

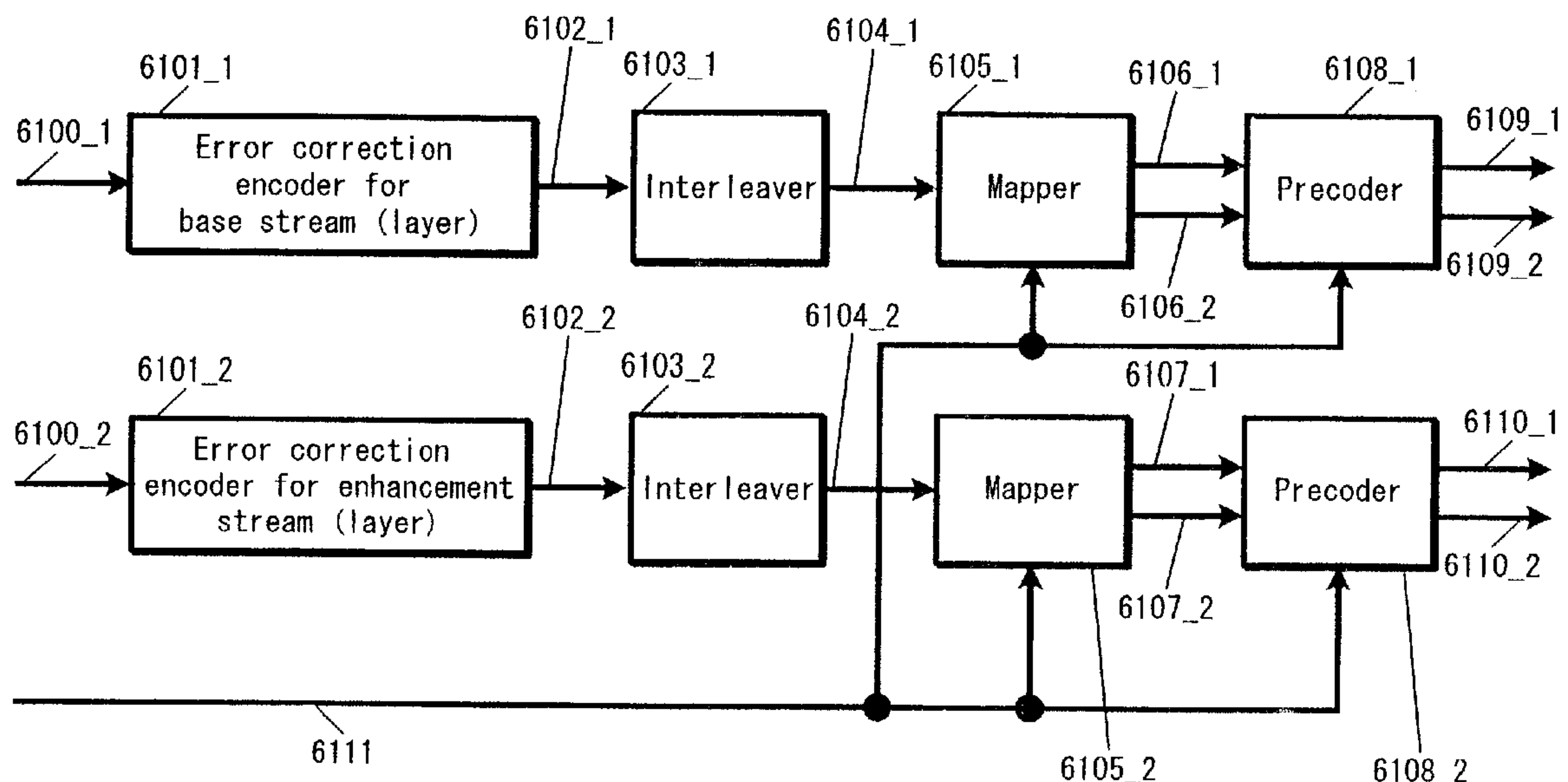


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(54) **Titre : PROCÉDE DE TRANSMISSION, DISPOSITIF DE TRANSMISSION, PROCÉDE DE RÉCEPTION ET DISPOSITIF DE RÉCEPTION**

(54) **Title: TRANSMISSION METHOD, TRANSMISSION DEVICE, RECEPTION METHOD, AND RECEPTION DEVICE**



(57) **Abrégé/Abstract:**

Provided is a precoding method for generating, from a plurality of baseband signals, a plurality of precoded signals to be transmitted over the same frequency bandwidth at the same time, including the steps of selecting a matrix $F[i]$ from among N matrices, which define precoding performed on the plurality of baseband signals, while switching between the N matrices, i being an integer from 0 to $N - 1$. and N being an integer at least two, generating a first precoded signal z_1 and a second precoded signal z_2 , generating a first encoded block and a second encoded block using a predetermined error correction block encoding method, generating a baseband signal with M symbols from the first encoded block and a baseband signal with M symbols the second encoded block, and precoding a combination of the generated baseband signals to generate a precoded signal having M slots.

ABSTRACT

Provided is a precoding method for generating, from a plurality of baseband signals, a plurality of precoded signals to be transmitted over the same frequency bandwidth at the same time, including the steps of selecting a matrix $F[i]$ from
5 among N matrices, which define precoding performed on the plurality of baseband signals, while switching between the N matrices, i being an integer from 0 to $N - 1$, and N being an integer at least two, generating a first precoded signal $z1$ and a second precoded signal $z2$, generating a first encoded block and a second encoded block using a predetermined error correction block encoding method, generating a
10 baseband signal with M symbols from the first encoded block and a baseband signal with M symbols the second encoded block, and precoding a combination of the generated baseband signals to generate a precoded signal having M slots.

DEMANDES OU BREVETS VOLUMINEUX

**LA PRÉSENTE PARTIE DE CETTE DEMANDE OU CE BREVETS
COMPREND PLUS D'UN TOME.**

CECI EST LE TOME __1__ DE __2__

NOTE: Pour les tomes additionels, veuillez contacter le Bureau Canadien des Brevets.

JUMBO APPLICATIONS / PATENTS

**THIS SECTION OF THE APPLICATION / PATENT CONTAINS MORE
THAN ONE VOLUME.**

THIS IS VOLUME __1__ OF __2__

NOTE: For additional volumes please contact the Canadian Patent Office.

DESCRIPTION

[Title of Invention]

[0001]

5 TRANSMISSION METHOD, TRANSMISSION DEVICE, RECEPTION
METHOD, AND RECEPTION DEVICE

[Technical Field]

[0002]

10 The present invention relates to a precoding method, a precoding device, a
transmission method, a transmission device, a reception method, and a reception
device that in particular perform communication using a multi-antenna.

[Background Art]

15 [0003]

Multiple-Input Multiple-Output (MIMO) is a conventional example of a
communication method using a multi-antenna. In multi-antenna communication, of
which MIMO is representative, multiple transmission signals are each modulated,
and each modulated signal is transmitted from a different antenna simultaneously in
20 order to increase the transmission speed of data.

[0004]

Fig. 28 shows an example of the structure of a transmission and reception
device when the number of transmit antennas is two, the number of receive antennas
is two, and the number of modulated signals for transmission (transmission streams)
25 is two. In the transmission device, encoded data is interleaved, the interleaved data is
modulated, and frequency conversion and the like is performed to generate

transmission signals, and the transmission signals are transmitted from antennas. In this case, the method for simultaneously transmitting different modulated signals from different transmit antennas at the same time and at the same frequency is spatial multiplexing MIMO.

5 [0005]

In this context, it has been suggested in Patent Literature 1 to use a transmission device provided with a different interleave pattern for each transmit antenna. In other words, the transmission device in Fig. 28 would have two different interleave patterns with respective interleaves (π_a , π_b). As shown in Non-Patent Literature 1 and Non-Patent Literature 2, reception quality is improved in the reception device by iterative performance of a phase detection method that uses soft values (the MIMO detector in Fig. 28).

[0006]

Models of actual propagation environments in wireless communications include non-line of sight (NLOS), of which a Rayleigh fading environment is representative, and line of sight (LOS), of which a Rician fading environment is representative. When the transmission device transmits a single modulated signal, and the reception device performs maximal ratio combining on the signals received by a plurality of antennas and then demodulates and decodes the signal resulting from maximal ratio combining, excellent reception quality can be achieved in an LOS environment, in particular in an environment where the Rician factor is large, which indicates the ratio of the received power of direct waves versus the received power of scattered waves. However, depending on the transmission system (for example, spatial multiplexing MIMO system), a problem occurs in that the reception quality deteriorates as the Rician factor increases (see Non-Patent Literature 3).

Figs. 29A and 29B show an example of simulation results of the Bit Error Rate (BER) characteristics (vertical axis: BER, horizontal axis: signal-to-noise power ratio (SNR)) for data encoded with low-density parity-check (LDPC) code

and transmitted over a 2×2 (two transmit antennas, two receive antennas) spatial multiplexing MIMO system in a Rayleigh fading environment and in a Rician fading environment with Rician factors of $K = 3, 10$, and 16 dB. Fig. 29A shows the BER characteristics of Max-log A Posteriori Probability (APP) without iterative detection
 5 (see Non-Patent Literature 1 and Non-Patent Literature 2), and Fig. 29B shows the BER characteristics of Max-log-APP with iterative detection (see Non-Patent Literature 1 and Non-Patent Literature 2) (number of iterations: five). As is clear from Figs. 29A and 29B, regardless of whether iterative phase detection is performed, reception quality degrades in the spatial multiplexing MIMO system as
 10 the Rician factor increases. It is thus clear that the unique problem of “degradation of reception quality upon stabilization of the propagation environment in the spatial multiplexing MIMO system”, which does not exist in a conventional single modulation signal transmission system, occurs in the spatial multiplexing MIMO system.

15 [0007]

Broadcast or multicast communication is a service directed towards line-of-sight users. The radio wave propagation environment between the broadcasting station and the reception devices belonging to the users is often an LOS environment. When using a spatial multiplexing MIMO system having the
 20 above problem for broadcast or multicast communication, a situation may occur in which the received electric field strength is high at the reception device, but degradation in reception quality makes it impossible to receive the service. In other words, in order to use a spatial multiplexing MIMO system in broadcast or multicast communication in both an NLOS environment and an LOS environment, there is a
 25 desire for development of a MIMO system that offers a certain degree of reception quality.

[0008]

Non-Patent Literature 8 describes a method to select a codebook used in precoding (i.e. a precoding matrix, also referred to as a precoding weight matrix) based on feedback information from a communication partner. Non-Patent Literature 8 does not at all disclose, however, a method for precoding in an environment in which feedback information cannot be acquired from the communication partner, such as in the above broadcast or multicast communication.

[0009]

On the other hand, Non-Patent Literature 4 discloses a method for switching the precoding matrix over time. This method can be applied even when no feedback information is available. Non-Patent Literature 4 discloses using a unitary matrix as the matrix for precoding and switching the unitary matrix at random but does not at all disclose a method applicable to degradation of reception quality in the above-described LOS environment. Non-Patent Literature 4 simply recites hopping between precoding matrices at random. Obviously, Non-Patent Literature 4 makes no mention whatsoever of a precoding method, or a structure of a precoding matrix, for remedying degradation of reception quality in an LOS environment.

[Citation List]

[Patent Literature]

[0010]

[Patent Literature 1]

WO 2005/050885

[Non-Patent Literature]

[0011]

[Non-Patent Literature 1]

“Achieving near-capacity on a multiple-antenna channel”, IEEE Transaction on Communications, vol. 51, no. 3, pp. 389–399, Mar. 2003.

[Non-Patent Literature 2]

“Performance analysis and design optimization of LDPC-coded MIMO OFDM systems”, IEEE Trans. Signal Processing, vol. 52, no. 2, pp. 348–361, Feb. 2004.

[Non-Patent Literature 3]

- 5 “BER performance evaluation in 2×2 MIMO spatial multiplexing systems under Rician fading channels”, IEICE Trans. Fundamentals, vol. E91-A, no. 10, pp. 2798–2807, Oct. 2008.

[Non-Patent Literature 4]

- “Turbo space-time codes with time varying linear transformations”, IEEE
10 Trans. Wireless communications, vol. 6, no. 2, pp. 486–493, Feb. 2007.

[Non-Patent Literature 5]

“Likelihood function for QR-MLD suitable for soft-decision turbo decoding and its performance”, IEICE Trans. Commun., vol. E88-B, no. 1, pp. 47–57, Jan. 2004.

- 15 [Non-Patent Literature 6]

“A tutorial on ‘parallel concatenated (Turbo) coding’, ‘Turbo (iterative) decoding’ and related topics”, The Institute of Electronics, Information, and Communication Engineers, Technical Report IT 98-51.

[Non-Patent Literature 7]

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[Non-Patent Literature 9]

DVB Document A122, Framing structure, channel coding and modulation for a second generation digital terrestrial television broadcasting system, (DVB-T2), Jun. 2008.

[Non-Patent Literature 10]

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[Non-Patent Literature 11]

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[Non-Patent Literature 12]

R. G. Gallager, “Low-density parity-check codes”, IRE Trans. Inform. Theory, IT-8, pp. 21–28, 1962.

- 15 [Non-Patent Literature 13]

D. J. C. Mackay, “Good error-correcting codes based on very sparse matrices”, IEEE Trans. Inform. Theory, vol. 45, no. 2, pp. 399–431, March 1999.

[Non-Patent Literature 14]

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[Non-Patent Literature 15]

- 25 Y.-L. Ueng, and C.-C. Cheng, “A fast-convergence decoding method and memory-efficient VLSI decoder architecture for irregular LDPC codes in the IEEE 802.16e standards”, IEEE VTC-2007 Fall, pp. 1255–1259.

[Summary of Invention]

[Technical Problem]

[0012]

It is an object of the present invention to provide a MIMO system that improves reception quality in an LOS environment.

[Solution to Problem]

[0013]

5 In order to solve the above problems, an aspect of the present invention is a precoding method for generating, from a plurality of baseband signals, a plurality of precoded signals to be transmitted over the same frequency bandwidth at the same time, comprising the steps of: selecting a matrix $F[i]$ from among N matrices while switching between the N matrices, the N matrices defining precoding
10 performed on the plurality of baseband signals, i being an integer from 0 to $N - 1$, and N being an integer at least two; and generating a first precoded signal $z1$ and a second precoded signal $z2$ by precoding, in accordance with the selected matrix $F[i]$, a first baseband signal $s1$ generated from a first plurality of bits and a second baseband signal $s2$ generated from a second plurality of bits, a first encoded block
15 and a second encoded block being generated respectively as the first plurality of bits and the second plurality of bits using a predetermined error correction block encoding method, the first baseband signal $s1$ and the second baseband signal $s2$ being generated respectively from the first encoded block and the second encoded block to have M symbols each, the first precoded signal $z1$ and the second
20 precoded signal $z2$ being generated to have M slots each by precoding a combination of the first baseband signal $s1$ and the second baseband signal $s2$, M being an integer at least two, the first precoded signal $z1$ and the second precoded signal $z2$ satisfying the equation $(z1, z2)^T = F[i](s1, s2)^T$, $(z1, z2)^T$ being a transposed matrix of $(z1, z2)$, and $(s1, s2)^T$ being a transposed matrix of $(s1, s2)$.

25 [0014]

Another aspect of the present invention is a precoding apparatus for generating, from a plurality of baseband signals, a plurality of precoded signals to be transmitted over the same frequency bandwidth at the same time, comprising: a weighting information generation unit configured to select a matrix $F[i]$ from
30 among N matrices while switching between the N matrices, the N matrices

defining precoding performed on the plurality of baseband signals, i being an integer from 0 to $N - 1$, and N being an integer at least two; a weighting unit configured to generate a first precoded signal $z1$ and a second precoded signal $z2$ by precoding, in accordance with the selected matrix $F[i]$, a first baseband signal $s1$ generated from a first plurality of bits and a second baseband signal $s2$ generated from a second plurality of bits; an error correction coding unit configured to generate a first encoded block as the first plurality of bits and a second encoded block as the second plurality of bits using a predetermined error correction block encoding method; and a mapper configured to generate a baseband signal with M symbols from the first encoded block and a baseband signal with M symbols from the second encoded block, M being an integer at least two, the first precoded signal $z1$ and the second precoded signal $z2$ satisfying the equation $(z1, z2)^T = F[i](s1, s2)^T$, $(z1, z2)^T$ being a transposed matrix of $(z1, z2)$, $(s1, s2)^T$ being a transposed matrix of $(s1, s2)$, and the weighting unit generating precoded signals with M slots by precoding a combination of the baseband signal generated from the first encoded block and the baseband signal generated from the second encoded block.

[0015]

With the above aspects of the present invention, a modulated signal is generated by performing precoding while hopping between precoding matrices so that among a plurality of precoding matrices, a precoding matrix used for at least one data symbol and precoding matrices that are used for data symbols that are adjacent to the data symbol in either the frequency domain or the time domain all differ. Therefore, reception quality in an LOS environment is improved in response to the design of the plurality of precoding matrices.

[Advantageous Effects of Invention]

[0016]

With the above structure, the present invention provides a transmission method, a reception method, a transmission device, and a reception device that

remedy degradation of reception quality in an LOS environment, thereby providing high-quality service to LOS users during broadcast or multicast communication.

[Brief Description of Drawings]

[0017]

5 Fig. 1 is an example of the structure of a transmission device and a reception device in a spatial multiplexing MIMO system.

Fig. 2 is an example of a frame structure.

Fig. 3 is an example of the structure of a transmission device when adopting a method of hopping between precoding weights.

10 Fig. 4 is an example of the structure of a transmission device when adopting a method of hopping between precoding weights.

Fig. 5 is an example of a frame structure.

Fig. 6 is an example of a method of hopping between precoding weights.

Fig. 7 is an example of the structure of a reception device.

15 Fig. 8 is an example of the structure of a signal processing unit in a reception device.

Fig. 9 is an example of the structure of a signal processing unit in a reception device.

Fig. 10 shows a decoding processing method.

20 Fig. 11 is an example of reception conditions.

Figs. 12A and 12B are examples of BER characteristics.

Fig. 13 is an example of the structure of a transmission device when adopting a method of hopping between precoding weights.

25 Fig. 14 is an example of the structure of a transmission device when adopting a method of hopping between precoding weights.

Figs. 15A and 15B are examples of a frame structure.

Figs. 16A and 16B are examples of a frame structure.

Figs. 17A and 17B are examples of a frame structure.

Figs. 18A and 18B are examples of a frame structure.

Figs. 19A and 19B are examples of a frame structure.

Fig. 20 shows positions of poor reception quality points.

Fig. 21 shows positions of poor reception quality points.

5 Fig. 22 is an example of a frame structure.

Fig. 23 is an example of a frame structure.

Figs. 24A and 24B are examples of mapping methods.

Figs. 25A and 25B are examples of mapping methods.

Fig. 26 is an example of the structure of a weighting unit.

10 Fig. 27 is an example of a method for reordering symbols.

Fig. 28 is an example of the structure of a transmission device and a reception device in a spatial multiplexing MIMO system.

Figs. 29A and 29B are examples of BER characteristics.

Fig. 30 is an example of a 2×2 MIMO spatial multiplexing MIMO system.

15 Figs. 31A and 31B show positions of poor reception points.

Fig. 32 shows positions of poor reception points.

Figs. 33A and 33B show positions of poor reception points.

Fig. 34 shows positions of poor reception points.

Figs. 35A and 35B show positions of poor reception points.

20 Fig. 36 shows an example of minimum distance characteristics of poor reception points in an imaginary plane.

Fig. 37 shows an example of minimum distance characteristics of poor reception points in an imaginary plane.

Figs. 38A and 38B show positions of poor reception points.

25 Figs. 39A and 39B show positions of poor reception points.

Fig. 40 is an example of the structure of a transmission device in Embodiment 7.

Fig. 41 is an example of the frame structure of a modulated signal transmitted by the transmission device.

Figs. 42A and 42B show positions of poor reception points.

Figs. 43A and 43B show positions of poor reception points.

5 Figs. 44A and 44B show positions of poor reception points.

Figs. 45A and 45B show positions of poor reception points.

Figs. 46A and 46B show positions of poor reception points.

Figs. 47A and 47B are examples of a frame structure in the time and frequency domains.

10 Figs. 48A and 48B are examples of a frame structure in the time and frequency domains.

Fig. 49 shows a signal processing method.

Fig. 50 shows the structure of modulated signals when using space-time block coding.

15 Fig. 51 is a detailed example of a frame structure in the time and frequency domains.

Fig. 52 is an example of the structure of a transmission device.

Fig. 53 is an example of a structure of the modulated signal generating units #1–#M in Fig. 52.

