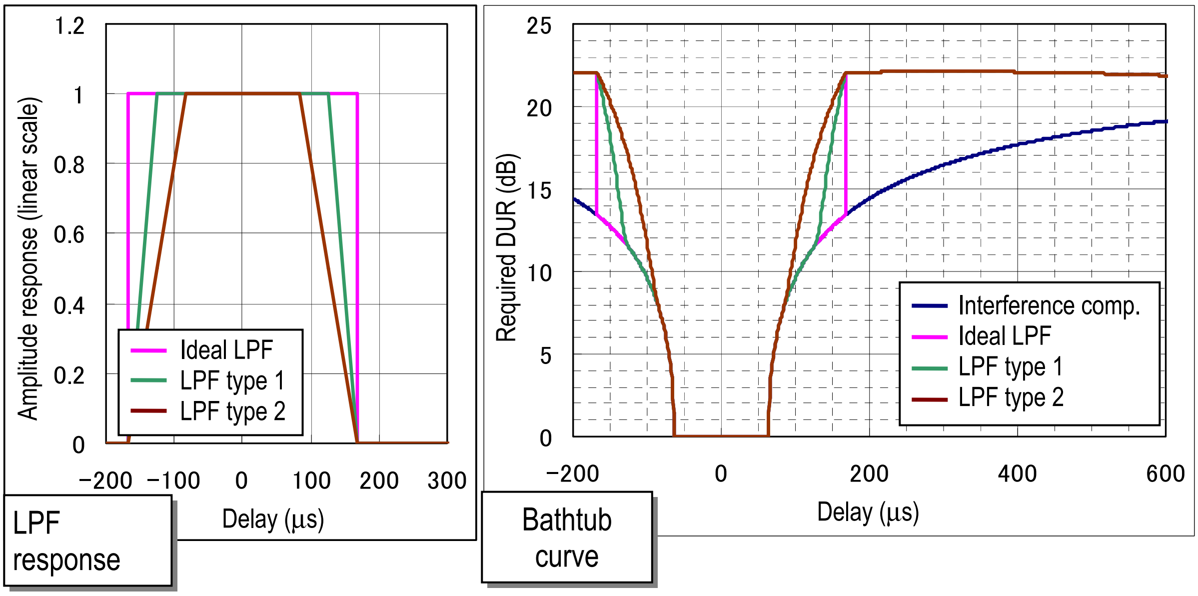
$$\left[CN_0 \cdot CN_{up}\left(U_{rms}^2\right)\right]^{-1} = N_{amp}^2 + \tau \cdot U_{rms}^2 + \tau\left(1-\tau\right)\cdot U_{rms}^2 + \underline{1.5}\cdot\left(1-\tau\right)\cdot A_e^2 \qquad (28)$$

here, the value for the new coefficient to be 1.5.

Figure 17 shows the examples calculated by equation (28).

FIGURE 17

Interpolation filter and corresponding DUR characteristics (by equation (28))



3.5 Setting of FFT window

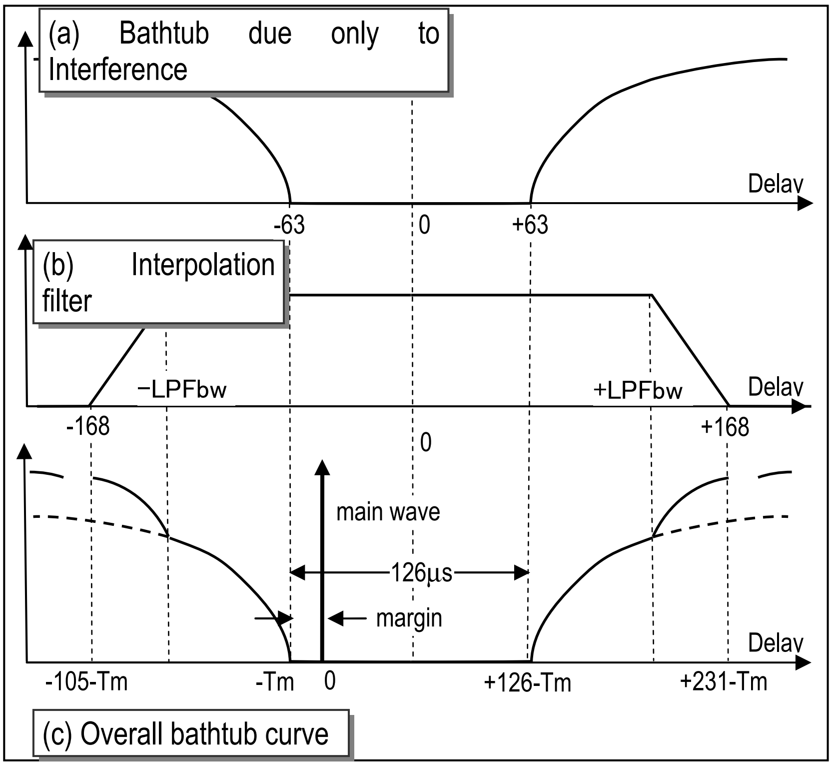
Figure 18 shows the relationship between the FFT window in the receiver and the required DUR characteristics or the bathtub curve. The figure is the case of guard interval of 126 µs (1/8 of FFT frame duration).