20 Fig. 54 shows the structure of the OFDM related processors (5207_1 and 5207_2) in Fig. 52.

Figs. 55A and 55B are detailed examples of a frame structure in the time and frequency domains.

Fig. 56 is an example of the structure of a reception device.

25 Fig. 57 shows the structure of the OFDM related processors (5600_X and 5600_Y) in Fig. 56.

Figs. 58A and 58B are detailed examples of a frame structure in the time and frequency domains.

Fig. 59 is an example of a broadcasting system.

Figs. 60A and 60B show positions of poor reception points.

Fig. 61 is an example of the structure of a transmission device when adopting hierarchical transmission.

5 Fig. 62 is an example of the structure of a transmission device when adopting hierarchical transmission.

Fig. 63 is an example of precoding of a base stream.

Fig. 64 is an example of precoding of an enhancement stream.

10 Figs. 65A and 65B are examples of arrangements of symbols in modulated signals when adopting hierarchical transmission.

Fig. 66 is an example of the structure of a signal processing unit in a transmission device when adopting hierarchical transmission.

Fig. 67 is an example of the structure of a transmission device when adopting hierarchical transmission.

15 Fig. 68 is an example of the structure of a transmission device when adopting hierarchical transmission.

Fig. 69 is an example of a structure of symbols in a baseband signal.

Figs. 70A and 70B are examples of arrangements of symbols in modulated signals when adopting hierarchical transmission.

20 Fig. 71 is an example of the structure of a transmission device when adopting hierarchical transmission.

Fig. 72 is an example of the structure of a transmission device when adopting hierarchical transmission.

25 Fig. 73 is an example of a structure of symbols in space-time block coded baseband signals.

Figs. 74A and 74B are examples of arrangements of symbols in modulated signals when adopting hierarchical transmission.

Figs. 75A and 75B are examples of arrangements of symbols in modulated signals when adopting hierarchical transmission.

Fig. 76 is an example of a modification of the number of symbols and of slots necessary for one encoded block when using block coding.

5 Fig. 77 is an example of a modification of the number of symbols and of slots necessary for two encoded blocks when using block coding.

Fig. 78 shows the overall structure of a digital broadcasting system.

Fig. 79 is a block diagram showing an example of the structure of a reception device.

10 Fig. 80 shows the structure of multiplexed data.

Fig. 81 schematically shows how each stream is multiplexed in the multiplexed data.

Fig. 82 shows in detail how a video stream is stored in a sequence of PES packets.

15 Fig. 83 shows the structure of a TS packet and a source packet in multiplexed data.

Fig. 84 shows the data structure of a PMT.

Fig. 85 shows the internal structure of multiplexed data information.

Fig. 86 shows the internal structure of stream attribute information.

20 Fig. 87 is a structural diagram of a video display / audio output device.

Fig. 88 shows the structure of a baseband signal switching unit.

[Description of Embodiments]

[0018]

25 The following describes embodiments of the present invention with reference to the drawings.

(Embodiment 1)

The following describes the transmission method, transmission device, reception method, and reception device of the present embodiment.

[0019]

Prior to describing the present embodiment, an overview is provided of a transmission method and decoding method in a conventional spatial multiplexing MIMO system.

- 5 Fig. 1 shows the structure of an $N_t \times N_r$ spatial multiplexing MIMO system. An information vector z is encoded and interleaved. As output of the interleaving, an encoded bit vector $u = (u_1, \dots, u_{N_t})$ is acquired. Note that $u_i = (u_{i1}, \dots, u_{iM})$ (where M is the number of transmission bits per symbol). Letting the transmission vector $s = (s_1, \dots, s_{N_t})^T$ and the transmission signal from transmit antenna #1 be represented as
- 10 $s_i = \text{map}(u_i)$, the normalized transmission energy is represented as $E\{|s_i|^2\} = E_s/N_t$ (E_s being the total energy per channel). Furthermore, letting the received vector be $y = (y_1, \dots, y_{N_r})^T$, the received vector is represented as in Equation 1.

[0020]

Math 1

- 15 Equation 1

$$\begin{aligned} \mathbf{y} &= \left(y_1, \dots, y_{N_r} \right)^T \\ &= \mathbf{H}_{N_t N_r} \mathbf{s} + \mathbf{n} \end{aligned}$$

[0021]

- In this Equation, $\mathbf{H}_{N_t N_r}$ is the channel matrix, $\mathbf{n} = (n_1, \dots, n_{N_r})^T$ is the noise
- 20 vector, and n_i is the i.i.d. complex Gaussian random noise with an average value 0 and variance σ^2 . From the relationship between transmission symbols and reception symbols that is induced at the reception device, the probability for the received vector may be provided as a multi-dimensional Gaussian distribution, as in Equation
2. 25

[0022]

Math 2

Equation 2

$$p(\mathbf{y} | \mathbf{u}) = \frac{1}{(2\pi\sigma^2)^{N_r}} \exp\left(-\frac{1}{2\sigma^2} \|\mathbf{y} - \mathbf{H}\mathbf{s}(\mathbf{u})\|^2\right)$$

5 [0023]

Here, a reception device that performs iterative decoding composed of an outer soft-in/soft-out decoder and a MIMO detector, as in Fig. 1, is considered. The vector of a log-likelihood ratio (L-value) in Fig. 1 is represented as in Equations 3–5.

10 [0024]

Math 3

Equation 3

$$L(\mathbf{u}) = \left(L(u_1), \dots, L(u_{N_t})\right)^T$$

15 [0025]

Math 4

Equation 4

$$L(\mathbf{u}_i) = \left(L(u_{i1}), \dots, L(u_{iM})\right)$$

20 [0026]

Math 5

Equation 5

$$L(u_{ij}) = \ln \frac{P(u_{ij} = +1)}{P(u_{ij} = -1)}$$

[0027]

<Iterative Detection Method>

5 The following describes iterative detection of MIMO signals in the $N_t \times N_r$ spatial multiplexing MIMO system.

The log-likelihood ratio of u_{mn} is defined as in Equation 6.

[0028]

Math 6

10 Equation 6

$$L(\mathbf{u}_{mn} | \mathbf{y}) = \ln \frac{P(\mathbf{u}_{mn} = +1 | \mathbf{y})}{P(\mathbf{u}_{mn} = -1 | \mathbf{y})}$$

[0029]

From Bayes' theorem, Equation 6 can be expressed as Equation 7.

15 [0030]

Math 7

Equation 7

$$\begin{aligned}
L(\mathbf{u}_{mn} | \mathbf{y}) &= \ln \frac{p(\mathbf{y} | \mathbf{u}_{mn} = +1)P(\mathbf{u}_{mn} = +1)/p(\mathbf{y})}{p(\mathbf{y} | \mathbf{u}_{mn} = -1)P(\mathbf{u}_{mn} = -1)/p(\mathbf{y})} \\
&= \ln \frac{P(\mathbf{u}_{mn} = +1)}{P(\mathbf{u}_{mn} = -1)} + \ln \frac{p(\mathbf{y} | \mathbf{u}_{mn} = +1)}{p(\mathbf{y} | \mathbf{u}_{mn} = -1)} \\
&= \ln \frac{P(\mathbf{u}_{mn} = +1)}{P(\mathbf{u}_{mn} = -1)} + \ln \frac{\sum_{U_{mn,+1}} p(\mathbf{y} | \mathbf{u})p(\mathbf{u} | \mathbf{u}_{mn})}{\sum_{U_{mn,-1}} p(\mathbf{y} | \mathbf{u})p(\mathbf{u} | \mathbf{u}_{mn})}
\end{aligned}$$

[0031]

Let $U_{mn,\pm 1} = \{\mathbf{u} | \mathbf{u}_{mn} = \pm 1\}$. When approximating $\ln \Sigma a_j \sim \max \ln a_j$, an
5 approximation of Equation 7 can be sought as Equation 8. Note that the above
symbol “ $\sim$ ” indicates approximation.

[0032]

Math 8

Equation 8

$$\begin{aligned}
L(\mathbf{u}_{mn} | \mathbf{y}) &\approx \ln \frac{P(\mathbf{u}_{mn} = +1)}{P(\mathbf{u}_{mn} = -1)} + \max_{U_{mn,+1}} \{ \ln p(\mathbf{y} | \mathbf{u}) + P(\mathbf{u} | \mathbf{u}_{mn}) \} \\
10 \quad &\quad - \max_{U_{mn,-1}} \{ \ln p(\mathbf{y} | \mathbf{u}) + P(\mathbf{u} | \mathbf{u}_{mn}) \}
\end{aligned}$$

[0033]

$P(\mathbf{u} | \mathbf{u}_{mn})$ and $\ln P(\mathbf{u} | \mathbf{u}_{mn})$ in Equation 8 are represented as follows.

[0034]

15 Math 9

Equation 9

$$\begin{aligned}
P(\mathbf{u} | \mathcal{U}_{mn}) &= \prod_{(ij) \neq (mn)} P(\mathcal{U}_{ij}) \\
&= \prod_{(ij) \neq (mn)} \frac{\exp\left(\frac{\mathcal{U}_{ij} L(\mathcal{U}_{ij})}{2}\right)}{\exp\left(\frac{L(\mathcal{U}_{ij})}{2}\right) + \exp\left(-\frac{L(\mathcal{U}_{ij})}{2}\right)}
\end{aligned}$$

[0035]

Math 10

5 Equation 10

$$\ln P(\mathbf{u} | \mathcal{U}_{mn}) = \left(\sum_{ij} \ln P(\mathcal{U}_{ij}) \right) - \ln P(\mathcal{U}_{mn})$$

[0036]

Math 11

10 Equation 11

$$\begin{aligned}
\ln P(u_{ij}) &= \frac{1}{2} u_{ij} P(u_{ij}) - \ln \left(\exp\left(\frac{L(u_{ij})}{2}\right) + \exp\left(-\frac{L(u_{ij})}{2}\right) \right) \\
&\approx \frac{1}{2} u_{ij} L(u_{ij}) - \frac{1}{2} |L(u_{ij})| \quad \text{for } |L(u_{ij})| > 2 \\
&= \left| \frac{L(u_{ij})}{2} \right| (u_{ij} \operatorname{sign}(L(u_{ij})) - 1)
\end{aligned}$$

[0037]

Incidentally, the logarithmic probability of the equation defined in Equation 2 is represented in Equation 12.

[0038]

Math 12

5 Equation 12

$$\ln P(\mathbf{y} | \mathbf{u}) = -\frac{N_r}{2} \ln(2\pi \sigma^2) - \frac{1}{2\sigma^2} \|\mathbf{y} - \mathbf{H}\mathbf{s}(\mathbf{u})\|^2$$

[0039]

Accordingly, from Equations 7 and 13, in MAP or A Posteriori Probability (APP), the a posteriori L-value is represented as follows.

[0040]

Math 13

Equation 13

$$L(\mathbf{u}_{mn} | \mathbf{y}) = \ln \frac{\sum_{U_{mn,+1}} \exp \left\{ -\frac{1}{2\sigma^2} \|\mathbf{y} - \mathbf{H}\mathbf{s}(\mathbf{u})\|^2 + \sum_{ij} \ln P(\mathbf{u}_{ij}) \right\}}{\sum_{U_{mn,-1}} \exp \left\{ -\frac{1}{2\sigma^2} \|\mathbf{y} - \mathbf{H}\mathbf{s}(\mathbf{u})\|^2 + \sum_{ij} \ln P(\mathbf{u}_{ij}) \right\}}$$

15

[0041]

Hereinafter, this is referred to as iterative APP decoding. From Equations 8 and 12, in the log-likelihood ratio utilizing Max-Log approximation (Max-Log APP), the a posteriori L-value is represented as follows.

20 [0042]

Math 14

Equation 14

$$L(\mathbf{u}_{mn} | \mathbf{y}) \approx \max_{U_{mn,+1}} \{\Psi(\mathbf{u}, \mathbf{y}, L(\mathbf{u}))\} - \max_{U_{mn,-1}} \{\Psi(\mathbf{u}, \mathbf{y}, L(\mathbf{u}))\}$$

[0043]

Math 15

5 Equation 15

$$\Psi(\mathbf{u}, \mathbf{y}, L(\mathbf{u})) = -\frac{1}{2\sigma^2} \|\mathbf{y} - \mathbf{H}\mathbf{s}(\mathbf{u})\|^2 + \sum_{ij} \ln P(\mathbf{u}_{ij})$$

[0044]

Hereinafter, this is referred to as iterative Max-log APP decoding. The
 10 extrinsic information required in an iterative decoding system can be sought by
 subtracting prior inputs from Equations 13 and 14.

<System Model>

Fig. 28 shows the basic structure of the system that is related to the
 subsequent description. This system is a 2×2 spatial multiplexing MIMO system.
 15 There is an outer encoder for each of streams A and B. The two outer encoders are
 identical LDPC encoders. (Here, a structure using LDPC encoders as the outer
 encoders is described as an example, but the error correction coding used by the
 outer encoder is not limited to LDPC coding. The present invention may similarly be
 embodied using other error correction coding such as turbo coding, convolutional
 20 coding, LDPC convolutional coding, and the like. Furthermore, each outer encoder
 is described as having a transmit antenna, but the outer encoders are not limited to
 this structure. A plurality of transmit antennas may be used, and the number of outer
 encoders may be one. Also, a greater number of outer encoders may be used than the
 number of transmit antennas.) The streams A and B respectively have interleavers

(π_a, π_b) . Here, the modulation scheme is 2^h -QAM (with h bits transmitted in one symbol).

[0045]

The reception device performs iterative detection on the above MIMO
5 signals (iterative APP (or iterative Max-log APP) decoding). Decoding of LDPC codes is performed by, for example, sum-product decoding.

[0046]

Fig. 2 shows a frame structure and lists the order of symbols after interleaving. In this case, (i_a, j_a) , (i_b, j_b) are represented by the following Equations.

10 [0047]

Math 16

Equation 16

$$(i_a, j_a) = \pi_a(\Omega_{ia,ja}^a)$$

15 [0048]

Math 17

Equation 17

$$(i_b, j_b) = \pi_b(\Omega_{ib,jb}^a)$$

20 [0049]

In this case, i^a, i^b indicate the order of symbols after interleaving, j^a, j^b indicate the bit positions ($j^a, j^b = 1, \dots, h$) in the modulation scheme, π^a, π^b indicate the interleavers for the streams A and B, and $\Omega_{ia,ja}^a, \Omega_{ib,jb}^b$ indicate the order of data in streams A and B before interleaving. Note that Fig. 2 shows the frame structure
25 for $i_a = i_b$.

<Iterative Decoding>

The following is a detailed description of the algorithms for sum-product decoding used in decoding of LDPC codes and for iterative detection of MIMO signals in the reception device.

[0050]

5 Sum-Product Decoding

Let a two-dimensional $M \times N$ matrix $H = \{H_{mn}\}$ be the check matrix for LDPC codes that are targeted for decoding. Subsets $A(m)$, $B(n)$ of the set $[1, N] = \{1, 2, \dots, N\}$ are defined by the following Equations.

[0051]

10 Math 18

Equation 18

$$A(m) \equiv \{n : H_{mn} = 1\}$$

[0052]

15 Math 19

Equation 19

$$B(n) \equiv \{m : H_{mn} = 1\}$$

[0053]

20 In these Equations, $A(m)$ represents the set of column indices of 1's in the m^{th} column of the check matrix H , and $B(n)$ represents the set of row indices of 1's in the n^{th} row of the check matrix H . The algorithm for sum-product decoding is as follows.

Step A.1 (initialization): let a priori value logarithmic ratio $\beta_{mn} = 0$ for all
 25 combinations (m, n) satisfying $H_{mn} = 1$. Assume that the loop variable (the number of iterations) $l_{\text{sum}} = 1$ and the maximum number of loops is set to $l_{\text{sum, max}}$.

Step A.2 (row processing): the extrinsic value logarithmic ratio α_{mn} is updated for all combinations (m, n) satisfying $H_{mn} = 1$ in the order of $m = 1, 2, \dots, M$, using the following updating Equations.

[0054]

5 Math 20

Equation 20

$$\alpha_{mn} = \left(\prod_{n' \in A(m) \setminus n} \text{sign}(\lambda_{n'} + \beta_{mn'}) \right) \times f \left(\sum_{n' \in A(m) \setminus n} f(\lambda_{n'} + \beta_{mn'}) \right)$$

[0055]

10 Math 21

Equation 21

$$\text{sign}(x) \equiv \begin{cases} 1 & x \geq 0 \\ -1 & x < 0 \end{cases}$$

[0056]

15 Math 22

Equation 22

$$f(x) \equiv \ln \frac{\exp(x) + 1}{\exp(x) - 1}$$

[0057]

20 In these Equations, f represents a Gallager function. Furthermore, the method of seeking λ_n is described in detail later.

Step A·3 (column processing): the extrinsic value logarithmic ratio β_{mn} is updated for all combinations (m, n) satisfying $H_{mn} = 1$ in the order of $n = 1, 2, \dots, N$, using the following updating Equation.

[0058]

5 Math 23

Equation 23

$$\beta_{mn} = \sum_{m' \in B(n) \setminus m} \alpha_{m'n}$$

[0059]

10 Step A·4 (calculating a log-likelihood ratio): the log-likelihood ratio L_n is sought for $n \in [1, N]$ by the following Equation.

[0060]

Math 24

Equation 24

15
$$L_n = \sum_{m' \in B(n) \setminus m} \alpha_{m'n} + \lambda_n$$

[0061]

Step A·5 (count of the number of iterations): if $l_{\text{sum}} < l_{\text{sum, max}}$, then l_{sum} is incremented, and processing returns to step A·2. If $l_{\text{sum}} = l_{\text{sum, max}}$, the sum-product
20 decoding in this round is finished.

[0062]

The operations in one sum-product decoding have been described. Subsequently, iterative MIMO signal detection is performed. In the variables $m, n, \alpha_{mn}, \beta_{mn}, \lambda_n$, and L_n , used in the above description of the operations of sum-product

decoding, the variables in stream A are m_a , n_a , α_{mana}^a , β_{mana}^a , λ_{na} , and L_{na} , and the variables in stream B are $m_b, n_b, \alpha_{mbnb}^b, \beta_{mbnb}^b, \lambda_{nb}$, and L_{nb} .

<Iterative MIMO Signal Detection>

The following describes the method of seeking λ_n in iterative MIMO signal
5 detection in detail.

[0063]

The following Equation holds from Equation 1.

[0064]

Math 25

10 Equation 25

$$\begin{aligned} \mathbf{y}(t) &= \left(y_1(t), y_2(t) \right)^T \\ &= \mathbf{H}_{22}(t) \mathbf{s}(t) + \mathbf{n}(t) \end{aligned}$$

[0065]

The following Equations are defined from the frame structures of Fig. 2 and
15 from Equations 16 and 17.

[0066]

Math 26

Equation 26

$$n_a = \Omega_{ia,ja}^a$$

20

[0067]

Math 27

Equation 27

$$n_b = \Omega_{ib,jb}^b$$

[0068]

In this case, $n_a, n_b \in [1, N]$. Hereinafter, λ_{na} , L_{na} , λ_{nb} , and L_{nb} , where the
 5 number of iterations of iterative MIMO signal detection is k , are represented as $\lambda_{k, na}$,
 $L_{k, na}$, $\lambda_{k, nb}$, and $L_{k, nb}$.

[0069]

Step B.1 (initial detection; $k = 0$): $\lambda_{0, na}$ and $\lambda_{0, nb}$ are sought as follows in the
 case of initial detection.

10 In iterative APP decoding:

[0070]

Math 28

Equation 28

$$\lambda_{0, n_X} = \ln \frac{\sum_{U_{0, n_X, +1}} \exp \left\{ -\frac{1}{2\sigma^2} \left\| \mathbf{y}(i_X) - \mathbf{H}_{22}(i_X) \mathbf{s}(\mathbf{u}(i_X)) \right\|^2 \right\}}{\sum_{U_{0, n_X, -1}} \exp \left\{ -\frac{1}{2\sigma^2} \left\| \mathbf{y}(i_X) - \mathbf{H}_{22}(i_X) \mathbf{s}(\mathbf{u}(i_X)) \right\|^2 \right\}}$$

15

[0071]

In iterative Max-log APP decoding:

[0072]

Math 29

20 Equation 29

$$\lambda_{0, n_X} = \max_{U_{0, n_X, +1}} \left\{ \Psi(\mathbf{u}(i_X), \mathbf{y}(i_X)) \right\} - \max_{U_{0, n_X, -1}} \left\{ \Psi(\mathbf{u}(i_X), \mathbf{y}(i_X)) \right\}$$

[0073]

Math 30

Equation 30

$$\Psi(\mathbf{u}(i_X), \mathbf{y}(i_X)) = -\frac{1}{2\sigma^2} \left\| \mathbf{y}(i_X) - \mathbf{H}_{22}(i_X) \mathbf{s}(\mathbf{u}(i_X)) \right\|^2$$

[0074]

Here, let $X = a, b$. Then, assume that the number of iterations of iterative MIMO signal detection is $l_{\text{mimo}} = 0$ and the maximum number of iterations is set to

10 $l_{\text{mimo, max}}$.