Figure 18 a) shows the bathtub curve due only to the interference components, i.e., inter-symbol interference and inter-carrier interference. Figure 18 b) represents the characteristics of interpolation filter, which has a unity response within the bandwidth of ±LPFbw and zero response outside the Nyquist band (±168 µs). Figure 18 c) shows the overall bathtub curve, which includes all the factors.

The receiver adjusts its FFT window so that the main wave is positioned at the bottom of the bathtub curve. Otherwise a large inter-symbol interference takes place. Therefore, the range of the FFT window is limited within $\pm GI/2$ (GI denotes guard interval duration). In the actual hardware, some margins are necessary to avoid miss-adjustment, and the FFT window range is limited within $\pm(GI/2 - T_m)$ as shown in the figure. Note that the position of “0 µs” in Fig. 18 c) is different from Figs 18 a) and 18 b), as it is a custom to express the delays relative to the main wave.

FIGURE 18

FFT window and bathtub curve



3.5.1 Optimum position of the FFT window

Here we summarize how to estimate the reception failure.

1. To split each SFN wave into effective inner-GI component and effective outer-GI component.

   Regarding effective inner-GI component, we assume one equivalent SFN wave of which power is equal to the sum of each effective inner-GI component, that is,

$$U_{eq}^2 = \sum_k (1-\tau_k) U_k^2 \tag{29}$$

   Regarding effective outer-GI component, we assume equivalent noise, which is the sum of inter-symbol interference, inter-carrier interference and carrier recovery error of each SFN wave, that is,

$$N_{eq}^2 = \sum_k \tau_k U_k^2 + \sum_k \tau_k (1-\tau_k) U_k^2 + 1.5 \cdot \sum_k (1-\tau_k)(1-LPF(DL_k))^2 U_k^2 \tag{30}$$

2. To sum up all of the non-correlated noise powers, which are fixed noise, co-channel digital waves (with different programs) and co-channel analogue waves in addition to equation (30), that is,

$$N_{total}^2 = N_{fix}^2 + N_{eq}^2 + \sum_k CoD_k^2 + \sum_k CoA_k^2 \big/ AtoD^2 \tag{31}$$

where, N^2_{fix} denotes the fixed noise, CoD^2_k the power of co-channel digital waves with different programs, CoA^2_k the power of co-channel analogue waves, and $AtoD^2$ the weighting factor with which the analogue signal is dealt as equivalent random noise.

3. To calculate the probability of failure occurrence, using equations (11) and (14), in which the total noise power is given by equation (31). In this case, the optimum position of the FFT window is defined to be such one that minimizes the value of equation (31).

4. To apply the following method instead of the above § 3, as it is not convenient to calculate equation (11).

5. To plot the signal voltages of the desired wave and the equivalent inner-GI SFN wave on the 2-dimensional voltage diagram, as shown in Fig. 19 (green coloured mark). To shift this input point to the point shown by brown coloured mark (left down direction). The value of the sift is given by the following equation:

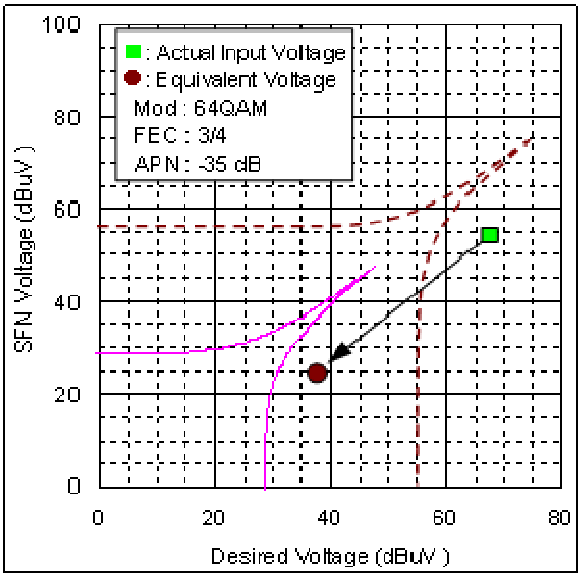
$$10 \log\left(N^2_{total} / N^2_{fix}\right) \quad \text{dB} \tag{32}$$

6. To apply equation (14) against the equivalent input voltage i.e., the shifted point in the figure.

That is the algorithm used in the “digital reception simulator”.

FIGURE 19

Equivalent input signal voltages

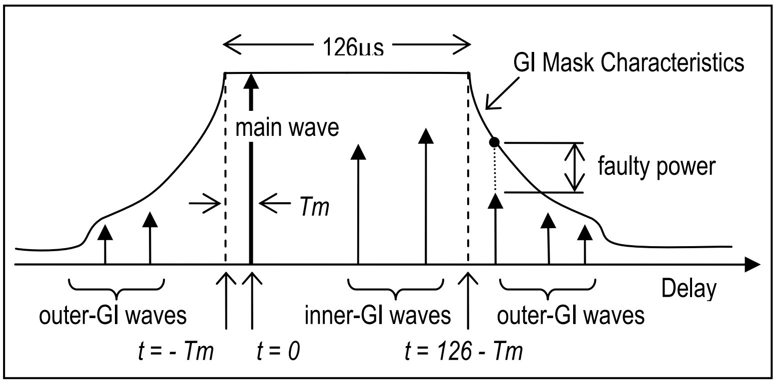


The optimum position of FFT window is stated in the above § 3, but it is not useful to apply equation (31) from the view point of computation hours. Therefore we introduce the following alternative method.

Figure 20 shows the relationship between the input waves and FFT window position. The curve named “GI Mask characteristics” in the figure is the inverse bathtub curve (opposite polarity in dB), which indicates the maximum allowable levels of delayed signals. The receiver adjusts its FFT window so that all delayed waves become below the mask. If a delayed signal exceeds the mask, the receiver cannot receive the signal correctly.

FIGURE 20

Optimum setting of FFT window



When there are many delayed signals of which amplitudes are close to (still below) the mask, the receiver may fail the correct reception. So, we introduce a new concept of "faulty power", which is defined to be the difference in dB between the amplitude of delayed signal and the corresponding mask. When the sum of the faulty power is larger than 0 dB, we regard the delayed signals to exceed the mask as a whole. We also assume that the receiver adjusts its FFT window to minimize the total faulty power, as written in the following equations:

$$
\begin{aligned}
&PdB_k = UdB_k^2 - MaskdB(DL_k) \qquad (\mathrm{dB}) \\
&P_{und} = \sum_k 10^{PdB_k/10} \rightarrow \text{minimize}
\end{aligned}
\tag{33}
$$

3.6 Protection ratios for analogue to digital interference

The affects of interference from analogue signal depends not only on the receiver characteristics but also the analogue signal or analogue program contents. We consider here the protection ratio against the worst-case analogue signals.

We have tested several receivers on the protection ratios, in which we apply colour-bar with 100% modulated stereo signal as the worst-case analogue signal. The measured results for the worst receiver is as follows.

for 64-QAM-FEC3/4	protection ratio of 5 dB
for 64-QAM-FEC7/8	protection ratio of 13 dB.

3.7 Receiver characteristics to be specified

As discussed above, many of the factors related to SFN reception are caused by OFDM itself, and are automatically defined by the signal parameters applied. The factors caused by hardware, which should be specified for the reference receiver are listed in Table 2 together with ARIB specifications.