[0075]

Step B.2 (iterative detection; the number of iterations k): $\lambda_{k, na}$ and $\lambda_{k, nb}$, where the number of iterations is k , are represented as in Equations 31–34, from Equations 11, 13–15, 16, and 17. Let $(X, Y) = (a, b)(b, a)$.

15 In iterative APP decoding:

[0076]

Math 31

Equation 31

$$\lambda_{k, r_X} = L_{k-1, \Omega_{X,Y}^X}(\mathbf{u}_{\Omega_{X,Y}^X}) + \ln \frac{\sum_{U_{k, n_X, +1}} \exp \left\{ -\frac{1}{2\sigma^2} \left\| \mathbf{y}(i_X) - \mathbf{H}_{22}(i_X) \mathbf{s}(\mathbf{u}(i_X)) \right\|^2 + \rho(\mathbf{u}_{\Omega_{X,Y}^X}) \right\}}{\sum_{U_{k, n_X, -1}} \exp \left\{ -\frac{1}{2\sigma^2} \left\| \mathbf{y}(i_X) - \mathbf{H}_{22}(i_X) \mathbf{s}(\mathbf{u}(i_X)) \right\|^2 - \rho(\mathbf{u}_{\Omega_{X,Y}^X}) \right\}}$$

20

[0077]

Math 32

Equation 32

$$\rho(u_{\Omega_{iX,jX}^X}) = \sum_{\substack{\gamma=1 \\ \gamma \neq jX}}^h \left| \frac{L_{k-1, \Omega_{iX,\gamma}^X}(u_{\Omega_{iX,\gamma}^X})}{2} \right| \left(u_{\Omega_{iX,\gamma}^X} \text{sign}(L_{k-1, \Omega_{iX,\gamma}^X}(u_{\Omega_{iX,\gamma}^X})) - 1 \right) \\ + \sum_{\gamma=1}^h \left| \frac{L_{k-1, \Omega_{iX,\gamma}^Y}(u_{\Omega_{iX,\gamma}^Y})}{2} \right| \left(u_{\Omega_{iX,\gamma}^Y} \text{sign}(L_{k-1, \Omega_{iX,\gamma}^Y}(u_{\Omega_{iX,\gamma}^Y})) - 1 \right)$$

[0078]

In iterative Max-log APP decoding:

5 [0079]

Math 33

Equation 33

$$\lambda_{k,n_X} = L_{k-1, \Omega_{iX,jX}^X}(u_{\Omega_{iX,jX}^X}) + \max_{U_{k,n_X}+1} \left\{ \Psi(\mathbf{u}(i_X), \mathbf{y}(i_X), \rho(u_{\Omega_{iX,jX}^X})) \right\} - \max_{U_{k,n_X}-1} \left\{ \Psi(\mathbf{u}(i_X), \mathbf{y}(i_X), \rho(u_{\Omega_{iX,jX}^X})) \right\}$$

10 [0080]

Math 34

Equation 34

$$\Psi(\mathbf{u}(i_X), \mathbf{y}(i_X), \rho(u_{\Omega_{iX,jX}^X})) = -\frac{1}{2\sigma^2} \|\mathbf{y}(i_X) - \mathbf{H}_{22}(i_X)\mathbf{s}(\mathbf{u}(i_X))\|^2 + \rho(u_{\Omega_{iX,jX}^X})$$

15 [0081]

Step B.3 (counting the number of iterations and estimating a codeword):

increment l_{mimo} if $l_{\text{mimo}} < l_{\text{mimo, max}}$, and return to step B.2. Assuming that $l_{\text{mimo}} = l_{\text{mimo, max}}$, the estimated codeword is sought as in the following Equation.

[0082]

20 Math 35

Equation 35

$$\hat{u}_{n_X} = \begin{cases} 1 & L_{l_{mimo}, n_X} \geq 0 \\ -1 & L_{l_{mimo}, n_X} < 0 \end{cases}$$

[0083]

Here, let X = a, b.

5 [0084]

Fig. 3 is an example of the structure of a transmission device 300 in the present embodiment. An encoder 302A receives information (data) 301A and a frame structure signal 313 as inputs and, in accordance with the frame structure signal 313, performs error correction coding such as convolutional coding, LDPC coding, turbo coding, or the like, outputting encoded data 303A. (The frame structure signal 313 includes information such as the error correction method used for error correction coding of data, the encoding ratio, the block length, and the like. The encoder 302A uses the error correction method indicated by the frame structure signal 313. Furthermore, the error correction method may be switched.)

15 [0085]

An interleaver 304A receives the encoded data 303A and the frame structure signal 313 as inputs and performs interleaving, i.e. changing the order of the data, to output interleaved data 305A. (The method of interleaving may be switched based on the frame structure signal 313.)

20 A mapper 306A receives the interleaved data 305A and the frame structure signal 313 as inputs, performs modulation such as Quadrature Phase Shift Keying (QPSK), 16 Quadrature Amplitude Modulation (16QAM), 64 Quadrature Amplitude Modulation (64QAM), or the like, and outputs a resulting baseband signal 307A. (The method of modulation may be switched based on the frame structure signal 313.)

25

Figs. 24A and 24B are an example of a mapping method over an IQ plane, having an in-phase component I and a quadrature component Q, to form a baseband signal in QPSK modulation. For example, as shown in Fig. 24A, if the input data is “00”, the output is $I = 1.0$, $Q = 1.0$. Similarly, for input data of “01”, the output is $I = -1.0$, $Q = 1.0$, and so forth. Fig. 24B is an example of a different method of mapping in an IQ plane for QPSK modulation than Fig. 24A. The difference between Fig. 24B and Fig. 24A is that the signal points in Fig. 24A have been rotated around the origin to yield the signal points of Fig. 24B. Non-Patent Literature 9 and Non-Patent Literature 10 describe such a constellation rotation method, and the Cyclic Q Delay described in Non-Patent Literature 9 and Non-Patent Literature 10 may also be adopted. As another example apart from Figs. 24A and 24B, Figs. 25A and 25B show signal point layout in the IQ plane for 16QAM. The example corresponding to Fig. 24A is shown in Fig. 25A, and the example corresponding to Fig. 24B is shown in Fig. 25B.

15 [0086]

An encoder 302B receives information (data) 301B and the frame structure signal 313 as inputs and, in accordance with the frame structure signal 313, performs error correction coding such as convolutional coding, LDPC coding, turbo coding, or the like, outputting encoded data 303B. (The frame structure signal 313 includes information such as the error correction method used, the encoding ratio, the block length, and the like. The error correction method indicated by the frame structure signal 313 is used. Furthermore, the error correction method may be switched.)

An interleaver 304B receives the encoded data 303B and the frame structure signal 313 as inputs and performs interleaving, i.e. changing the order of the data, to output interleaved data 305B. (The method of interleaving may be switched based on the frame structure signal 313.)

A mapper 306B receives the interleaved data 305B and the frame structure signal 313 as inputs, performs modulation such as Quadrature Phase Shift Keying

(QPSK), 16 Quadrature Amplitude Modulation (16QAM), 64 Quadrature Amplitude Modulation (64QAM), or the like, and outputs a resulting baseband signal 307B. (The method of modulation may be switched based on the frame structure signal 313.)

5 [0087]

A weighting information generating unit 314 receives the frame structure signal 313 as an input and outputs information 315 regarding a weighting method based on the frame structure signal 313. The weighting method is characterized by regular hopping between weights.

10 [0088]

A weighting unit 308A receives the baseband signal 307A, the baseband signal 307B, and the information 315 regarding the weighting method, and based on the information 315 regarding the weighting method, performs weighting on the baseband signal 307A and the baseband signal 307B and outputs a signal 309A resulting from the weighting. Details on the weighting method are provided later.

[0089]

A wireless unit 310A receives the signal 309A resulting from the weighting as an input and performs processing such as orthogonal modulation, band limiting, frequency conversion, amplification, and the like, outputting a transmission signal 311A. A transmission signal 311A is output as a radio wave from an antenna 312A.

[0090]

A weighting unit 308B receives the baseband signal 307A, the baseband signal 307B, and the information 315 regarding the weighting method, and based on the information 315 regarding the weighting method, performs weighting on the baseband signal 307A and the baseband signal 307B and outputs a signal 309B resulting from the weighting.

[0091]

Fig. 26 shows the structure of a weighting unit. The baseband signal 307A is multiplied by $w_{11}(t)$, yielding $w_{11}(t)s_1(t)$, and is multiplied by $w_{21}(t)$, yielding $w_{21}(t)s_1(t)$. Similarly, the baseband signal 307B is multiplied by $w_{12}(t)$ to generate $w_{12}(t)s_2(t)$ and is multiplied by $w_{22}(t)$ to generate $w_{22}(t)s_2(t)$. Next, $z_1(t) =$
 5 $w_{11}(t)s_1(t) + w_{12}(t)s_2(t)$ and $z_2(t) = w_{21}(t)s_1(t) + w_{22}(t)s_2(t)$ are obtained.

[0092]

Details on the weighting method are provided later.

[0093]

A wireless unit 310B receives the signal 309B resulting from the weighting
 10 as an input and performs processing such as orthogonal modulation, band limiting, frequency conversion, amplification, and the like, outputting a transmission signal 311B. A transmission signal 311B is output as a radio wave from an antenna 312B.

[0094]

Fig. 4 shows an example of the structure of a transmission device 400 that
 15 differs from Fig. 3. The differences in Fig. 4 from Fig. 3 are described.

[0095]

An encoder 402 receives information (data) 401 and the frame structure signal 313 as inputs and, in accordance with the frame structure signal 313, performs error correction coding and outputs encoded data 402.

20 [0096]

A distribution unit 404 receives the encoded data 403 as an input, distributes the data 403, and outputs data 405A and data 405B. Note that in Fig. 4, one encoder is shown, but the number of encoders is not limited in this way. The present invention may similarly be embodied when the number of encoders is m (where m is
 25 an integer greater than or equal to one) and the distribution unit divides encoded data generated by each encoder into two parts and outputs the divided data.

[0097]

Fig. 5 shows an example of a frame structure in the time domain for a transmission device according to the present embodiment. A symbol 500_1 is a symbol for notifying the reception device of the transmission method. For example, the symbol 500_1 conveys information such as the error correction method used for transmitting data symbols, the encoding ratio, and the modulation method used for transmitting data symbols.

[0098]

The symbol 501_1 is for estimating channel fluctuation for the modulated signal $z1(t)$ (where t is time) transmitted by the transmission device. The symbol 502_1 is the data symbol transmitted as symbol number u (in the time domain) by the modulated signal $z1(t)$, and the symbol 503_1 is the data symbol transmitted as symbol number $u + 1$ by the modulated signal $z1(t)$.

[0099]

The symbol 501_2 is for estimating channel fluctuation for the modulated signal $z2(t)$ (where t is time) transmitted by the transmission device. The symbol 502_2 is the data symbol transmitted as symbol number u by the modulated signal $z2(t)$, and the symbol 503_2 is the data symbol transmitted as symbol number $u + 1$ by the modulated signal $z2(t)$.

[0100]

The following describes the relationships between the modulated signals $z1(t)$ and $z2(t)$ transmitted by the transmission device and the received signals $r1(t)$ and $r2(t)$ received by the reception device.

[0101]

In Fig. 5, 504#1 and 504#2 indicate transmit antennas in the transmission device, and 505#1 and 505#2 indicate receive antennas in the reception device. The transmission device transmits the modulated signal $z1(t)$ from transmit antenna 504#1 and transmits the modulated signal $z2(t)$ from transmit antenna 504#2. In this case, the modulated signal $z1(t)$ and the modulated signal $z2(t)$ are assumed to

occupy the same (a shared/common) frequency (bandwidth). Letting the channel fluctuation for the transmit antennas of the transmission device and the antennas of the reception device be $h_{11}(t)$, $h_{12}(t)$, $h_{21}(t)$, and $h_{22}(t)$, the signal received by the receive antenna 505#1 of the reception device be $r1(t)$, and the signal received by the receive antenna 505#2 of the reception device be $r2(t)$, the following relationship holds.

[0102]

Math 36

Equation 36

$$\begin{pmatrix} r1(t) \\ r2(t) \end{pmatrix} = \begin{pmatrix} h_{11}(t) & h_{12}(t) \\ h_{21}(t) & h_{22}(t) \end{pmatrix} \begin{pmatrix} z1(t) \\ z2(t) \end{pmatrix}$$

[0103]

Fig. 6 relates to the weighting method (precoding method) in the present embodiment. A weighting unit 600 integrates the weighting units 308A and 308B in Fig. 3. As shown in Fig. 6, a stream $s1(t)$ and a stream $s2(t)$ correspond to the baseband signals 307A and 307B in Fig. 3. In other words, the streams $s1(t)$ and $s2(t)$ are the baseband signal in-phase components I and quadrature components Q when mapped according to a modulation scheme such as QPSK, 16QAM, 64QAM, or the like. As indicated by the frame structure of Fig. 6, the stream $s1(t)$ is represented as $s1(u)$ at symbol number u , as $s1(u + 1)$ at symbol number $u + 1$, and so forth. Similarly, the stream $s2(t)$ is represented as $s2(u)$ at symbol number u , as $s2(u + 1)$ at symbol number $u + 1$, and so forth. The weighting unit 600 receives the baseband signals 307A ($s1(t)$) and 307B ($s2(t)$) and the information 315 regarding weighting information in Fig. 3 as inputs, performs weighting in accordance with the information 315 regarding weighting, and outputs the signals 309A ($z1(t)$) and 309B

($z_2(t)$) after weighting in Fig. 3. In this case, $z_1(t)$ and $z_2(t)$ are represented as follows.

For symbol number $4i$ (where i is an integer greater than or equal to zero):

[0104]

5 Math 37

Equation 37

$$\begin{pmatrix} z_1(4i) \\ z_2(4i) \end{pmatrix} = \frac{1}{\sqrt{2}} \begin{pmatrix} e^{j0} & e^{j0} \\ e^{j0} & e^{j\frac{3}{4}\pi} \end{pmatrix} \begin{pmatrix} s_1(4i) \\ s_2(4i) \end{pmatrix}$$

[0105]

10 Here, j is an imaginary unit.

For symbol number $4i + 1$:

[0106]

Math 38

Equation 38

$$15 \quad \begin{pmatrix} z_1(4i + 1) \\ z_2(4i + 1) \end{pmatrix} = \frac{1}{\sqrt{2}} \begin{pmatrix} e^{j0} & e^{j0} \\ e^{j\frac{3}{4}\pi} & e^{j0} \end{pmatrix} \begin{pmatrix} s_1(4i + 1) \\ s_2(4i + 1) \end{pmatrix}$$

[0107]

For symbol number $4i + 2$:

[0108]

20 Math 39

Equation 39

$$\begin{pmatrix} z1(4i+2) \\ z2(4i+2) \end{pmatrix} = \frac{1}{\sqrt{2}} \begin{pmatrix} e^{j0} & e^{j\frac{3}{4}\pi} \\ e^{j0} & e^{j0} \end{pmatrix} \begin{pmatrix} s1(4i+2) \\ s2(4i+2) \end{pmatrix}$$

[0109]

For symbol number $4i+3$:

5 [0110]

Math 40

Equation 40

$$\begin{pmatrix} z1(4i+3) \\ z2(4i+3) \end{pmatrix} = \frac{1}{\sqrt{2}} \begin{pmatrix} e^{j\frac{3}{4}\pi} & e^{j0} \\ e^{j0} & e^{j0} \end{pmatrix} \begin{pmatrix} s1(4i+3) \\ s2(4i+3) \end{pmatrix}$$

10 [0111]

In this way, the weighting unit in Fig. 6 regularly hops between precoding weights over a four-slot period (cycle). (While precoding weights have been described as being hopped between regularly over four slots, the number of slots for regular hopping is not limited to four.)

15 Incidentally, Non-Patent Literature 4 describes switching the precoding weights for each slot. This switching of precoding weights is characterized by being random. On the other hand, in the present embodiment, a certain period (cycle) is provided, and the precoding weights are hopped between regularly. Furthermore, in each 2×2 precoding weight matrix composed of four precoding weights, the
20 absolute value of each of the four precoding weights is equivalent to $(1/\sqrt{2})$, and hopping is regularly performed between precoding weight matrices having this characteristic.

[0112]

In an LOS environment, if a special precoding matrix is used, reception quality may greatly improve, yet the special precoding matrix differs depending on the conditions of direct waves. In an LOS environment, however, a certain tendency
 5 exists, and if precoding matrices are hopped between regularly in accordance with this tendency, the reception quality of data greatly improves. On the other hand, when precoding matrices are hopped between at random, a precoding matrix other than the above-described special precoding matrix may exist, and the possibility of performing precoding only with biased precoding matrices that are not suitable for
 10 the LOS environment also exists. Therefore, in an LOS environment, excellent reception quality may not always be obtained. Accordingly, there is a need for a precoding hopping method suitable for an LOS environment. The present invention proposes such a precoding method.

[0113]

15 Fig. 7 is an example of the structure of a reception device 700 in the present embodiment. A wireless unit 703_X receives, as an input, a received signal 702_X received by an antenna 701_X, performs processing such as frequency conversion, quadrature demodulation, and the like, and outputs a baseband signal 704_X.

[0114]

20 A channel fluctuation estimating unit 705_1 for the modulated signal z_1 transmitted by the transmission device receives the baseband signal 704_X as an input, extracts a reference symbol 501_1 for channel estimation as in Fig. 5, estimates a value corresponding to h_{11} in Equation 36, and outputs a channel estimation signal 706_1.

25 [0115]

A channel fluctuation estimating unit 705_2 for the modulated signal z_2 transmitted by the transmission device receives the baseband signal 704_X as an input, extracts a reference symbol 501_2 for channel estimation as in Fig. 5,

estimates a value corresponding to h_{12} in Equation 36, and outputs a channel estimation signal 706_2.

[0116]

A wireless unit 703_Y receives, as input, a received signal 702_Y received
5 by an antenna 701_Y, performs processing such as frequency conversion, quadrature demodulation, and the like, and outputs a baseband signal 704_Y.

[0117]

A channel fluctuation estimating unit 707_1 for the modulated signal z_1 transmitted by the transmission device receives the baseband signal 704_Y as an
10 input, extracts a reference symbol 501_1 for channel estimation as in Fig. 5, estimates a value corresponding to h_{21} in Equation 36, and outputs a channel estimation signal 708_1.

[0118]

A channel fluctuation estimating unit 707_2 for the modulated signal z_2 transmitted by the transmission device receives the baseband signal 704_Y as an
15 input, extracts a reference symbol 501_2 for channel estimation as in Fig. 5, estimates a value corresponding to h_{22} in Equation 36, and outputs a channel estimation signal 708_2.

[0119]

20 A control information decoding unit 709 receives the baseband signal 704_X and the baseband signal 704_Y as inputs, detects the symbol 500_1 that indicates the transmission method as in Fig. 5, and outputs a signal 710 regarding information on the transmission method indicated by the transmission device.

[0120]

25 A signal processing unit 711 receives, as inputs, the baseband signals 704_X and 704_Y, the channel estimation signals 706_1, 706_2, 708_1, and 708_2, and the signal 710 regarding information on the transmission method indicated by

the transmission device, performs detection and decoding, and outputs received data 712_1 and 712_2.

[0121]

Next, operations by the signal processing unit 711 in Fig. 7 are described in
 5 detail. Fig. 8 is an example of the structure of the signal processing unit 711 in the
 present embodiment. Fig. 8 shows an INNER MIMO detector, a soft-in/soft-out
 decoder, and a weighting coefficient generating unit as the main elements.
 Non-Patent Literature 2 and Non-Patent Literature 3 describe the method of iterative
 decoding with this structure. The MIMO system described in Non-Patent Literature
 10 2 and Non-Patent Literature 3 is a spatial multiplexing MIMO system, whereas the
 present embodiment differs from Non-Patent Literature 2 and Non-Patent Literature
 3 by describing a MIMO system that changes precoding weights with time. Letting
 the (channel) matrix in Equation 36 be $H(t)$, the precoding weight matrix in Fig. 6 be
 $W(t)$ (where the precoding weight matrix changes over t), the received vector be $R(t)$
 15 $= (r1(t), r2(t))^T$, and the stream vector be $S(t) = (s1(t), s2(t))^T$, the following Equation
 holds.