TABLE 2

Factors to be specified

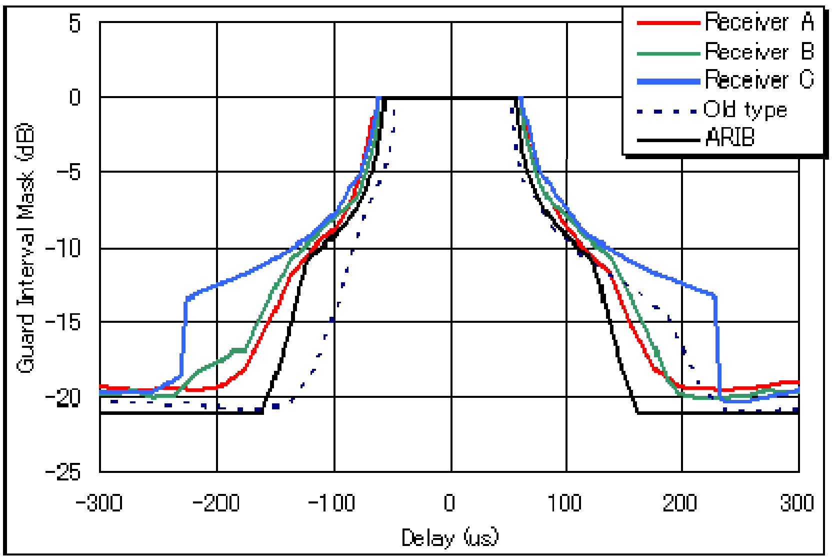
Item	ARIB specification	Remarks
Amplitude proportional noise	–35 dB	Relative to input signal level
Interpolation filter for carrier recovery	Flat	–126 μs ~ 126 μs
	Transition	–168~ –126 μs and 126~168 μs
FFT window setting margin	6 μs	
Protection ratio interfered by analogue TV	5 dB	64-QAM-FEC3/4
	13 dB	64-QAM-FEC7/8

3.8 GI Mask characteristics of receivers on the market

Figure 21 shows examples of guard interval mask of receivers available on the market together with the characteristics (denoted by "ARIB" in the figure) specified in § 3.7. Receivers except for the one expressed by dot line in the figure exhibit better characteristics than the ARIB specification. The dot-lined receiver is one that was produced before the ARIB specification has been established. The characteristic difference among receivers comes from the characteristics of interpolation filter used for reference carrier recovery (see § 3.3).

FIGURE 21

GI Mask of receivers available on the market



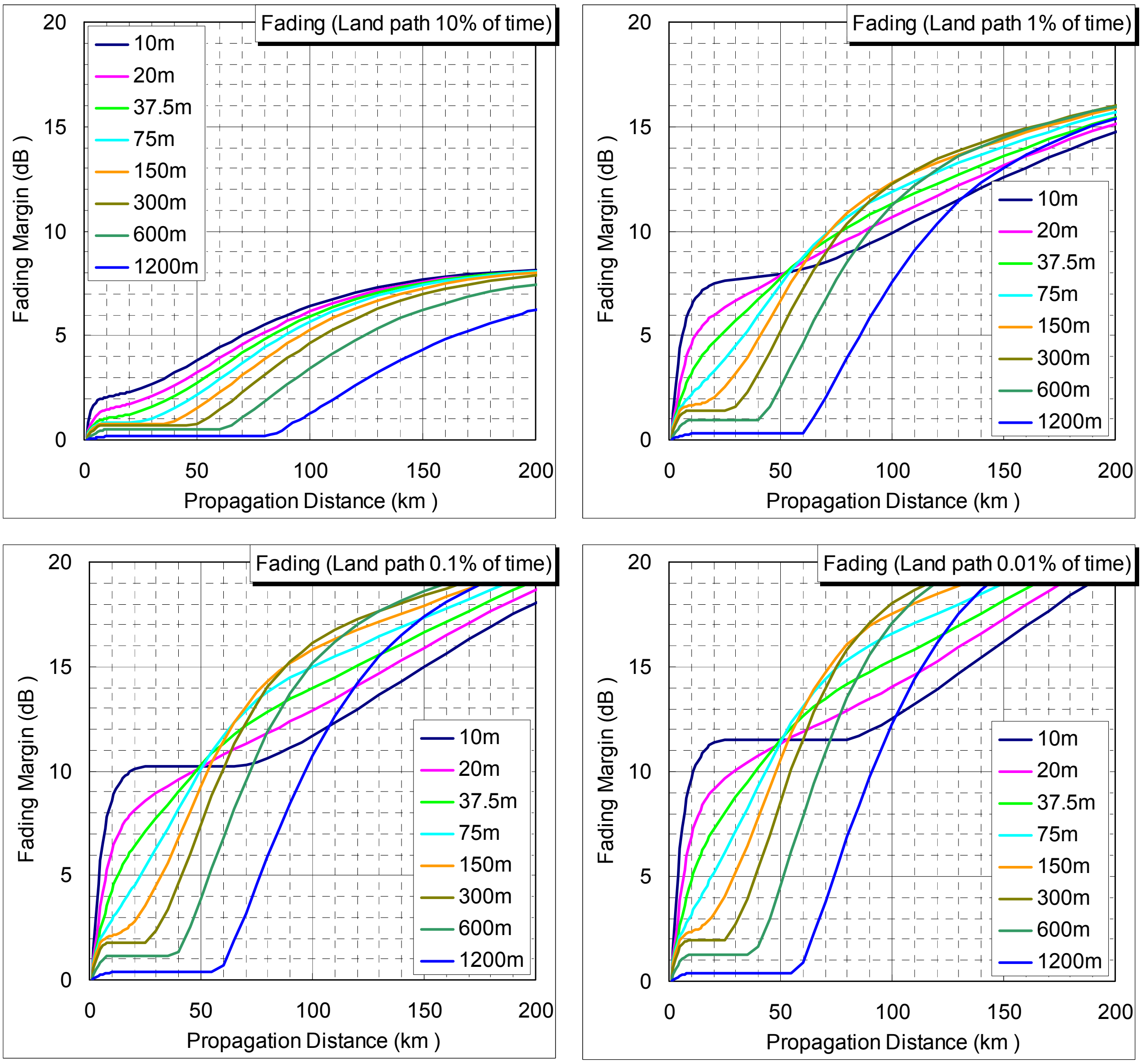
Chapter III – Fading

Recommendation ITU-R P.1546 gives the relationship of reception field strength vs. propagation distance for various time availabilities. We assume here fading margins to be the difference between the field strengths of the designated time availability and that of 50% time (annual mean value). The Recommendation states that the data is valid for the time availability only between 50% and 1%. For the range less than 1% of time, we assume so-called Kumada's law, details of which are described in Appendix 1.

Figure 22 shows the fading margins derived from the above. The curves in figures correspond to the transmission antenna heights. Since these fading margins are derived from the Recommendation, the general treatments, such as interpolation of time availability, antenna height, etc. are to follow the Recommendation.

FIGURE 22

Fading margins (600 MHz)



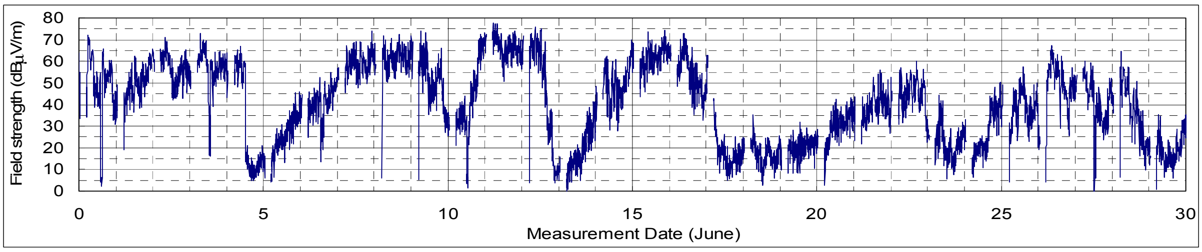
Appendix 1

Fading margin for less than 1% of time (Kumada's law)