[0122]

Math 41

Equation 41

$$20 \quad R(t) = H(t)W(t)S(t)$$

[0123]

In this case, the reception device can apply the decoding method in
 Non-Patent Literature 2 and Non-Patent Literature 3 to the received vector $R(t)$ by
 25 considering $H(t)W(t)$ as the channel matrix.

[0124]

Therefore, a weighting coefficient generating unit 819 in Fig. 8 receives, as input, a signal 818 regarding information on the transmission method indicated by the transmission device (corresponding to 710 in Fig. 7) and outputs a signal 820 regarding information on weighting coefficients.

5 [0125]

An INNER MIMO detector 803 receives the signal 820 regarding information on weighting coefficients as input and, using the signal 820, performs the calculation in Equation 41. Iterative detection and decoding is thus performed. The following describes operations thereof.

10 [0126]

In the signal processing unit in Fig. 8, a processing method such as that shown in Fig. 10 is necessary for iterative decoding (iterative detection). First, one codeword (or one frame) of the modulated signal (stream) s1 and one codeword (or one frame) of the modulated signal (stream) s2 are decoded. As a result, the Log-Likelihood Ratio (LLR) of each bit of the one codeword (or one frame) of the modulated signal (stream) s1 and of the one codeword (or one frame) of the modulated signal (stream) s2 is obtained from the soft-in/soft-out decoder. Detection and decoding is performed again using the LLR. These operations are performed multiple times (these operations being referred to as iterative decoding (iterative detection)). Hereinafter, description focuses on the method of generating the log-likelihood ratio (LLR) of a symbol at a particular time in one frame.

[0127]

In Fig. 8, a storage unit 815 receives, as inputs, a baseband signal 801X (corresponding to the baseband signal 704_X in Fig. 7), a channel estimation signal group 802X (corresponding to the channel estimation signals 706_1 and 706_2 in Fig. 7), a baseband signal 801Y (corresponding to the baseband signal 704_Y in Fig. 7), and a channel estimation signal group 802Y (corresponding to the channel estimation signals 708_1 and 708_2 in Fig. 7). In order to achieve iterative decoding

(iterative detection), the storage unit 815 calculates $H(t)W(t)$ in Equation 41 and stores the calculated matrix as a transformed channel signal group. The storage unit 815 outputs the above signals when necessary as a baseband signal 816X, a transformed channel estimation signal group 817X, a baseband signal 816Y, and a transformed channel estimation signal group 817Y.

[0128]

Subsequent operations are described separately for initial detection and for iterative decoding (iterative detection).

[0129]

10 <Initial Detection>

The INNER MIMO detector 803 receives, as inputs, the baseband signal 801X, the channel estimation signal group 802X, the baseband signal 801Y, and the channel estimation signal group 802Y. Here, the modulation method for the modulated signal (stream) s1 and the modulated signal (stream) s2 is described as 16QAM.

[0130]

The INNER MIMO detector 803 first calculates $H(t)W(t)$ from the channel estimation signal group 802X and the channel estimation signal group 802Y to seek candidate signal points corresponding to the baseband signal 801X. Fig. 11 shows such calculation. In Fig. 11, each black dot (●) is a candidate signal point in the IQ plane. Since the modulation method is 16QAM, there are 256 candidate signal points. (Since Fig. 11 is only for illustration, not all 256 candidate signal points are shown.) Here, letting the four bits transferred by modulated signal s1 be b0, b1, b2, and b3, and the four bits transferred by modulated signal s2 be b4, b5, b6, and b7, candidate signal points corresponding to (b0, b1, b2, b3, b4, b5, b6, b7) in Fig. 11 exist. The squared Euclidian distance is sought between a received signal point 1101 (corresponding to the baseband signal 801X) and each candidate signal point. Each squared Euclidian distance is divided by the noise variance σ^2 . Accordingly, $E_x(b0,$

b1, b2, b3, b4, b5, b6, b7), i.e. the value of the squared Euclidian distance between a candidate signal point corresponding to (b0, b1, b2, b3, b4, b5, b6, b7) and a received signal point, divided by the noise variance, is sought. Note that the baseband signals and the modulated signals s1 and s2 are each complex signals.

5 [0131]

Similarly, $H(t)W(t)$ is calculated from the channel estimation signal group 802X and the channel estimation signal group 802Y, candidate signal points corresponding to the baseband signal 801Y are sought, the squared Euclidian distance for the received signal point (corresponding to the baseband signal 801Y) is
 10 sought, and the squared Euclidian distance is divided by the noise variance σ^2 . Accordingly, $E_Y(b0, b1, b2, b3, b4, b5, b6, b7)$, i.e. the value of the squared Euclidian distance between a candidate signal point corresponding to (b0, b1, b2, b3, b4, b5, b6, b7) and a received signal point, divided by the noise variance, is sought.

[0132]

15 Then $E_X(b0, b1, b2, b3, b4, b5, b6, b7) + E_Y(b0, b1, b2, b3, b4, b5, b6, b7) = E(b0, b1, b2, b3, b4, b5, b6, b7)$ is sought.

[0133]

The INNER MIMO detector 803 outputs $E(b0, b1, b2, b3, b4, b5, b6, b7)$ as a signal 804.

20 [0134]

A log-likelihood calculating unit 805A receives the signal 804 as input, calculates the log likelihood for bits b0, b1, b2, and b3, and outputs a log-likelihood signal 806A. Note that during calculation of the log likelihood, the log likelihood for “1” and the log likelihood for “0” are calculated. The calculation method is as shown
 25 in Equations 28, 29, and 30. Details can be found in Non-Patent Literature 2 and Non-Patent Literature 3.

[0135]

Similarly, a log-likelihood calculating unit 805B receives the signal 804 as input, calculates the log likelihood for bits b4, b5, b6, and b7, and outputs a log-likelihood signal 806B.

[0136]

5 A deinterleaver (807A) receives the log-likelihood signal 806A as an input, performs deinterleaving corresponding to the interleaver (the interleaver (304A) in Fig. 3), and outputs a deinterleaved log-likelihood signal 808A.

[0137]

10 Similarly, a deinterleaver (807B) receives the log-likelihood signal 806B as an input, performs deinterleaving corresponding to the interleaver (the interleaver (304B) in Fig. 3), and outputs a deinterleaved log-likelihood signal 808B.

[0138]

A log-likelihood ratio calculating unit 809A receives the interleaved log-likelihood signal 808A as an input, calculates the log-likelihood ratio (LLR) of the bits encoded by the encoder 302A in Fig. 3, and outputs a log-likelihood ratio signal 810A.

[0139]

20 Similarly, a log-likelihood ratio calculating unit 809B receives the interleaved log-likelihood signal 808B as an input, calculates the log-likelihood ratio (LLR) of the bits encoded by the encoder 302B in Fig. 3, and outputs a log-likelihood ratio signal 810B.

[0140]

A soft-in/soft-out decoder 811A receives the log-likelihood ratio signal 810A as an input, performs decoding, and outputs a decoded log-likelihood ratio 25 812A.

[0141]

Similarly, a soft-in/soft-out decoder 811B receives the log-likelihood ratio signal 810B as an input, performs decoding, and outputs a decoded log-likelihood ratio 812B.

[0142]

5 <Iterative Decoding (Iterative Detection), Number of Iterations k >

An interleaver (813A) receives the log-likelihood ratio 812A decoded by the soft-in/soft-out decoder in the $(k - 1)^{\text{th}}$ iteration as an input, performs interleaving, and outputs an interleaved log-likelihood ratio 814A. The interleaving pattern in the interleaver (813A) is similar to the interleaving pattern in the
10 interleaver (304A) in Fig. 3.

[0143]

An interleaver (813B) receives the log-likelihood ratio 812B decoded by the soft-in/soft-out decoder in the $(k - 1)^{\text{th}}$ iteration as an input, performs interleaving, and outputs an interleaved log-likelihood ratio 814B. The interleaving pattern in the
15 interleaver (813B) is similar to the interleaving pattern in the interleaver (304B) in Fig. 3.

[0144]

The INNER MIMO detector 803 receives, as inputs, the baseband signal 816X, the transformed channel estimation signal group 817X, the baseband signal
20 816Y, the transformed channel estimation signal group 817Y, the interleaved log-likelihood ratio 814A, and the interleaved log-likelihood ratio 814B. The reason for using the baseband signal 816X, the transformed channel estimation signal group 817X, the baseband signal 816Y, and the transformed channel estimation signal group 817Y instead of the baseband signal 801X, the channel estimation signal group 802X, the baseband signal 801Y, and the channel estimation signal group
25 802Y is because a delay occurs due to iterative decoding.

[0145]

The difference between operations by the INNER MIMO detector 803 for iterative decoding and for initial detection is the use of the interleaved log-likelihood ratio 814A and the interleaved log-likelihood ratio 814B during signal processing. The INNER MIMO detector 803 first seeks $E(b_0, b_1, b_2, b_3, b_4, b_5, b_6, b_7)$, as
 5 during initial detection. Additionally, coefficients corresponding to Equations 11 and 32 are sought from the interleaved log-likelihood ratio 814A and the interleaved log-likelihood ratio 914B. The value $E(b_0, b_1, b_2, b_3, b_4, b_5, b_6, b_7)$ is adjusted using the sought coefficients, and the resulting value $E'(b_0, b_1, b_2, b_3, b_4, b_5, b_6, b_7)$ is output as the signal 804.

10 [0146]

The log-likelihood calculating unit 805A receives the signal 804 as input, calculates the log likelihood for bits b_0, b_1, b_2 , and b_3 , and outputs the log-likelihood signal 806A. Note that during calculation of the log likelihood, the log likelihood for “1” and the log likelihood for “0” are calculated. The calculation
 15 method is as shown in Equations 31, 32, 33, 34, and 35. Details can be found in Non-Patent Literature 2 and Non-Patent Literature 3.

[0147]

Similarly, the log-likelihood calculating unit 805B receives the signal 804 as input, calculates the log likelihood for bits b_4, b_5, b_6 , and b_7 , and outputs the
 20 log-likelihood signal 806B. Operations by the deinterleaver onwards are similar to initial detection.

[0148]

Note that while Fig. 8 shows the structure of the signal processing unit when performing iterative detection, iterative detection is not always essential for
 25 obtaining excellent reception quality, and a structure not including the interleavers 813A and 813B, which are necessary only for iterative detection, is possible. In such a case, the INNER MIMO detector 803 does not perform iterative detection.

[0149]

The main part of the present embodiment is calculation of $H(t)W(t)$. Note that as shown in Non-Patent Literature 5 and the like, QR decomposition may be used to perform initial detection and iterative detection.

[0150]

5 Furthermore, as shown in Non-Patent Literature 11, based on $H(t)W(t)$, linear operation of the Minimum Mean Squared Error (MMSE) and Zero Forcing (ZF) may be performed in order to perform initial detection.

[0151]

Fig. 9 is the structure of a different signal processing unit than Fig. 8 and is for the modulated signal transmitted by the transmission device in Fig. 4. The difference with Fig. 8 is the number of soft-in/soft-out decoders. A soft-in/soft-out decoder 901 receives, as inputs, the log-likelihood ratio signals 810A and 810B, performs decoding, and outputs a decoded log-likelihood ratio 902. A distribution unit 903 receives the decoded log-likelihood ratio 902 as an input and distributes the log-likelihood ratio 902. Other operations are similar to Fig. 8.

[0152]

Figs. 12A and 12B show BER characteristics for a transmission method using the precoding weights of the present embodiment under similar conditions to Figs. 29A and 29B. Fig. 12A shows the BER characteristics of Max-log A Posteriori Probability (APP) without iterative detection (see Non-Patent Literature 1 and Non-Patent Literature 2), and Fig. 12B shows the BER characteristics of Max-log-APP with iterative detection (see Non-Patent Literature 1 and Non-Patent Literature 2) (number of iterations: five). Comparing Figs. 12A, 12B, 29A, and 29B shows how if the transmission method of the present embodiment is used, the BER characteristics when the Rician factor is large greatly improve over the BER characteristics when using spatial multiplexing MIMO system, thereby confirming the usefulness of the method in the present embodiment.

[0153]

As described above, when a transmission device transmits a plurality of modulated signals from a plurality of antennas in a MIMO system, the advantageous effect of improved transmission quality, as compared to conventional spatial multiplexing MIMO system, is achieved in an LOS environment in which direct waves dominate by hopping between precoding weights regularly over time, as in the present embodiment.

[0154]

In the present embodiment, and in particular with regards to the structure of the reception device, operations have been described for a limited number of antennas, but the present invention may be embodied in the same way even if the number of antennas increases. In other words, the number of antennas in the reception device does not affect the operations or advantageous effects of the present embodiment. Furthermore, in the present embodiment, the example of LDPC coding has particularly been explained, but the present invention is not limited to LDPC coding. Furthermore, with regards to the decoding method, the soft-in/soft-out decoders are not limited to the example of sum-product decoding. Another soft-in/soft-out decoding method may be used, such as a BCJR algorithm, a SOVA algorithm, a Max-log-MAP algorithm, and the like. Details are provided in Non-Patent Literature 6.

[0155]

Additionally, in the present embodiment, the example of a single carrier method has been described, but the present invention is not limited in this way and may be similarly embodied for multi-carrier transmission. Accordingly, when using a method such as spread spectrum communication, Orthogonal Frequency-Division Multiplexing (OFDM), Single Carrier Frequency Division Multiple Access (SC-FDMA), Single Carrier Orthogonal Frequency-Division Multiplexing (SC-OFDM), or wavelet OFDM as described in Non-Patent Literature 7 and the like, for example, the present invention may be similarly embodied. Furthermore, in the

present embodiment, symbols other than data symbols, such as pilot symbols (preamble, unique word, and the like), symbols for transmission of control information, and the like, may be arranged in the frame in any way.

[0156]

5 The following describes an example of using OFDM as an example of a multi-carrier method.

[0157]

Fig. 13 shows the structure of a transmission device when using OFDM. In Fig. 13, elements that operate in a similar way to Fig. 3 bear the same reference
10 signs.

[0158]

An OFDM related processor 1301A receives, as input, the weighted signal 309A, performs processing related to OFDM, and outputs a transmission signal 1302A. Similarly, an OFDM related processor 1301B receives, as input, the
15 weighted signal 309B, performs processing related to OFDM, and outputs a transmission signal 1302B.

[0159]

Fig. 14 shows an example of a structure from the OFDM related processors 1301A and 1301B in Fig. 13 onwards. The part from 1401A to 1410A is related to
20 the part from 1301A to 312A in Fig. 13, and the part from 1401B to 1410B is related to the part from 1301B to 312B in Fig. 13.

[0160]

A serial/parallel converter 1402A performs serial/parallel conversion on a weighted signal 1401A (corresponding to the weighted signal 309A in Fig. 13) and
25 outputs a parallel signal 1403A.

[0161]

A reordering unit 1404A receives a parallel signal 1403A as input, performs reordering, and outputs a reordered signal 1405A. Reordering is described in detail later.

[0162]

5 An inverse fast Fourier transformer 1406A receives the reordered signal 1405A as an input, performs a fast Fourier transform, and outputs a fast Fourier transformed signal 1407A.

[0163]

10 A wireless unit 1408A receives the fast Fourier transformed signal 1407A as an input, performs processing such as frequency conversion, amplification, and the like, and outputs a modulated signal 1409A. The modulated signal 1409A is output as a radio wave from an antenna 1410A.

[0164]

15 A serial/parallel converter 1402B performs serial/parallel conversion on a weighted signal 1401B (corresponding to the weighted signal 309B in Fig. 13) and outputs a parallel signal 1403B.

[0165]

20 A reordering unit 1404B receives a parallel signal 1403B as input, performs reordering, and outputs a reordered signal 1405B. Reordering is described in detail later.

[0166]

An inverse fast Fourier transformer 1406B receives the reordered signal 1405B as an input, performs a fast Fourier transform, and outputs a fast Fourier transformed signal 1407B.

25 [0167]

A wireless unit 1408B receives the fast Fourier transformed signal 1407B as an input, performs processing such as frequency conversion, amplification, and the

like, and outputs a modulated signal 1409B. The modulated signal 1409B is output as a radio wave from an antenna 1410B.

[0168]

In the transmission device of Fig. 3, since the transmission method does not use multi-carrier, precoding hops to form a four-slot period (cycle), as shown in Fig. 6, and the precoded symbols are arranged in the time domain. When using a multi-carrier transmission method as in the OFDM method shown in Fig. 13, it is of course possible to arrange the precoded symbols in the time domain as in Fig. 3 for each (sub)carrier. In the case of a multi-carrier transmission method, however, it is possible to arrange symbols in the frequency domain, or in both the frequency and time domains. The following describes these arrangements.

[0169]

Figs. 15A and 15B show an example of a method of reordering symbols by reordering units 1401A and 1401B in Fig. 14, the horizontal axis representing frequency, and the vertical axis representing time. The frequency domain runs from (sub)carrier 0 through (sub)carrier 9. The modulated signals z1 and z2 use the same frequency bandwidth at the same time. Fig. 15A shows the reordering method for symbols of the modulated signal z1, and Fig. 15B shows the reordering method for symbols of the modulated signal z2. Numbers #1, #2, #3, #4, ... are assigned in order to the symbols of the weighted signal 1401A which is input into the serial/parallel converter 1402A. At this point, symbols are assigned regularly, as shown in Fig. 15A. The symbols #1, #2, #3, #4, ... are arranged in order starting from carrier 0. The symbols #1 through #9 are assigned to time \$1, and subsequently, the symbols #10 through #19 are assigned to time \$2.

[0170]

Similarly, numbers #1, #2, #3, #4, ... are assigned in order to the symbols of the weighted signal 1401B which is input into the serial/parallel converter 1402B. At this point, symbols are assigned regularly, as shown in Fig. 15B. The symbols #1,

#2, #3, #4, ... are arranged in order starting from carrier 0. The symbols #1 through #9 are assigned to time \$1, and subsequently, the symbols #10 through #19 are assigned to time \$2. Note that the modulated signals $z1$ and $z2$ are complex signals.

[0171]

5 The symbol group 1501 and the symbol group 1502 shown in Figs. 15A and 15B are the symbols for one period (cycle) when using the precoding weight hopping method shown in Fig. 6. Symbol #0 is the symbol when using the precoding weight of slot $4i$ in Fig. 6. Symbol #1 is the symbol when using the precoding weight of slot $4i + 1$ in Fig. 6. Symbol #2 is the symbol when using the precoding weight of slot $4i + 2$ in Fig. 6. Symbol #3 is the symbol when using the precoding weight of slot $4i + 3$ in Fig. 6. Accordingly, symbol # x is as follows. When $x \bmod 4$ is 0, the symbol # x is the symbol when using the precoding weight of slot $4i$ in Fig. 6. When $x \bmod 4$ is 1, the symbol # x is the symbol when using the precoding weight of slot $4i + 1$ in Fig. 6. When $x \bmod 4$ is 2, the symbol # x is the symbol when using the precoding weight of slot $4i + 2$ in Fig. 6. When $x \bmod 4$ is 3, the symbol # x is the symbol when using the precoding weight of slot $4i + 3$ in Fig. 6.

[0172]

20 In this way, when using a multi-carrier transmission method such as OFDM, unlike during single carrier transmission, symbols can be arranged in the frequency domain. Furthermore, the ordering of symbols is not limited to the ordering shown in Figs. 15A and 15B. Other examples are described with reference to Figs. 16A, 16B, 17A, and 17B.

[0173]

25 Figs. 16A and 16B show an example of a method of reordering symbols by the reordering units 1404A and 1404B in Fig. 14, the horizontal axis representing frequency, and the vertical axis representing time, that differs from Figs. 15A and 15B. Fig. 16A shows the reordering method for symbols of the modulated signal $z1$, and Fig. 16B shows the reordering method for symbols of the modulated signal $z2$.

The difference in Figs. 16A and 16B as compared to Figs. 15A and 15B is that the reordering method of the symbols of the modulated signal z1 differs from the reordering method of the symbols of the modulated signal z2. In Fig. 16B, symbols #0 through #5 are assigned to carriers 4 through 9, and symbols #6 through #9 are assigned to carriers 0 through 3. Subsequently, symbols #10 through #19 are assigned regularly in the same way. At this point, as in Figs. 15A and 15B, the symbol group 1601 and the symbol group 1602 shown in Figs. 16A and 16B are the symbols for one period (cycle) when using the precoding weight hopping method shown in Fig. 6.

10 [0174]

Figs. 17A and 17B show an example of a method of reordering symbols by the reordering units 1404A and 1404B in Fig. 14, the horizontal axis representing frequency, and the vertical axis representing time, that differs from Figs. 15A and 15B. Fig. 17A shows the reordering method for symbols of the modulated signal z1, and Fig. 17B shows the reordering method for symbols of the modulated signal z2. The difference in Figs. 17A and 17B as compared to Figs. 15A and 15B is that whereas the symbols are arranged in order by carrier in Figs. 15A and 15B, the symbols are not arranged in order by carrier in Figs. 17A and 17B. It is obvious that, in Figs. 17A and 17B, the reordering method of the symbols of the modulated signal z1 may differ from the reordering method of the symbols of the modulated signal z2, as in Figs. 16A and 16B.

[0175]

Figs. 18A and 18B show an example of a method of reordering symbols by the reordering units 1404A and 1404B in Fig. 14, the horizontal axis representing frequency, and the vertical axis representing time, that differs from Figs. 15A through 17B. Fig. 18A shows the reordering method for symbols of the modulated signal z1, and Fig. 18B shows the reordering method for symbols of the modulated signal z2. In Figs. 15A through 17B, symbols are arranged in the frequency domain,

whereas in Figs. 18A and 18B, symbols are arranged in both the frequency and time domains.

[0176]

In Fig. 6, an example has been described of hopping between precoding weights over four slots. Here, however, an example of hopping over eight slots is described. The symbol groups 1801 and 1802 shown in Figs. 18A and 18B are the symbols for one period (cycle) when using the precoding weight hopping method (and are therefore eight-symbol groups). Symbol #0 is the symbol when using the precoding weight of slot $8i$. Symbol #1 is the symbol when using the precoding weight of slot $8i + 1$. Symbol #2 is the symbol when using the precoding weight of slot $8i + 2$. Symbol #3 is the symbol when using the precoding weight of slot $8i + 3$. Symbol #4 is the symbol when using the precoding weight of slot $8i + 4$. Symbol #5 is the symbol when using the precoding weight of slot $8i + 5$. Symbol #6 is the symbol when using the precoding weight of slot $8i + 6$. Symbol #7 is the symbol when using the precoding weight of slot $8i + 7$. Accordingly, symbol # x is as follows. When $x \bmod 8$ is 0, the symbol # x is the symbol when using the precoding weight of slot $8i$. When $x \bmod 8$ is 1, the symbol # x is the symbol when using the precoding weight of slot $8i + 1$. When $x \bmod 8$ is 2, the symbol # x is the symbol when using the precoding weight of slot $8i + 2$. When $x \bmod 8$ is 3, the symbol # x is the symbol when using the precoding weight of slot $8i + 3$. When $x \bmod 8$ is 4, the symbol # x is the symbol when using the precoding weight of slot $8i + 4$. When $x \bmod 8$ is 5, the symbol # x is the symbol when using the precoding weight of slot $8i + 5$. When $x \bmod 8$ is 6, the symbol # x is the symbol when using the precoding weight of slot $8i + 6$. When $x \bmod 8$ is 7, the symbol # x is the symbol when using the precoding weight of slot $8i + 7$. In the symbol ordering in Figs. 18A and 18B, four slots in the time domain and two slots in the frequency domain for a total of $4 \times 2 = 8$ slots are used to arrange symbols for one period (cycle). In this case, letting the number of symbols in one period (cycle) be $m \times n$ symbols (in other words, $m \times$

n precoding weights exist), the number of slots (the number of carriers) in the frequency domain used to arrange symbols in one period (cycle) be n , and the number of slots used in the time domain be m , m should be greater than n . This is because the phase of direct waves fluctuates more slowly in the time domain than in the frequency domain. Therefore, since the precoding weights are changed in the present embodiment to minimize the influence of steady direct waves, it is preferable to reduce the fluctuation in direct waves in the period (cycle) for changing the precoding weights. Accordingly, m should be greater than n . Furthermore, considering the above points, rather than reordering symbols only in the frequency domain or only in the time domain, direct waves are more likely to become stable when symbols are reordered in both the frequency and the time domains as in Figs. 18A and 18B, thereby making it easier to achieve the advantageous effects of the present invention. When symbols are ordered in the frequency domain, however, fluctuations in the frequency domain are abrupt, leading to the possibility of yielding diversity gain. Therefore, reordering in both the frequency and the time domains is not necessarily always the best method.

[0177]

Figs. 19A and 19B show an example of a method of reordering symbols by the reordering units 1404A and 1404B in Fig. 14, the horizontal axis representing frequency, and the vertical axis representing time, that differs from Figs. 18A and 18B. Fig. 19A shows the reordering method for symbols of the modulated signal z_1 , and Fig. 19B shows the reordering method for symbols of the modulated signal z_2 . As in Figs. 18A and 18B, Figs. 19A and 19B show arrangement of symbols using both the frequency and the time axes. The difference as compared to Figs. 18A and 18B is that, whereas symbols are arranged first in the frequency domain and then in the time domain in Figs. 18A and 18B, symbols are arranged first in the time domain and then in the frequency domain in Figs. 19A and 19B. In Figs. 19A and

19B, the symbol group 1901 and the symbol group 1902 are the symbols for one period (cycle) when using the precoding hopping method.

[0178]

Note that in Figs. 18A, 18B, 19A, and 19B, as in Figs. 16A and 16B, the present invention may be similarly embodied, and the advantageous effect of high reception quality achieved, with the symbol arranging method of the modulated signal z1 differing from the symbol arranging method of the modulated signal z2. Furthermore, in Figs. 18A, 18B, 19A, and 19B, as in Figs. 17A and 17B, the present invention may be similarly embodied, and the advantageous effect of high reception quality achieved, without arranging the symbols in order.

[0179]

Fig. 27 shows an example of a method of reordering symbols by the reordering units 1404A and 1404B in Fig. 14, the horizontal axis representing frequency, and the vertical axis representing time. that differs from the above examples. The case of hopping between precoding matrix regularly over four slots, as in Equations 37–40, is considered. The characteristic feature of Fig. 27 is that symbols are arranged in order in the frequency domain, but when progressing in the time domain, symbols are cyclically shifted by n symbols (in the example in Fig. 27, $n = 1$). In the four symbols shown in the symbol group 2710 in the frequency domain in Fig. 27, precoding hops between the precoding matrices of Equations 37–40.

[0180]

In this case, symbol #0 is precoded using the precoding matrix in Equation 37, symbol #1 is precoded using the precoding matrix in Equation 38, symbol #2 is precoded using the precoding matrix in Equation 39, and symbol #3 is precoded using the precoding matrix in Equation 40.

[0181]

Similarly, for the symbol group 2720 in the frequency domain, symbol #4 is precoded using the precoding matrix in Equation 37, symbol #5 is precoded using the precoding matrix in Equation 38, symbol #6 is precoded using the precoding matrix in Equation 39, and symbol #7 is precoded using the precoding matrix in
 5 Equation 40.

[0182]

For the symbols at time \$1, precoding hops between the above precoding matrices, but in the time domain, symbols are cyclically shifted. Therefore, precoding hops between precoding matrices for the symbol groups 2701, 2702, 2703,
 10 and 2704 as follows.

[0183]

In the symbol group 2701 in the time domain, symbol #0 is precoded using the precoding matrix in Equation 37, symbol #9 is precoded using the precoding matrix in Equation 38, symbol #18 is precoded using the precoding matrix in
 15 Equation 39, and symbol #27 is precoded using the precoding matrix in Equation 40.

[0184]

In the symbol group 2702 in the time domain, symbol #28 is precoded using the precoding matrix in Equation 37, symbol #1 is precoded using the precoding matrix in Equation 38, symbol #10 is precoded using the precoding matrix in
 20 Equation 39, and symbol #19 is precoded using the precoding matrix in Equation 40.

[0185]

In the symbol group 2703 in the time domain, symbol #20 is precoded using the precoding matrix in Equation 37, symbol #29 is precoded using the precoding matrix in Equation 38, symbol #2 is precoded using the precoding matrix in
 25 Equation 39, and symbol #11 is precoded using the precoding matrix in Equation 40.

[0186]

In the symbol group 2704 in the time domain, symbol #12 is precoded using the precoding matrix in Equation 37, symbol #21 is precoded using the precoding

matrix in Equation 38, symbol #30 is precoded using the precoding matrix in Equation 39, and symbol #3 is precoded using the precoding matrix in Equation 40.

[0187]

The characteristic of Fig. 27 is that, for example focusing on symbol #11, the symbols on either side in the frequency domain at the same time (symbols #10 and #12) are both precoded with a different precoding matrix than symbol #11, and the symbols on either side in the time domain in the same carrier (symbols #2 and #20) are both precoded with a different precoding matrix than symbol #11. This is true not only for symbol #11. Any symbol having symbols on either side in the frequency domain and the time domain is characterized in the same way as symbol #11. As a result, precoding matrices are effectively hopped between, and since the influence on stable conditions of direct waves is reduced, the possibility of improved reception quality of data increases.

[0188]

In Fig. 27, the case of $n = 1$ has been described, but n is not limited in this way. The present invention may be similarly embodied with $n = 3$. Furthermore, in Fig. 27, when symbols are arranged in the frequency domain and time progresses in the time domain, the above characteristic is achieved by cyclically shifting the number of the arranged symbol, but the above characteristic may also be achieved by randomly (or regularly) arranging the symbols.

[0189]

(Embodiment 2)

In Embodiment 1, regular hopping of the precoding weights as shown in Fig. 6 has been described. In the present embodiment, a method for designing specific precoding weights that differ from the precoding weights in Fig. 6 is described.

[0190]

In Fig. 6, the method for hopping between the precoding weights in Equations 37–40 has been described. By generalizing this method, the precoding

weights may be changed as follows. (The hopping period (cycle) for the precoding weights has four slots, and Equations are listed similarly to Equations 37–40.)

For symbol number $4i$ (where i is an integer greater than or equal to zero):

[0191]

5 Math 42

Equation 42

$$\begin{pmatrix} z1(4i) \\ z2(4i) \end{pmatrix} = \frac{1}{\sqrt{2}} \begin{pmatrix} e^{j\theta_{11}(4i)} & e^{j(\theta_{11}(4i)+\lambda)} \\ e^{j\theta_{21}(4i)} & e^{j(\theta_{21}(4i)+\lambda+\delta)} \end{pmatrix} \begin{pmatrix} s1(4i) \\ s2(4i) \end{pmatrix}$$

[0192]

10 Here, j is an imaginary unit.

For symbol number $4i + 1$:

[0193]

Math 43

Equation 43

$$15 \quad \begin{pmatrix} z1(4i+1) \\ z2(4i+1) \end{pmatrix} = \frac{1}{\sqrt{2}} \begin{pmatrix} e^{j\theta_{11}(4i+1)} & e^{j(\theta_{11}(4i+1)+\lambda)} \\ e^{j\theta_{21}(4i+1)} & e^{j(\theta_{21}(4i+1)+\lambda+\delta)} \end{pmatrix} \begin{pmatrix} s1(4i+1) \\ s2(4i+1) \end{pmatrix}$$

[0194]

For symbol number $4i + 2$:

[0195]

20 Math 44

Equation 44

$$\begin{pmatrix} z1(4i+2) \\ z2(4i+2) \end{pmatrix} = \frac{1}{\sqrt{2}} \begin{pmatrix} e^{j\theta_{11}(4i+2)} & e^{j(\theta_{11}(4i+2)+\lambda)} \\ e^{j\theta_{21}(4i+2)} & e^{j(\theta_{21}(4i+2)+\lambda+\delta)} \end{pmatrix} \begin{pmatrix} s1(4i+2) \\ s2(4i+2) \end{pmatrix}$$

[0196]

For symbol number $4i + 3$:

[0197]

5 Math 45

Equation 45

$$\begin{pmatrix} z1(4i+3) \\ z2(4i+3) \end{pmatrix} = \frac{1}{\sqrt{2}} \begin{pmatrix} e^{j\theta_{11}(4i+3)} & e^{j(\theta_{11}(4i+3)+\lambda)} \\ e^{j\theta_{21}(4i+3)} & e^{j(\theta_{21}(4i+3)+\lambda+\delta)} \end{pmatrix} \begin{pmatrix} s1(4i+3) \\ s2(4i+3) \end{pmatrix}$$

[0198]

10 From Equations 36 and 41, the received vector $R(t) = (r1(t), r2(t))^T$ can be represented as follows.

For symbol number $4i$:

[0199]

Math 46

15 Equation 46

$$\begin{pmatrix} r1(4i) \\ r2(4i) \end{pmatrix} = \frac{1}{\sqrt{2}} \begin{pmatrix} h_{11}(4i) & h_{12}(4i) \\ h_{21}(4i) & h_{22}(4i) \end{pmatrix} \begin{pmatrix} e^{j\theta_{11}(4i)} & e^{j(\theta_{11}(4i)+\lambda)} \\ e^{j\theta_{21}(4i)} & e^{j(\theta_{21}(4i)+\lambda+\delta)} \end{pmatrix} \begin{pmatrix} s1(4i) \\ s2(4i) \end{pmatrix}$$

[0200]

For symbol number $4i + 1$:

20 [0201]

Math 47

Equation 47

$$\begin{pmatrix} r1(4i+1) \\ r2(4i+1) \end{pmatrix} = \frac{1}{\sqrt{2}} \begin{pmatrix} h_{11}(4i+1) & h_{12}(4i+1) \\ h_{21}(4i+1) & h_{22}(4i+1) \end{pmatrix} \begin{pmatrix} e^{j\theta_{11}(4i+1)} & e^{j(\theta_{11}(4i+1)+\lambda)} \\ e^{j\theta_{21}(4i+1)} & e^{j(\theta_{21}(4i+1)+\lambda+\delta)} \end{pmatrix} \begin{pmatrix} s1(4i+1) \\ s2(4i+1) \end{pmatrix}$$

[0202]

For symbol number $4i+2$:

5 [0203]

Math 48

Equation 48

$$\begin{pmatrix} r1(4i+2) \\ r2(4i+2) \end{pmatrix} = \frac{1}{\sqrt{2}} \begin{pmatrix} h_{11}(4i+2) & h_{12}(4i+2) \\ h_{21}(4i+2) & h_{22}(4i+2) \end{pmatrix} \begin{pmatrix} e^{j\theta_{11}(4i+2)} & e^{j(\theta_{11}(4i+2)+\lambda)} \\ e^{j\theta_{21}(4i+2)} & e^{j(\theta_{21}(4i+2)+\lambda+\delta)} \end{pmatrix} \begin{pmatrix} s1(4i+2) \\ s2(4i+2) \end{pmatrix}$$

10 [0204]

For symbol number $4i+3$:

[0205]

Math 49

Equation 49

$$15 \quad \begin{pmatrix} r1(4i+3) \\ r2(4i+3) \end{pmatrix} = \frac{1}{\sqrt{2}} \begin{pmatrix} h_{11}(4i+3) & h_{12}(4i+3) \\ h_{21}(4i+3) & h_{22}(4i+3) \end{pmatrix} \begin{pmatrix} e^{j\theta_{11}(4i+3)} & e^{j(\theta_{11}(4i+3)+\lambda)} \\ e^{j\theta_{21}(4i+3)} & e^{j(\theta_{21}(4i+3)+\lambda+\delta)} \end{pmatrix} \begin{pmatrix} s1(4i+3) \\ s2(4i+3) \end{pmatrix}$$

[0206]

In this case, it is assumed that only components of direct waves exist in the channel elements $h_{11}(t)$, $h_{12}(t)$, $h_{21}(t)$, and $h_{22}(t)$, that the amplitude components of the direct waves are all equal, and that fluctuations do not occur over time. With these assumptions, Equations 46–49 can be represented as follows.

For symbol number $4i$:

[0207]

Math 50

Equation 50

$$\begin{pmatrix} r1(4i) \\ r2(4i) \end{pmatrix} = \frac{1}{\sqrt{2}} \begin{pmatrix} A e^{j0} & q \\ A e^{j0} & q \end{pmatrix} \begin{pmatrix} e^{j\theta_{11}(4i)} & e^{j(\theta_{11}(4i)+\lambda)} \\ e^{j\theta_{21}(4i)} & e^{j(\theta_{21}(4i)+\lambda+\delta)} \end{pmatrix} \begin{pmatrix} s1(4i) \\ s2(4i) \end{pmatrix}$$

[0208]

5 For symbol number $4i + 1$:

[0209]

Math 51

Equation 51

$$\begin{pmatrix} r1(4i+1) \\ r2(4i+1) \end{pmatrix} = \frac{1}{\sqrt{2}} \begin{pmatrix} A e^{j0} & q \\ A e^{j0} & q \end{pmatrix} \begin{pmatrix} e^{j\theta_{11}(4i+1)} & e^{j(\theta_{11}(4i+1)+\lambda)} \\ e^{j\theta_{21}(4i+1)} & e^{j(\theta_{21}(4i+1)+\lambda+\delta)} \end{pmatrix} \begin{pmatrix} s1(4i+1) \\ s2(4i+1) \end{pmatrix}$$

10

[0210]

For symbol number $4i + 2$:

[0211]

Math 52

15 Equation 52

$$\begin{pmatrix} r1(4i+2) \\ r2(4i+2) \end{pmatrix} = \frac{1}{\sqrt{2}} \begin{pmatrix} A e^{j0} & q \\ A e^{j0} & q \end{pmatrix} \begin{pmatrix} e^{j\theta_{11}(4i+2)} & e^{j(\theta_{11}(4i+2)+\lambda)} \\ e^{j\theta_{21}(4i+2)} & e^{j(\theta_{21}(4i+2)+\lambda+\delta)} \end{pmatrix} \begin{pmatrix} s1(4i+2) \\ s2(4i+2) \end{pmatrix}$$

[0212]

For symbol number $4i + 3$:

20 [0213]

Math 53

Equation 53

$$\begin{pmatrix} r1(4i+3) \\ r2(4i+3) \end{pmatrix} = \frac{1}{\sqrt{2}} \begin{pmatrix} A e^{j0} & q \\ A e^{j0} & q \end{pmatrix} \begin{pmatrix} e^{j\theta_{11}(4i+3)} & e^{j(\theta_{11}(4i+3)+\lambda)} \\ e^{j\theta_{21}(4i+3)} & e^{j(\theta_{21}(4i+3)+\lambda+\delta)} \end{pmatrix} \begin{pmatrix} s1(4i+3) \\ s2(4i+3) \end{pmatrix}$$

[0214]

In Equations 50–53, let A be a positive real number and q be a complex number. The values of A and q are determined in accordance with the positional relationship between the transmission device and the reception device. Equations 50–53 can be represented as follows.

For symbol number 4i:

[0215]

10 Math 54

Equation 54

$$\begin{pmatrix} r1(4i) \\ r2(4i) \end{pmatrix} = \frac{1}{\sqrt{2}} \begin{pmatrix} e^{j0} \\ e^{j0} \end{pmatrix} \begin{pmatrix} A e^{j0} & q \\ A e^{j0} & q \end{pmatrix} \begin{pmatrix} e^{j\theta_{11}(4i)} & e^{j(\theta_{11}(4i)+\lambda)} \\ e^{j\theta_{21}(4i)} & e^{j(\theta_{21}(4i)+\lambda+\delta)} \end{pmatrix} \begin{pmatrix} s1(4i) \\ s2(4i) \end{pmatrix}$$

[0216]

15 For symbol number 4i + 1:

[0217]

Math 55

Equation 55

$$\begin{pmatrix} r1(4i+1) \\ r2(4i+1) \end{pmatrix} = \frac{1}{\sqrt{2}} \begin{pmatrix} e^{j0} \\ e^{j0} \end{pmatrix} \begin{pmatrix} A e^{j0} & q \\ A e^{j0} & q \end{pmatrix} \begin{pmatrix} e^{j\theta_{11}(4i+1)} & e^{j(\theta_{11}(4i+1)+\lambda)} \\ e^{j\theta_{21}(4i+1)} & e^{j(\theta_{21}(4i+1)+\lambda+\delta)} \end{pmatrix} \begin{pmatrix} s1(4i+1) \\ s2(4i+1) \end{pmatrix}$$

20

[0218]

For symbol number 4i + 2:

[0219]

Math 56

Equation 56

$$\begin{pmatrix} r1(4i+2) \\ r2(4i+2) \end{pmatrix} = \frac{1}{\sqrt{2}} \begin{pmatrix} e^{j0} \\ e^{j0} \end{pmatrix} \left(A e^{j0} \quad q \begin{pmatrix} e^{j\theta_{11}(4i+2)} & e^{j(\theta_{11}(4i+2)+\lambda)} \\ e^{j\theta_{21}(4i+2)} & e^{j(\theta_{21}(4i+2)+\lambda+\delta)} \end{pmatrix} \begin{pmatrix} s1(4i+2) \\ s2(4i+2) \end{pmatrix} \right)$$

5

[0220]

For symbol number $4i+3$:

[0221]

Math 57

10 Equation 57

$$\begin{pmatrix} r1(4i+3) \\ r2(4i+3) \end{pmatrix} = \frac{1}{\sqrt{2}} \begin{pmatrix} e^{j0} \\ e^{j0} \end{pmatrix} \left(A e^{j0} \quad q \begin{pmatrix} e^{j\theta_{11}(4i+3)} & e^{j(\theta_{11}(4i+3)+\lambda)} \\ e^{j\theta_{21}(4i+3)} & e^{j(\theta_{21}(4i+3)+\lambda+\delta)} \end{pmatrix} \begin{pmatrix} s1(4i+3) \\ s2(4i+3) \end{pmatrix} \right)$$

[0222]

As a result, when q is represented as follows, a signal component based on one of $s1$ and $s2$ is no longer included in $r1$ and $r2$, and therefore one of the signals $s1$ and $s2$ can no longer be obtained.