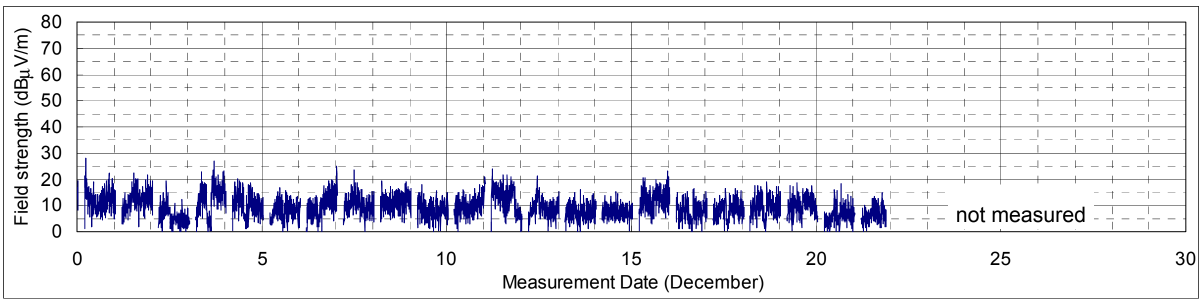
Figure 23 shows examples of measurement results of field strength coming form Korea to Japan. The analogue TV waves emitted at several cities in Korea were continuously measured for one year period. Figure 23 a) is the results measured during June and Figure 23 b) during December. The field strengths present big changes in June while almost constant in December. In this area, fading phenomenon takes places very often during May to October and it is rarely during December to March.

FIGURE 23

Measurement examples of field strengths coming from Korea (sea path)



(a) Fading season (June 2001)



(b) Non-fading season (December 2001)

FIGURE 24

Cumulative percentage of field strengths coming from Korea (sea path)

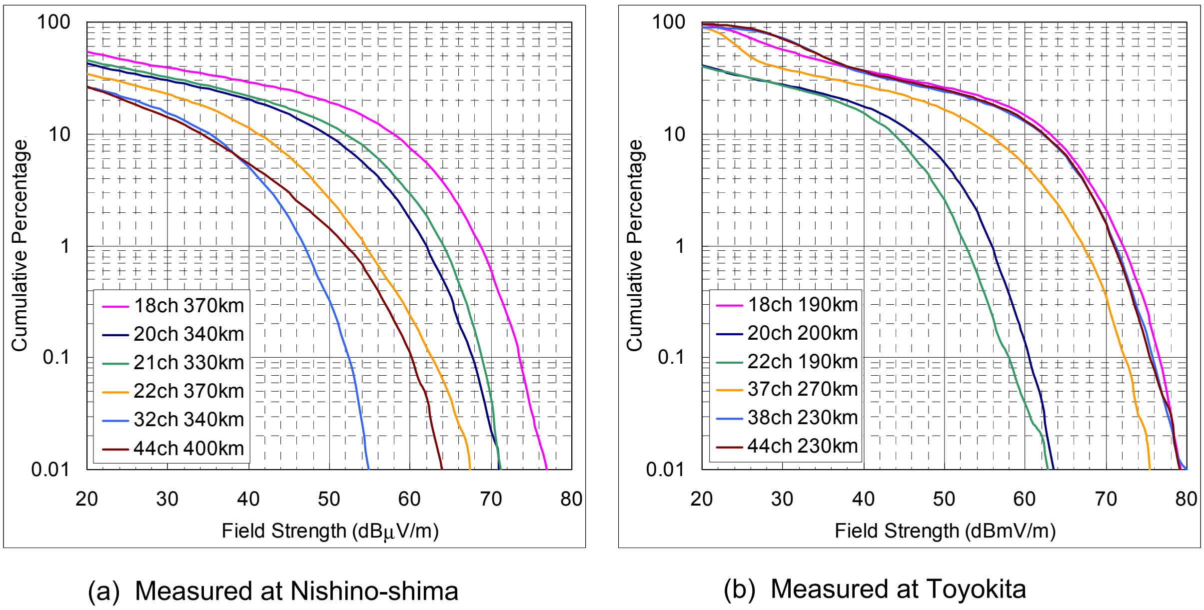


Figure 24 shows the cumulative percentage of field strength measured in different places. Although the field strengths corresponding to a certain percentage of time are different by propagation distances, frequencies (channels), etc., we can see a common feature from the measurement results, as follows.

When the field strengths for 10% of time and 1% of time are to be FS_{10} and FS_1 (dB(µV/m)), respectively, the field strength for 0.1% of time can be obtained by the following equation.

$$FS_{01} = FS_1 + \frac{1}{2}(FS_{10} - FS_1) \qquad \text{dB} \qquad (34)$$

$$FS_{001} = FS_{01} + \frac{1}{2}(FS_1 - FS_{01}) \qquad \text{dB} \qquad (35)$$

where, FS_{01} and FS_{001} represent the field strengths for 0.1% of time and 0.01% of time respectively. The above simple equations are so-called Kumada's law. An example of it is shown in Fig. 25.

FIGURE 25

Example of Kumada's law