For symbol number $4i$:

[0223]

Math 58

20 Equation 58

$$q = -A e^{j(\theta_{11}(4i)-\theta_{21}(4i))}, -A e^{j(\theta_{11}(4i)-\theta_{21}(4i)-\delta)}$$

[0224]

For symbol number $4i + 1$:

[0225]

Math 59

Equation 59

$$5 \quad q = -A e^{j(\theta_{11}(4i+1)-\theta_{21}(4i+1))}, -A e^{j(\theta_{11}(4i+1)-\theta_{21}(4i+1)-\delta)}$$

[0226]

For symbol number $4i + 2$:

[0227]

10 Math 60

Equation 60

$$q = -A e^{j(\theta_{11}(4i+2)-\theta_{21}(4i+2))}, -A e^{j(\theta_{11}(4i+2)-\theta_{21}(4i+2)-\delta)}$$

[0228]

15 For symbol number $4i + 3$:

[0229]

Math 61

Equation 61

$$q = -A e^{j(\theta_{11}(4i+3)-\theta_{21}(4i+3))}, -A e^{j(\theta_{11}(4i+3)-\theta_{21}(4i+3)-\delta)}$$

20

[0230]

In this case, if q has the same solution in symbol numbers $4i$, $4i + 1$, $4i + 2$, and $4i + 3$, then the channel elements of the direct waves do not greatly fluctuate. Therefore, a reception device having channel elements in which the value of q is
25 equivalent to the same solution can no longer obtain excellent reception quality for

any of the symbol numbers. Therefore, it is difficult to achieve the ability to correct errors, even if error correction codes are introduced. Accordingly, for q not to have the same solution, the following condition is necessary from Equations 58–61 when focusing on one of two solutions of q which does not include δ .

5 [0231]

Math 62

Condition #1

$$e^{j(\theta_{11}(4i+x)-\theta_{21}(4i+x))} \neq e^{j(\theta_{11}(4i+y)-\theta_{21}(4i+y))} \quad \text{for } \forall x, \forall y (x \neq y; x, y = 0,1,2,3)$$

10 [0232]

(x is 0, 1, 2, 3; y is 0, 1, 2, 3; and $x \neq y$.)

In an example fulfilling Condition #1, values are set as follows:

(Example #1)

(1) $\theta_{11}(4i) = \theta_{11}(4i + 1) = \theta_{11}(4i + 2) = \theta_{11}(4i + 3) = 0$ radians,

15 (2) $\theta_{21}(4i) = 0$ radians,

(3) $\theta_{21}(4i + 1) = \pi/2$ radians,

(4) $\theta_{21}(4i + 2) = \pi$ radians, and

(5) $\theta_{21}(4i + 3) = 3\pi/2$ radians.

(The above is an example. It suffices for one each of zero radians, $\pi/2$ radians, π radians, and $3\pi/2$ radians to exist for the set $(\theta_{21}(4i), \theta_{21}(4i + 1), \theta_{21}(4i + 2), \theta_{21}(4i + 3))$.) In this case, in particular under condition (1), there is no need to perform signal processing (rotation processing) on the baseband signal $S1(t)$, which therefore offers the advantage of a reduction in circuit size. Another example is to set values as follows.

25 (Example #2)

(6) $\theta_{11}(4i) = 0$ radians,

(7) $\theta_{11}(4i + 1) = \pi/2$ radians,

(8) $\theta_{11}(4i + 2) = \pi$ radians,

(9) $\theta_{11}(4i + 3) = 3\pi/2$ radians, and

(10) $\theta_{21}(4i) = \theta_{21}(4i + 1) = \theta_{21}(4i + 2) = \theta_{21}(4i + 3) = 0$ radians.

(The above is an example. It suffices for one each of zero radians, $\pi/2$ radians, π radians, and $3\pi/2$ radians to exist for the set $(\theta_{11}(4i), \theta_{11}(4i + 1), \theta_{11}(4i + 2), \theta_{11}(4i + 3))$.) In this case, in particular under condition (6), there is no need to perform signal processing (rotation processing) on the baseband signal $S2(t)$, which therefore offers the advantage of a reduction in circuit size. Yet another example is as follows.

(Example #3)

10 (11) $\theta_{11}(4i) = \theta_{11}(4i + 1) = \theta_{11}(4i + 2) = \theta_{11}(4i + 3) = 0$ radians,

(12) $\theta_{21}(4i) = 0$ radians,

(13) $\theta_{21}(4i + 1) = \pi/4$ radians,

(14) $\theta_{21}(4i + 2) = \pi/2$ radians, and

(15) $\theta_{21}(4i + 3) = 3\pi/4$ radians.

15 (The above is an example. It suffices for one each of zero radians, $\pi/4$ radians, $\pi/2$ radians, and $3\pi/4$ radians to exist for the set $(\theta_{21}(4i), \theta_{21}(4i + 1), \theta_{21}(4i + 2), \theta_{21}(4i + 3))$.)

(Example #4)

(16) $\theta_{11}(4i) = 0$ radians,

20 (17) $\theta_{11}(4i + 1) = \pi/4$ radians,

(18) $\theta_{11}(4i + 2) = \pi/2$ radians,

(19) $\theta_{11}(4i + 3) = 3\pi/4$ radians, and

(20) $\theta_{21}(4i) = \theta_{21}(4i + 1) = \theta_{21}(4i + 2) = \theta_{21}(4i + 3) = 0$ radians.

(The above is an example. It suffices for one each of zero radians, $\pi/4$ radians, $\pi/2$ radians, and $3\pi/4$ radians to exist for the set $(\theta_{11}(4i), \theta_{11}(4i + 1), \theta_{11}(4i + 2), \theta_{11}(4i + 3))$.)

While four examples have been shown, the method of satisfying Condition #1 is not limited to these examples.

[0233]

Next, design requirements for not only θ_{11} and θ_{12} , but also for λ and δ are described. It suffices to set λ to a certain value; it is then necessary to establish requirements for δ . The following describes the design method for δ when λ is set to
 5 zero radians.

[0234]

In this case, by defining δ so that $\pi/2$ radians $\leq |\delta| \leq \pi$ radians, excellent reception quality is achieved, particularly in an LOS environment.

[0235]

10 Incidentally, for each of the symbol numbers $4i$, $4i + 1$, $4i + 2$, and $4i + 3$, two points q exist where reception quality becomes poor. Therefore, a total of $2 \times 4 = 8$ such points exist. In an LOS environment, in order to prevent reception quality from degrading in a specific reception terminal, these eight points should each have a different solution. In this case, in addition to Condition #1, Condition #2 is
 15 necessary.

[0236]

Math 63

Condition #2

$$e^{j(\theta_{11}(4i+x)-\theta_{21}(4i+x))} \neq e^{j(\theta_{11}(4i+y)-\theta_{21}(4i+y)-\delta)} \quad \text{for } \forall x, \forall y (x, y = 0,1,2,3)$$

20 and

$$e^{j(\theta_{11}(4i+x)-\theta_{21}(4i+x)-\delta)} \neq e^{j(\theta_{11}(4i+y)-\theta_{21}(4i+y)-\delta)} \quad \text{for } \forall x, \forall y (x \neq y; x, y = 0,1,2,3)$$

[0237]

Additionally, the phase of these eight points should be evenly distributed
 25 (since the phase of a direct wave is considered to have a high probability of even distribution). The following describes the design method for δ to satisfy this requirement.

[0238]

In the case of example #1 and example #2, the phase becomes even at the points at which reception quality is poor by setting δ to $\pm 3\pi/4$ radians. For example, letting δ be $3\pi/4$ radians in example #1 (and letting A be a positive real number), then each of the four slots, points at which reception quality becomes poor exist
5 once, as shown in Fig. 20. In the case of example #3 and example #4, the phase becomes even at the points at which reception quality is poor by setting δ to $\pm \pi$ radians. For example, letting δ be π radians in example #3, then in each of the four slots, points at which reception quality becomes poor exist once, as shown in Fig. 21. (If the element q in the channel matrix H exists at the points shown in Figs. 20 and
10 21, reception quality degrades.)

With the above structure, excellent reception quality is achieved in an LOS environment. Above, an example of changing precoding weights in a four-slot period (cycle) is described, but below, changing precoding weights in an N -slot period (cycle) is described. Making the same considerations as in Embodiment 1 and
15 in the above description, processing represented as below is performed on each symbol number.

For symbol number Ni (where i is an integer greater than or equal to zero):

[0239]

Math 64

20 Equation 62

$$\begin{pmatrix} z1(Ni) \\ z2(Ni) \end{pmatrix} = \frac{1}{\sqrt{2}} \begin{pmatrix} e^{j\theta_{11}(Ni)} & e^{j(\theta_{11}(Ni)+\lambda)} \\ e^{j\theta_{21}(Ni)} & e^{j(\theta_{21}(Ni)+\lambda+\delta)} \end{pmatrix} \begin{pmatrix} s1(Ni) \\ s2(Ni) \end{pmatrix}$$

[0240]

Here, j is an imaginary unit.

25 For symbol number $Ni + 1$:

[0241]

Math 65

Equation 63

$$\begin{pmatrix} z1(Ni+1) \\ z2(Ni+1) \end{pmatrix} = \frac{1}{\sqrt{2}} \begin{pmatrix} e^{j\theta_{11}(Ni+1)} & e^{j(\theta_{11}(Ni+1)+\lambda)} \\ e^{j\theta_{21}(Ni+1)} & e^{j(\theta_{21}(Ni+1)+\lambda+\delta)} \end{pmatrix} \begin{pmatrix} s1(Ni+1) \\ s2(Ni+1) \end{pmatrix}$$

5 [0242]

When generalized, this equation is as follows.

For symbol number $Ni + k$ ($k = 0, 1, \dots, N - 1$):

[0243]

Math 66

10 Equation 64

$$\begin{pmatrix} z1(Ni+k) \\ z2(Ni+k) \end{pmatrix} = \frac{1}{\sqrt{2}} \begin{pmatrix} e^{j\theta_{11}(Ni+k)} & e^{j(\theta_{11}(Ni+k)+\lambda)} \\ e^{j\theta_{21}(Ni+k)} & e^{j(\theta_{21}(Ni+k)+\lambda+\delta)} \end{pmatrix} \begin{pmatrix} s1(Ni+k) \\ s2(Ni+k) \end{pmatrix}$$

[0244]

Furthermore, for symbol number $Ni + N - 1$:

15 [0245]

Math 67

Equation 65

$$\begin{pmatrix} z1(Ni+N-1) \\ z2(Ni+N-1) \end{pmatrix} = \frac{1}{\sqrt{2}} \begin{pmatrix} e^{j\theta_{11}(Ni+N-1)} & e^{j(\theta_{11}(Ni+N-1)+\lambda)} \\ e^{j\theta_{21}(Ni+N-1)} & e^{j(\theta_{21}(Ni+N-1)+\lambda+\delta)} \end{pmatrix} \begin{pmatrix} s1(Ni+N-1) \\ s2(Ni+N-1) \end{pmatrix}$$

20 [0246]

Accordingly, $r1$ and $r2$ are represented as follows.

For symbol number Ni (where i is an integer greater than or equal to zero):

[0247]

Math 68

Equation 66

$$\begin{pmatrix} r1(Ni) \\ r2(Ni) \end{pmatrix} = \frac{1}{\sqrt{2}} \begin{pmatrix} h_{11}(Ni) & h_{12}(Ni) \\ h_{21}(Ni) & h_{22}(Ni) \end{pmatrix} \begin{pmatrix} e^{j\theta_{11}(Ni)} & e^{j(\theta_{11}(Ni)+\lambda)} \\ e^{j\theta_{21}(Ni)} & e^{j(\theta_{21}(Ni)+\lambda+\delta)} \end{pmatrix} \begin{pmatrix} s1(Ni) \\ s2(Ni) \end{pmatrix}$$

5

[0248]

Here, j is an imaginary unit.

For symbol number Ni + 1:

[0249]

10 Math 69

Equation 67

$$\begin{pmatrix} r1(Ni+1) \\ r2(Ni+1) \end{pmatrix} = \frac{1}{\sqrt{2}} \begin{pmatrix} h_{11}(Ni+1) & h_{12}(Ni+1) \\ h_{21}(Ni+1) & h_{22}(Ni+1) \end{pmatrix} \begin{pmatrix} e^{j\theta_{11}(Ni+1)} & e^{j(\theta_{11}(Ni+1)+\lambda)} \\ e^{j\theta_{21}(Ni+1)} & e^{j(\theta_{21}(Ni+1)+\lambda+\delta)} \end{pmatrix} \begin{pmatrix} s1(Ni+1) \\ s2(Ni+1) \end{pmatrix}$$

[0250]

15 When generalized, this equation is as follows.

For symbol number Ni + k (k = 0, 1, ..., N - 1):

[0251]

Math 70

Equation 68

$$20 \quad \begin{pmatrix} r1(Ni+k) \\ r2(Ni+k) \end{pmatrix} = \frac{1}{\sqrt{2}} \begin{pmatrix} h_{11}(Ni+k) & h_{12}(Ni+k) \\ h_{21}(Ni+k) & h_{22}(Ni+k) \end{pmatrix} \begin{pmatrix} e^{j\theta_{11}(Ni+k)} & e^{j(\theta_{11}(Ni+k)+\lambda)} \\ e^{j\theta_{21}(Ni+k)} & e^{j(\theta_{21}(Ni+k)+\lambda+\delta)} \end{pmatrix} \begin{pmatrix} s1(Ni+k) \\ s2(Ni+k) \end{pmatrix}$$

[0252]

Furthermore, for symbol number Ni + N - 1:

[0253]

Math 71

Equation 69

$$\begin{pmatrix} r1(Ni+N-1) \\ r2(Ni+N-1) \end{pmatrix} = \frac{1}{\sqrt{2}} \begin{pmatrix} h_{11}(Ni+N-1) & h_{12}(Ni+N-1) \\ h_{21}(Ni+N-1) & h_{22}(Ni+N-1) \end{pmatrix} \begin{pmatrix} e^{j\theta_{11}(Ni+N-1)} & e^{j(\theta_{11}(Ni+N-1)+\lambda)} \\ e^{j\theta_{21}(Ni+N-1)} & e^{j(\theta_{21}(Ni+N-1)+\lambda+\delta)} \end{pmatrix} \begin{pmatrix} s1(Ni+N-1) \\ s2(Ni+N-1) \end{pmatrix}$$

5 [0254]

In this case, it is assumed that only components of direct waves exist in the channel elements $h_{11}(t)$, $h_{12}(t)$, $h_{21}(t)$, and $h_{22}(t)$, that the amplitude components of the direct waves are all equal, and that fluctuations do not occur over time. With these assumptions, Equations 66–69 can be represented as follows.

10 For symbol number Ni (where i is an integer greater than or equal to zero):

[0255]

Math 72

Equation 70

$$\begin{pmatrix} r1(Ni) \\ r2(Ni) \end{pmatrix} = \frac{1}{\sqrt{2}} \begin{pmatrix} A e^{j0} & q \\ A e^{j0} & q \end{pmatrix} \begin{pmatrix} e^{j\theta_{11}(Ni)} & e^{j(\theta_{11}(Ni)+\lambda)} \\ e^{j\theta_{21}(Ni)} & e^{j(\theta_{21}(Ni)+\lambda+\delta)} \end{pmatrix} \begin{pmatrix} s1(Ni) \\ s2(Ni) \end{pmatrix}$$

15

[0256]

Here, j is an imaginary unit.

For symbol number $Ni + 1$:

[0257]

20 Math 73

Equation 71

$$\begin{pmatrix} r1(Ni+1) \\ r2(Ni+1) \end{pmatrix} = \frac{1}{\sqrt{2}} \begin{pmatrix} A e^{j0} & q \\ A e^{j0} & q \end{pmatrix} \begin{pmatrix} e^{j\theta_{11}(Ni+1)} & e^{j(\theta_{11}(Ni+1)+\lambda)} \\ e^{j\theta_{21}(Ni+1)} & e^{j(\theta_{21}(Ni+1)+\lambda+\delta)} \end{pmatrix} \begin{pmatrix} s1(Ni+1) \\ s2(Ni+1) \end{pmatrix}$$

[0258]

When generalized, this equation is as follows.

For symbol number $Ni + k$ ($k = 0, 1, \dots, N - 1$):

[0259]

5 Math 74

Equation 72

$$\begin{pmatrix} r1(Ni + k) \\ r2(Ni + k) \end{pmatrix} = \frac{1}{\sqrt{2}} \begin{pmatrix} A e^{j0} & q \\ A e^{j0} & q \end{pmatrix} \begin{pmatrix} e^{j\theta_{11}(Ni+k)} & e^{j(\theta_{11}(Ni+k)+\lambda)} \\ e^{j\theta_{21}(Ni+k)} & e^{j(\theta_{21}(Ni+k)+\lambda+\delta)} \end{pmatrix} \begin{pmatrix} s1(Ni + k) \\ s2(Ni + k) \end{pmatrix}$$

[0260]

10 Furthermore, for symbol number $Ni + N - 1$:

[0261]

Math 75

Equation 73

$$\begin{pmatrix} r1(Ni + N - 1) \\ r2(Ni + N - 1) \end{pmatrix} = \frac{1}{\sqrt{2}} \begin{pmatrix} A e^{j0} & q \\ A e^{j0} & q \end{pmatrix} \begin{pmatrix} e^{j\theta_{11}(Ni+N-1)} & e^{j(\theta_{11}(Ni+N-1)+\lambda)} \\ e^{j\theta_{21}(Ni+N-1)} & e^{j(\theta_{21}(Ni+N-1)+\lambda+\delta)} \end{pmatrix} \begin{pmatrix} s1(Ni + N - 1) \\ s2(Ni + N - 1) \end{pmatrix}$$

15

[0262]

In Equations 70–73, let A be a real number and q be a complex number. The values of A and q are determined in accordance with the positional relationship between the transmission device and the reception device. Equations 70–73 can be

20 represented as follows.

For symbol number Ni (where i is an integer greater than or equal to zero):

[0263]

Math 76

Equation 74

$$\begin{pmatrix} r1(Ni) \\ r2(Ni) \end{pmatrix} = \frac{1}{\sqrt{2}} \begin{pmatrix} e^{j0} \\ e^{j0} \end{pmatrix} \left(A e^{j0} \quad q \right) \begin{pmatrix} e^{j\theta_{11}(Ni)} & e^{j(\theta_{11}(Ni)+\lambda)} \\ e^{j\theta_{21}(Ni)} & e^{j(\theta_{21}(Ni)+\lambda+\delta)} \end{pmatrix} \begin{pmatrix} s1(Ni) \\ s2(Ni) \end{pmatrix}$$

[0264]

Here, j is an imaginary unit.

5 For symbol number Ni + 1:

[0265]

Math 77

Equation 75

$$\begin{pmatrix} r1(Ni+1) \\ r2(Ni+1) \end{pmatrix} = \frac{1}{\sqrt{2}} \begin{pmatrix} e^{j0} \\ e^{j0} \end{pmatrix} \left(A e^{j0} \quad q \right) \begin{pmatrix} e^{j\theta_{11}(Ni+1)} & e^{j(\theta_{11}(Ni+1)+\lambda)} \\ e^{j\theta_{21}(Ni+1)} & e^{j(\theta_{21}(Ni+1)+\lambda+\delta)} \end{pmatrix} \begin{pmatrix} s1(Ni+1) \\ s2(Ni+1) \end{pmatrix}$$

10

[0266]

When generalized, this equation is as follows.

For symbol number Ni + k (k = 0, 1, ..., N - 1):

[0267]

15 Math 78

Equation 76

$$\begin{pmatrix} r1(Ni+k) \\ r2(Ni+k) \end{pmatrix} = \frac{1}{\sqrt{2}} \begin{pmatrix} e^{j0} \\ e^{j0} \end{pmatrix} \left(A e^{j0} \quad q \right) \begin{pmatrix} e^{j\theta_{11}(Ni+k)} & e^{j(\theta_{11}(Ni+k)+\lambda)} \\ e^{j\theta_{21}(Ni+k)} & e^{j(\theta_{21}(Ni+k)+\lambda+\delta)} \end{pmatrix} \begin{pmatrix} s1(Ni+k) \\ s2(Ni+k) \end{pmatrix}$$

[0268]

20 Furthermore, for symbol number Ni + N - 1:

[0269]

Math 79

Equation 77

$$\begin{pmatrix} r1(Ni + N - 1) \\ r2(Ni + N - 1) \end{pmatrix} = \frac{1}{\sqrt{2}} \begin{pmatrix} e^{j0} \\ e^{j0} \end{pmatrix} \begin{pmatrix} A e^{j0} & q \end{pmatrix} \begin{pmatrix} e^{j\theta_{11}(Ni+N-1)} & e^{j(\theta_{11}(Ni+N-1)+\lambda)} \\ e^{j\theta_{21}(Ni+N-1)} & e^{j(\theta_{21}(Ni+N-1)+\lambda+\delta)} \end{pmatrix} \begin{pmatrix} s1(Ni + N - 1) \\ s2(Ni + N - 1) \end{pmatrix}$$

[0270]

- 5 As a result, when q is represented as follows, a signal component based on one of s1 and s2 is no longer included in r1 and r2, and therefore one of the signals s1 and s2 can no longer be obtained.

For symbol number Ni (where i is an integer greater than or equal to zero):

[0271]

10 Math 80

Equation 78

$$q = -A e^{j(\theta_{11}(Ni) - \theta_{21}(Ni))}, -A e^{j(\theta_{11}(Ni) - \theta_{21}(Ni) - \delta)}$$

[0272]

- 15 For symbol number Ni + 1:

[0273]

Math 81

Equation 79

$$q = -A e^{j(\theta_{11}(Ni+1) - \theta_{21}(Ni+1))}, -A e^{j(\theta_{11}(Ni+1) - \theta_{21}(Ni+1) - \delta)}$$

20

[0274]

When generalized, this equation is as follows.

For symbol number Ni + k (k = 0, 1, ..., N - 1):

[0275]

25 Math 82

Equation 80

$$q = -A e^{j(\theta_{11}(Ni+k)-\theta_{21}(Ni+k))}, -A e^{j(\theta_{11}(Ni+k)-\theta_{21}(Ni+k)-\delta)}$$

[0276]

5 Furthermore, for symbol number $Ni + N - 1$:

[0277]

Math 83

Equation 81

$$q = -A e^{j(\theta_{11}(Ni+N-1)-\theta_{21}(Ni+N-1))}, -A e^{j(\theta_{11}(Ni+N-1)-\theta_{21}(Ni+N-1)-\delta)}$$

10

[0278]

In this case, if q has the same solution in symbol numbers Ni through $Ni + N - 1$, then since the channel elements of the direct waves do not greatly fluctuate, a reception device having channel elements in which the value of q is equivalent to this same solution can no longer obtain excellent reception quality for any of the symbol numbers. Therefore, it is difficult to achieve the ability to correct errors, even if error correction codes are introduced. Accordingly, for q not to have the same solution, the following condition is necessary from Equations 78–81 when focusing on one of two solutions of q which does not include δ .

20 [0279]

Math 84

Condition #3

$$e^{j(\theta_{11}(Ni+x)-\theta_{21}(Ni+x))} \neq e^{j(\theta_{11}(Ni+y)-\theta_{21}(Ni+y))} \quad \text{for } \forall x, \forall y (x \neq y; x, y = 0, 1, 2, \dots, N-2, N-1)$$

25 [0280]

(x is 0, 1, 2, ..., $N - 2$, $N - 1$; y is 0, 1, 2, ..., $N - 2$, $N - 1$; and $x \neq y$.)

Next, design requirements for not only θ_{11} and θ_{12} , but also for λ and δ are described. It suffices to set λ to a certain value; it is then necessary to establish requirements for δ . The following describes the design method for δ when λ is set to zero radians.

5 [0281]

In this case, similar to the method of changing the precoding weights in a four-slot period (cycle), by defining δ so that $\pi/2$ radians $\leq |\delta| \leq \pi$ radians, excellent reception quality is achieved, particularly in an LOS environment.

[0282]

10 In each symbol number N_i through $N_i + N - 1$, two points labeled q exist where reception quality becomes poor, and therefore $2N$ such points exist. In an LOS environment, in order to achieve excellent characteristics, these $2N$ points should each have a different solution. In this case, in addition to Condition #3, Condition #4 is necessary.

15 [0283]

Math 85

Condition #4

$$e^{j(\theta_{11}(N_i+x)-\theta_{21}(N_i+x))} \neq e^{j(\theta_{11}(N_i+y)-\theta_{21}(N_i+y)-\delta)} \quad \text{for } \forall x, \forall y (x, y = 0, 1, 2, \dots, N-2, N-1)$$

and

20
$$e^{j(\theta_{11}(N_i+x)-\theta_{21}(N_i-x)-\delta)} \neq e^{j(\theta_{11}(N_i+y)-\theta_{21}(N_i+y)-\delta)} \quad \text{for } \forall x, \forall y (x \neq y; x, y = 0, 1, 2, \dots, N-2, N-1)$$

[0284]

25 Additionally, the phase of these $2N$ points should be evenly distributed (since the phase of a direct wave at each reception device is considered to have a high probability of even distribution).

As described above, when a transmission device transmits a plurality of modulated signals from a plurality of antennas in a MIMO system, the advantageous effect of improved transmission quality, as compared to conventional spatial

multiplexing MIMO, is achieved in an LOS environment in which direct waves dominate by hopping between precoding weights regularly over time.

[0285]

In the present embodiment, the structure of the reception device is as described in Embodiment 1, and in particular with regards to the structure of the reception device, operations have been described for a limited number of antennas, but the present invention may be embodied in the same way even if the number of antennas increases. In other words, the number of antennas in the reception device does not affect the operations or advantageous effects of the present embodiment. Furthermore, in the present embodiment, similar to Embodiment 1, the error correction codes are not limited.

[0286]

In the present embodiment, in contrast with Embodiment 1, the method of changing the precoding weights in the time domain has been described. As described in Embodiment 1, however, the present invention may be similarly embodied by changing the precoding weights by using a multi-carrier transmission method and arranging symbols in the frequency domain and the frequency-time domain. Furthermore, in the present embodiment, symbols other than data symbols, such as pilot symbols (preamble, unique word, and the like), symbols for control information, and the like, may be arranged in the frame in any way.

[0287]

(Embodiment 3)

In Embodiment 1 and Embodiment 2, the method of regularly hopping between precoding weights has been described for the case where the amplitude of each element in the precoding weight matrix is equivalent. In the present embodiment, however, an example that does not satisfy this condition is described.

[0288]

For the sake of contrast with Embodiment 2, the case of changing precoding weights over an N-slot period (cycle) is described. Making the same considerations as in Embodiment 1 and Embodiment 2, processing represented as below is performed on each symbol number. Let β be a positive real number, and $\beta \neq 1$.

5 For symbol number Ni (where i is an integer greater than or equal to zero):

[0289]

Math 86

Equation 82

$$\begin{pmatrix} z1(Ni) \\ z2(Ni) \end{pmatrix} = \frac{1}{\sqrt{\beta^2 + 1}} \begin{pmatrix} e^{j\theta_{11}(Ni)} & \beta \times e^{j(\theta_{11}(Ni) + \lambda)} \\ \beta \times e^{j\theta_{21}(Ni)} & e^{j(\theta_{21}(Ni) + \lambda + \delta)} \end{pmatrix} \begin{pmatrix} s1(Ni) \\ s2(Ni) \end{pmatrix}$$

10

[0290]

Here, j is an imaginary unit.

For symbol number $Ni + 1$:

[0291]

15 Math 87

Equation 83

$$\begin{pmatrix} z1(Ni + 1) \\ z2(Ni + 1) \end{pmatrix} = \frac{1}{\sqrt{\beta^2 + 1}} \begin{pmatrix} e^{j\theta_{11}(Ni+1)} & \beta \times e^{j(\theta_{11}(Ni+1) + \lambda)} \\ \beta \times e^{j\theta_{21}(Ni+1)} & e^{j(\theta_{21}(Ni+1) + \lambda + \delta)} \end{pmatrix} \begin{pmatrix} s1(Ni + 1) \\ s2(Ni + 1) \end{pmatrix}$$

[0292]

20 When generalized, this equation is as follows.

For symbol number $Ni + k$ ($k = 0, 1, \dots, N - 1$):

[0293]

Math 88

Equation 84

$$\begin{pmatrix} z1(Ni + k) \\ z2(Ni + k) \end{pmatrix} = \frac{1}{\sqrt{\beta^2 + 1}} \begin{pmatrix} e^{j\theta_{11}(Ni+k)} & \beta \times e^{j(\theta_{11}(Ni+k)+\lambda)} \\ \beta \times e^{j\theta_{21}(Ni+k)} & e^{j(\theta_{21}(Ni+k)+\lambda+\delta)} \end{pmatrix} \begin{pmatrix} s1(Ni + k) \\ s2(Ni + k) \end{pmatrix}$$

[0294]

5 Furthermore, for symbol number $Ni + N - 1$:

[0295]

Math 89

Equation 85

$$\begin{pmatrix} z1(Ni + N - 1) \\ z2(Ni + N - 1) \end{pmatrix} = \frac{1}{\sqrt{\beta^2 + 1}} \begin{pmatrix} e^{j\theta_{11}(Ni+N-1)} & \beta \times e^{j(\theta_{11}(Ni+N-1)+\lambda)} \\ \beta \times e^{j\theta_{21}(Ni+N-1)} & e^{j(\theta_{21}(Ni+N-1)+\lambda+\delta)} \end{pmatrix} \begin{pmatrix} s1(Ni + N - 1) \\ s2(Ni + N - 1) \end{pmatrix}$$

10

[0296]

Accordingly, r1 and r2 are represented as follows.

For symbol number Ni (where i is an integer greater than or equal to zero):

[0297]

15 Math 90

Equation 86

$$\begin{pmatrix} r1(Ni) \\ r2(Ni) \end{pmatrix} = \frac{1}{\sqrt{\beta^2 + 1}} \begin{pmatrix} h_{11}(Ni) & h_{12}(Ni) \\ h_{21}(Ni) & h_{22}(Ni) \end{pmatrix} \begin{pmatrix} e^{j\theta_{11}(Ni)} & \beta \times e^{j(\theta_{11}(Ni)+\lambda)} \\ \beta \times e^{j\theta_{21}(Ni)} & e^{j(\theta_{21}(Ni)+\lambda+\delta)} \end{pmatrix} \begin{pmatrix} s1(Ni) \\ s2(Ni) \end{pmatrix}$$

[0298]

20 Here, j is an imaginary unit.

For symbol number $Ni + 1$:

[0299]

Math 91

Equation 87

$$\begin{pmatrix} r1(Ni+1) \\ r2(Ni+1) \end{pmatrix} = \frac{1}{\sqrt{\beta^2 + 1}} \begin{pmatrix} h_{11}(Ni+1) & h_{12}(Ni+1) \\ h_{21}(Ni+1) & h_{22}(Ni+1) \end{pmatrix} \begin{pmatrix} e^{j\theta_{11}(Ni+1)} & \beta \times e^{j(\theta_{11}(Ni+1)+\lambda)} \\ \beta \times e^{j\theta_{21}(Ni+1)} & e^{j(\theta_{21}(Ni+1)+\lambda+\delta)} \end{pmatrix} \begin{pmatrix} s1(Ni+1) \\ s2(Ni+1) \end{pmatrix}$$

5 [0300]

When generalized, this equation is as follows.

For symbol number $Ni + k$ ($k = 0, 1, \dots, N - 1$):

[0301]

Math 92

10 Equation 88

$$\begin{pmatrix} r1(Ni+k) \\ r2(Ni+k) \end{pmatrix} = \frac{1}{\sqrt{\beta^2 + 1}} \begin{pmatrix} h_{11}(Ni+k) & h_{12}(Ni+k) \\ h_{21}(Ni+k) & h_{22}(Ni+k) \end{pmatrix} \begin{pmatrix} e^{j\theta_{11}(Ni+k)} & \beta \times e^{j(\theta_{11}(Ni+k)+\lambda)} \\ \beta \times e^{j\theta_{21}(Ni+k)} & e^{j(\theta_{21}(Ni+k)+\lambda+\delta)} \end{pmatrix} \begin{pmatrix} s1(Ni+k) \\ s2(Ni+k) \end{pmatrix}$$

[0302]

When generalized, this equation is as follows.

15 For symbol number $Ni - N - 1$:

[0303]

Math 93

Equation 89

$$\begin{pmatrix} r1(Ni+N-1) \\ r2(Ni+N-1) \end{pmatrix} = \frac{1}{\sqrt{\beta^2 + 1}} \begin{pmatrix} h_{11}(Ni+N-1) & h_{12}(Ni+N-1) \\ h_{21}(Ni+N-1) & h_{22}(Ni+N-1) \end{pmatrix} \begin{pmatrix} e^{j\theta_{11}(Ni+N-1)} & \beta \times e^{j(\theta_{11}(Ni+N-1)+\lambda)} \\ \beta \times e^{j\theta_{21}(Ni+N-1)} & e^{j(\theta_{21}(Ni+N-1)+\lambda+\delta)} \end{pmatrix} \begin{pmatrix} s1(Ni+N-1) \\ s2(Ni+N-1) \end{pmatrix}$$

20

[0304]

In this case, it is assumed that only components of direct waves exist in the channel elements $h_{11}(t)$, $h_{12}(t)$, $h_{21}(t)$, and $h_{22}(t)$, that the amplitude components of the direct waves are all equal, and that fluctuations do not occur over time. With these assumptions, Equations 86–89 can be represented as follows.

For symbol number Ni (where i is an integer greater than or equal to zero):

[0305]

Math 94

Equation 90

$$5 \quad \begin{pmatrix} r1(Ni) \\ r2(Ni) \end{pmatrix} = \frac{1}{\sqrt{\beta^2 + 1}} \begin{pmatrix} A e^{j0} & q \\ A e^{j0} & q \end{pmatrix} \begin{pmatrix} e^{j\theta_{11}(Ni)} & \beta \times e^{j(\theta_{11}(Ni) + \lambda)} \\ \beta \times e^{j\theta_{21}(Ni)} & e^{j(\theta_{21}(Ni) + \lambda + \delta)} \end{pmatrix} \begin{pmatrix} s1(Ni) \\ s2(Ni) \end{pmatrix}$$

[0306]

Here, j is an imaginary unit.

For symbol number $Ni + 1$:

10 [0307]

Math 95

Equation 91

$$\begin{pmatrix} r1(Ni + 1) \\ r2(Ni + 1) \end{pmatrix} = \frac{1}{\sqrt{\beta^2 + 1}} \begin{pmatrix} A e^{j0} & q \\ A e^{j0} & q \end{pmatrix} \begin{pmatrix} e^{j\theta_{11}(Ni+1)} & \beta \times e^{j(\theta_{11}(Ni+1) + \lambda)} \\ \beta \times e^{j\theta_{21}(Ni+1)} & e^{j(\theta_{21}(Ni+1) + \lambda + \delta)} \end{pmatrix} \begin{pmatrix} s1(Ni + 1) \\ s2(Ni + 1) \end{pmatrix}$$

15 [0308]

When generalized, this equation is as follows.

For symbol number $Ni + k$ ($k = 0, 1, \dots, N - 1$):

[0309]

Math 96

20 Equation 92

$$\begin{pmatrix} r1(Ni + k) \\ r2(Ni + k) \end{pmatrix} = \frac{1}{\sqrt{\beta^2 + 1}} \begin{pmatrix} A e^{j0} & q \\ A e^{j0} & q \end{pmatrix} \begin{pmatrix} e^{j\theta_{11}(Ni+k)} & \beta \times e^{j(\theta_{11}(Ni+k) + \lambda)} \\ \beta \times e^{j\theta_{21}(Ni+k)} & e^{j(\theta_{21}(Ni+k) + \lambda + \delta)} \end{pmatrix} \begin{pmatrix} s1(Ni + k) \\ s2(Ni + k) \end{pmatrix}$$

[0310]

Furthermore, for symbol number $Ni + N - 1$:

[0311]

Math 97

5 Equation 93

$$\begin{pmatrix} r1(Ni + N - 1) \\ r2(Ni + N - 1) \end{pmatrix} = \frac{1}{\sqrt{\beta^2 + 1}} \begin{pmatrix} A e^{j0} & q \\ A e^{j0} & q \end{pmatrix} \begin{pmatrix} e^{j\theta_{11}(Ni + N - 1)} & \beta \times e^{j(\theta_{11}(Ni + N - 1) + \lambda)} \\ \beta \times e^{j\theta_{21}(Ni + N - 1)} & e^{j(\theta_{21}(Ni + N - 1) + \lambda + \delta)} \end{pmatrix} \begin{pmatrix} s1(Ni + N - 1) \\ s2(Ni + N - 1) \end{pmatrix}$$

[0312]

In Equations 90–93, let A be a real number and q be a complex number.

10 Equations 90–93 can be represented as follows.

For symbol number Ni (where i is an integer greater than or equal to zero):

[0313]

Math 98

Equation 94

$$15 \quad \begin{pmatrix} r1(Ni) \\ r2(Ni) \end{pmatrix} = \frac{1}{\sqrt{\beta^2 + 1}} \begin{pmatrix} e^{j0} \\ e^{j0} \end{pmatrix} \begin{pmatrix} A e^{j0} & q \\ A e^{j0} & q \end{pmatrix} \begin{pmatrix} e^{j\theta_{11}(Ni)} & \beta \times e^{j(\theta_{11}(Ni) + \lambda)} \\ \beta \times e^{j\theta_{21}(Ni)} & e^{j(\theta_{21}(Ni) + \lambda + \delta)} \end{pmatrix} \begin{pmatrix} s1(Ni) \\ s2(Ni) \end{pmatrix}$$

[0314]

Here, j is an imaginary unit.For symbol number $Ni + 1$:

20 [0315]

Math 99

Equation 95

$$\begin{pmatrix} r1(Ni + 1) \\ r2(Ni + 1) \end{pmatrix} = \frac{1}{\sqrt{\beta^2 + 1}} \begin{pmatrix} e^{j0} \\ e^{j0} \end{pmatrix} \begin{pmatrix} A e^{j0} & q \\ A e^{j0} & q \end{pmatrix} \begin{pmatrix} e^{j\theta_{11}(Ni + 1)} & \beta \times e^{j(\theta_{11}(Ni + 1) + \lambda)} \\ \beta \times e^{j\theta_{21}(Ni + 1)} & e^{j(\theta_{21}(Ni + 1) + \lambda + \delta)} \end{pmatrix} \begin{pmatrix} s1(Ni + 1) \\ s2(Ni + 1) \end{pmatrix}$$

[0316]

When generalized, this equation is as follows.

For symbol number $Ni + k$ ($k = 0, 1, \dots, N - 1$):

5 [0317]

Math 100

Equation 96

$$\begin{pmatrix} r1(Ni + k) \\ r2(Ni + k) \end{pmatrix} = \frac{1}{\sqrt{\beta^2 + 1}} \begin{pmatrix} e^{j0} \\ e^{j0} \end{pmatrix} \left(A e^{j0} \quad q \right) \begin{pmatrix} e^{j\theta_{11}(Ni+k)} & \beta \times e^{j(\theta_{11}(Ni+k)+\lambda)} \\ \beta \times e^{j\theta_{21}(Ni+k)} & e^{j(\theta_{21}(Ni+k)+\lambda+\delta)} \end{pmatrix} \begin{pmatrix} s1(Ni + k) \\ s2(Ni + k) \end{pmatrix}$$

10 [0318]

Furthermore, for symbol number $Ni + N - 1$:

[0319]

Math 101

Equation 97

$$15 \quad \begin{pmatrix} r1(Ni + N - 1) \\ r2(Ni + N - 1) \end{pmatrix} = \frac{1}{\sqrt{\beta^2 + 1}} \begin{pmatrix} e^{j0} \\ e^{j0} \end{pmatrix} \left(A e^{j0} \quad q \right) \begin{pmatrix} e^{j\theta_{11}(Ni+N-1)} & \beta \times e^{j(\theta_{11}(Ni+N-1)+\lambda)} \\ \beta \times e^{j\theta_{21}(Ni+N-1)} & e^{j(\theta_{21}(Ni+N-1)+\lambda+\delta)} \end{pmatrix} \begin{pmatrix} s1(Ni + N - 1) \\ s2(Ni + N - 1) \end{pmatrix}$$

[0320]

As a result, when q is represented as follows, one of the signals $s1$ and $s2$ can no longer be obtained.

20 For symbol number Ni (where i is an integer greater than or equal to zero):

[0321]

Math 102

Equation 98

$$q = -\frac{A}{\beta} e^{j(\theta_{11}(Ni)-\theta_{21}(Ni))}, -A\beta e^{j(\theta_{11}(Ni)-\theta_{21}(Ni)-\delta)}$$

[0322]

For symbol number $Ni + 1$:

[0323]

5 Math 103

Equation 99

$$q = -\frac{A}{\beta} e^{j(\theta_{11}(Ni+1)-\theta_{21}(Ni+1))}, -A\beta e^{j(\theta_{11}(Ni+1)-\theta_{21}(Ni+1)-\delta)}$$

[0324]

10 When generalized, this equation is as follows.

For symbol number $Ni + k$ ($k = 0, 1, \dots, N - 1$):

[0325]

Math 104

Equation 100

$$15 \quad q = -\frac{A}{\beta} e^{j(\theta_{11}(Ni+k)-\theta_{21}(Ni+k))}, -A\beta e^{j(\theta_{11}(Ni+k)-\theta_{21}(Ni+k)-\delta)}$$

[0326]

Furthermore, for symbol number $Ni + N - 1$:

[0327]

20 Math 105

Equation 101

$$q = -\frac{A}{\beta} e^{j(\theta_{11}(Ni+N-1)-\theta_{21}(Ni+N-1))}, -A\beta e^{j(\theta_{11}(Ni+N-1)-\theta_{21}(Ni+N-1)-\delta)}$$

[0328]

In this case, if q has the same solution in symbol numbers N_i through $N_i + N - 1$, then since the channel elements of the direct waves do not greatly fluctuate, excellent reception quality can no longer be obtained for any of the symbol numbers. Therefore, it is difficult to achieve the ability to correct errors, even if error correction codes are introduced. Accordingly, for q not to have the same solution, the following condition is necessary from Equations 98–101 when focusing on one of two solutions of q which does not include δ .

[0329]

Math 106

10 Condition #5

$$e^{j(\theta_{11}(N_i+x)-\theta_{21}(N_i+x))} \neq e^{j(\theta_{11}(N_i+y)-\theta_{21}(N_i+y))} \quad \text{for } \forall x, \forall y (x \neq y; x, y = 0, 1, 2, \dots, N-2, N-1)$$

[0330]

(x is 0, 1, 2, ..., $N-2$, $N-1$; y is 0, 1, 2, ..., $N-2$, $N-1$; and $x \neq y$.)

15 Next, design requirements for not only θ_{11} and θ_{12} , but also for λ and δ are described. It suffices to set λ to a certain value; it is then necessary to establish requirements for δ . The following describes the design method for δ when λ is set to zero radians.

[0331]

20 In this case, similar to the method of changing the precoding weights in a four-slot period (cycle), by defining δ so that $\pi/2$ radians $\leq |\delta| \leq \pi$ radians, excellent reception quality is achieved, particularly in an LOS environment.

[0332]

25 In each of symbol numbers N_i through $N_i + N - 1$, two points q exist where reception quality becomes poor, and therefore $2N$ such points exist. In an LOS environment, in order to achieve excellent characteristics, these $2N$ points should each have a different solution. In this case, in addition to Condition #5, considering that β is a positive real number, and $\beta \neq 1$, Condition #6 is necessary.

[0333]

Math 107

Condition #6

$$e^{j(\theta_{11}(N+x)-\theta_{21}(N+x)-\delta)} \neq e^{j(\theta_{11}(N+y)-\theta_{21}(N+y)-\delta)} \quad \text{for } \forall x, \forall y (x \neq y; x, y = 0, 1, 2, \dots, N-2, N-1)$$

5

[0334]

As described above, when a transmission device transmits a plurality of modulated signals from a plurality of antennas in a MIMO system, the advantageous effect of improved transmission quality, as compared to conventional spatial multiplexing MIMO system, is achieved in an LOS environment in which direct waves dominate by hopping between precoding weights regularly over time.

10

[0335]

In the present embodiment, the structure of the reception device is as described in Embodiment 1, and in particular with regards to the structure of the reception device, operations have been described for a limited number of antennas, but the present invention may be embodied in the same way even if the number of antennas increases. In other words, the number of antennas in the reception device does not affect the operations or advantageous effects of the present embodiment. Furthermore, in the present embodiment, similar to Embodiment 1, the error correction codes are not limited.

15

20

[0336]

In the present embodiment, in contrast with Embodiment 1, the method of changing the precoding weights in the time domain has been described. As described in Embodiment 1, however, the present invention may be similarly embodied by changing the precoding weights by using a multi-carrier transmission method and arranging symbols in the frequency domain and the frequency-time domain. Furthermore, in the present embodiment, symbols other than data symbols,

25

such as pilot symbols (preamble, unique word, and the like), symbols for control information, and the like, may be arranged in the frame in any way.

[0337]

(Embodiment 4)

5 In Embodiment 3, the method of regularly hopping between precoding weights has been described for the example of two types of amplitudes for each element in the precoding weight matrix, 1 and β .

[0338]

In this case, the following

10 [0339]

Math 108

$$\frac{1}{\sqrt{\beta^2 + 1}}$$

[0340]

is ignored.

15 [0341]

Next, the example of changing the value of β by slot is described. For the sake of contrast with Embodiment 3, the case of changing precoding weights over a $2 \times N$ -slot period (cycle) is described.

[0342]

20 Making the same considerations as in Embodiment 1, Embodiment 2, and Embodiment 3, processing represented as below is performed on symbol numbers. Let β be a positive real number, and $\beta \neq 1$. Furthermore, let α be a positive real number, and $\alpha \neq \beta$.

For symbol number $2Ni$ (where i is an integer greater than or equal to zero):

25 [0343]

Math 109

Equation 102

$$\begin{pmatrix} z1(2Ni) \\ z2(2Ni) \end{pmatrix} = \frac{1}{\sqrt{\beta^2 + 1}} \begin{pmatrix} e^{j\theta_{11}(2Ni)} & \beta \times e^{j(\theta_{11}(2Ni) + \lambda)} \\ \beta \times e^{j\theta_{21}(2Ni)} & e^{j(\theta_{21}(2Ni) + \lambda + \delta)} \end{pmatrix} \begin{pmatrix} s1(2Ni) \\ s2(2Ni) \end{pmatrix}$$

[0344]

5 Here, j is an imaginary unit.

For symbol number $2Ni + 1$:

[0345]

Math 110

Equation 103

$$10 \quad \begin{pmatrix} z1(2Ni + 1) \\ z2(2Ni + 1) \end{pmatrix} = \frac{1}{\sqrt{\beta^2 + 1}} \begin{pmatrix} e^{j\theta_{11}(2Ni+1)} & \beta \times e^{j(\theta_{11}(2Ni+1) + \lambda)} \\ \beta \times e^{j\theta_{21}(2Ni+1)} & e^{j(\theta_{21}(2Ni+1) + \lambda + \delta)} \end{pmatrix} \begin{pmatrix} s1(2Ni + 1) \\ s2(2Ni + 1) \end{pmatrix}$$

[0346]

When generalized, this equation is as follows.

For symbol number $2Ni + k$ ($k = 0, 1, \dots, N - 1$):

15 [0347]

Math 111

Equation 104

$$\begin{pmatrix} z1(2Ni + k) \\ z2(2Ni + k) \end{pmatrix} = \frac{1}{\sqrt{\beta^2 + 1}} \begin{pmatrix} e^{j\theta_{11}(2Ni+k)} & \beta \times e^{j(\theta_{11}(2Ni+k) + \lambda)} \\ \beta \times e^{j\theta_{21}(2Ni+k)} & e^{j(\theta_{21}(2Ni+k) + \lambda + \delta)} \end{pmatrix} \begin{pmatrix} s1(2Ni + k) \\ s2(2Ni + k) \end{pmatrix}$$

20 [0348]

Furthermore, for symbol number $2Ni + N - 1$:

[0349]

Math 112

Equation 105

$$\begin{pmatrix} z1(2Ni + N - 1) \\ z2(2Ni + N - 1) \end{pmatrix} = \frac{1}{\sqrt{\beta^2 + 1}} \begin{pmatrix} e^{j\theta_{11}(2Ni + N - 1)} & \beta \times e^{j(\theta_{11}(2Ni + N - 1) + \lambda)} \\ \beta \times e^{j\theta_{21}(2Ni + N - 1)} & e^{j(\theta_{21}(2Ni + N - 1) + \lambda + \delta)} \end{pmatrix} \begin{pmatrix} s1(2Ni + N - 1) \\ s2(2Ni + N - 1) \end{pmatrix}$$

5 [0350]

For symbol number $2Ni + N$ (where i is an integer greater than or equal to zero):

[0351]

Math 113

Equation 106

$$10 \quad \begin{pmatrix} z1(2Ni + N) \\ z2(2Ni + N) \end{pmatrix} = \frac{1}{\sqrt{\alpha^2 + 1}} \begin{pmatrix} e^{j\theta_{11}(2Ni + N)} & \alpha \times e^{j(\theta_{11}(2Ni + N) + \lambda)} \\ \alpha \times e^{j\theta_{21}(2Ni + N)} & e^{j(\theta_{21}(2Ni + N) + \lambda + \delta)} \end{pmatrix} \begin{pmatrix} s1(2Ni + N) \\ s2(2Ni + N) \end{pmatrix}$$

[0352]

Here, j is an imaginary unit.

For symbol number $2Ni + N + 1$:

15 [0353]

Math 114

Equation 107

$$\begin{pmatrix} z1(2Ni + N + 1) \\ z2(2Ni + N + 1) \end{pmatrix} = \frac{1}{\sqrt{\alpha^2 + 1}} \begin{pmatrix} e^{j\theta_{11}(2Ni + N + 1)} & \alpha \times e^{j(\theta_{11}(2Ni + N + 1) + \lambda)} \\ \alpha \times e^{j\theta_{21}(2Ni + N + 1)} & e^{j(\theta_{21}(2Ni + N + 1) + \lambda + \delta)} \end{pmatrix} \begin{pmatrix} s1(2Ni + N + 1) \\ s2(2Ni + N + 1) \end{pmatrix}$$

20 [0354]

When generalized, this equation is as follows.

For symbol number $2Ni + N + k$ ($k = 0, 1, \dots, N - 1$):

[0355]

Math 115

Equation 108

$$\begin{pmatrix} z1(2Ni + N + k) \\ z2(2Ni + N + k) \end{pmatrix} = \frac{1}{\sqrt{\alpha^2 + 1}} \begin{pmatrix} e^{j\theta_{11}(2Ni + N + k)} & \alpha \times e^{j(\theta_{11}(2Ni + N + k) + \lambda)} \\ \alpha \times e^{j\theta_{21}(2Ni + N + k)} & e^{j(\theta_{21}(2Ni + N + k) + \lambda + \delta)} \end{pmatrix} \begin{pmatrix} s1(2Ni + N + k) \\ s2(2Ni + N + k) \end{pmatrix}$$

5 [0356]

Furthermore, for symbol number $2Ni + 2N - 1$:

[0357]

Math 116

Equation 109

$$10 \quad \begin{pmatrix} z1(2Ni + 2N - 1) \\ z2(2Ni + 2N - 1) \end{pmatrix} = \frac{1}{\sqrt{\alpha^2 + 1}} \begin{pmatrix} e^{j\theta_{11}(2Ni + 2N - 1)} & \alpha \times e^{j(\theta_{11}(2Ni + 2N - 1) + \lambda)} \\ \alpha \times e^{j\theta_{21}(2Ni + 2N - 1)} & e^{j(\theta_{21}(2Ni + 2N - 1) + \lambda + \delta)} \end{pmatrix} \begin{pmatrix} s1(2Ni + 2N - 1) \\ s2(2Ni + 2N - 1) \end{pmatrix}$$

[0358]

Accordingly, r1 and r2 are represented as follows.

For symbol number $2Ni$ (where i is an integer greater than or equal to zero):

15 [0359]

Math 117

Equation 110

$$\begin{pmatrix} r1(2Ni) \\ r2(2Ni) \end{pmatrix} = \frac{1}{\sqrt{\beta^2 + 1}} \begin{pmatrix} h_{11}(2Ni) & h_{12}(2Ni) \\ h_{21}(2Ni) & h_{22}(2Ni) \end{pmatrix} \begin{pmatrix} e^{j\theta_{11}(2Ni)} & \beta \times e^{j(\theta_{11}(2Ni) + \lambda)} \\ \beta \times e^{j\theta_{21}(2Ni)} & e^{j(\theta_{21}(2Ni) + \lambda + \delta)} \end{pmatrix} \begin{pmatrix} s1(2Ni) \\ s2(2Ni) \end{pmatrix}$$

20 [0360]

Here, j is an imaginary unit.

For symbol number $2Ni + 1$:

[0361]

Math 118

Equation 111

$$\begin{pmatrix} r1(2Ni+1) \\ r2(2Ni+1) \end{pmatrix} = \frac{1}{\sqrt{\beta^2+1}} \begin{pmatrix} h_{11}(2Ni+1) & h_{12}(2Ni+1) \\ h_{21}(2Ni+1) & h_{22}(2Ni+1) \end{pmatrix} \begin{pmatrix} e^{j\theta_{11}(2Ni+1)} & \beta \times e^{j(\theta_{11}(2Ni+1)+\lambda)} \\ \beta \times e^{j\theta_{21}(2Ni+1)} & e^{j(\theta_{21}(2Ni+1)+\lambda+\delta)} \end{pmatrix} \begin{pmatrix} s1(2Ni+1) \\ s2(2Ni+1) \end{pmatrix}$$

[0362]

5 When generalized, this equation is as follows.

For symbol number $2Ni + k$ ($k = 0, 1, \dots, N - 1$):

[0363]

Math 119

Equation 112

$$10 \quad \begin{pmatrix} r1(2Ni+k) \\ r2(2Ni+k) \end{pmatrix} = \frac{1}{\sqrt{\beta^2+1}} \begin{pmatrix} h_{11}(2Ni+k) & h_{12}(2Ni+k) \\ h_{21}(2Ni+k) & h_{22}(2Ni+k) \end{pmatrix} \begin{pmatrix} e^{j\theta_{11}(2Ni+k)} & \beta \times e^{j(\theta_{11}(2Ni+k)+\lambda)} \\ \beta \times e^{j\theta_{21}(2Ni+k)} & e^{j(\theta_{21}(2Ni+k)+\lambda+\delta)} \end{pmatrix} \begin{pmatrix} s1(2Ni+k) \\ s2(2Ni+k) \end{pmatrix}$$

[0364]

Furthermore, for symbol number $2Ni + N - 1$:

[0365]

15 Math 120

Equation 113

$$\begin{pmatrix} r1(2Ni+N-1) \\ r2(2Ni+N-1) \end{pmatrix} = \frac{1}{\sqrt{\beta^2+1}} \begin{pmatrix} h_{11}(2Ni+N-1) & h_{12}(2Ni+N-1) \\ h_{21}(2Ni+N-1) & h_{22}(2Ni+N-1) \end{pmatrix} \begin{pmatrix} e^{j\theta_{11}(2Ni+N-1)} & \beta \times e^{j(\theta_{11}(2Ni+N-1)+\lambda)} \\ \beta \times e^{j\theta_{21}(2Ni+N-1)} & e^{j(\theta_{21}(2Ni+N-1)+\lambda+\delta)} \end{pmatrix} \begin{pmatrix} s1(2Ni+N-1) \\ s2(2Ni+N-1) \end{pmatrix}$$

[0366]

20 For symbol number $2Ni + N$ (where i is an integer greater than or equal to zero):

[0367]

Math 121

Equation 114

$$\begin{pmatrix} r1(2Ni+N) \\ r2(2Ni+N) \end{pmatrix} = \frac{1}{\sqrt{\alpha^2+1}} \begin{pmatrix} h_{11}(2Ni+N) & h_{12}(2Ni+N) \\ h_{21}(2Ni+N) & h_{22}(2Ni+N) \end{pmatrix} \begin{pmatrix} e^{j\theta_{11}(2Ni+N)} & \alpha \times e^{j(\theta_{11}(2Ni+N)+\lambda)} \\ \alpha \times e^{j\theta_{21}(2Ni+N)} & e^{j(\theta_{21}(2Ni+N)+\lambda+\delta)} \end{pmatrix} \begin{pmatrix} s1(2Ni+N) \\ s2(2Ni+N) \end{pmatrix}$$

[0368]

Here, j is an imaginary unit.

For symbol number $2Ni + N + 1$:

5 [0369]

Math 122

Equation 115

$$\begin{pmatrix} r1(2Ni+N+1) \\ r2(2Ni+N+1) \end{pmatrix} = \frac{1}{\sqrt{\alpha^2+1}} \begin{pmatrix} h_{11}(2Ni+N+1) & h_{12}(2Ni+N+1) \\ h_{21}(2Ni+N+1) & h_{22}(2Ni+N+1) \end{pmatrix} \begin{pmatrix} e^{j\theta_{11}(2Ni+N+1)} & \alpha \times e^{j(\theta_{11}(2Ni+N+1)+\lambda)} \\ \alpha \times e^{j\theta_{21}(2Ni+N+1)} & e^{j(\theta_{21}(2Ni+N+1)+\lambda+\delta)} \end{pmatrix} \begin{pmatrix} s1(2Ni+N+1) \\ s2(2Ni+N+1) \end{pmatrix}$$

10 [0370]

When generalized, this equation is as follows.

For symbol number $2Ni + N + k$ ($k = 0, 1, \dots, N - 1$):

[0371]

Math 123

15 Equation 116

$$\begin{pmatrix} r1(2Ni+N+k) \\ r2(2Ni+N+k) \end{pmatrix} = \frac{1}{\sqrt{\alpha^2+1}} \begin{pmatrix} h_{11}(2Ni+N+k) & h_{12}(2Ni+N+k) \\ h_{21}(2Ni+N+k) & h_{22}(2Ni+N+k) \end{pmatrix} \begin{pmatrix} e^{j\theta_{11}(2Ni+N+k)} & \alpha \times e^{j(\theta_{11}(2Ni+N+k)+\lambda)} \\ \alpha \times e^{j\theta_{21}(2Ni+N+k)} & e^{j(\theta_{21}(2Ni+N+k)+\lambda+\delta)} \end{pmatrix} \begin{pmatrix} s1(2Ni+N+k) \\ s2(2Ni+N+k) \end{pmatrix}$$

[0372]

For symbol number $2Ni + 2N - 1$:

20 [0373]

Math 124

Equation 117

$$\begin{pmatrix} r1(2Ni+2N-1) \\ r2(2Ni+2N-1) \end{pmatrix} = \frac{1}{\sqrt{\alpha^2+1}} \begin{pmatrix} h_{11}(2Ni+2N-1) & h_{12}(2Ni+2N-1) \\ h_{21}(2Ni+2N-1) & h_{22}(2Ni+2N-1) \end{pmatrix} \begin{pmatrix} e^{j\theta_{11}(2Ni+2N-1)} & \alpha \times e^{j(\theta_{11}(2Ni+2N-1)+\lambda)} \\ \alpha \times e^{j\theta_{21}(2Ni+2N-1)} & e^{j(\theta_{21}(2Ni+2N-1)+\lambda+\delta)} \end{pmatrix} \begin{pmatrix} s1(2Ni+2N-1) \\ s2(2Ni+2N-1) \end{pmatrix}$$

25 [0374]

In this case, it is assumed that only components of direct waves exist in the channel elements $h_{11}(t)$, $h_{12}(t)$, $h_{21}(t)$, and $h_{22}(t)$, that the amplitude components of the direct waves are all equal, and that fluctuations do not occur over time. With these assumptions, Equations 110–117 can be represented as follows.

5 For symbol number $2Ni$ (where i is an integer greater than or equal to zero):

[0375]

Math 125

Equation 118

$$\begin{pmatrix} r1(2Ni) \\ r2(2Ni) \end{pmatrix} = \frac{1}{\sqrt{\beta^2 + 1}} \begin{pmatrix} Ae^{j0} & q \\ Ae^{j0} & q \end{pmatrix} \begin{pmatrix} e^{j\theta_{11}(2Ni)} & \beta \times e^{j(\theta_{11}(2Ni) + \lambda)} \\ \beta \times e^{j\theta_{21}(2Ni)} & e^{j(\theta_{21}(2Ni) + \lambda + \delta)} \end{pmatrix} \begin{pmatrix} s1(2Ni) \\ s2(2Ni) \end{pmatrix}$$

10

[0376]

Here, j is an imaginary unit.

For symbol number $2Ni + 1$:

[0377]

15 Math 126

Equation 119

$$\begin{pmatrix} r1(2Ni + 1) \\ r2(2Ni + 1) \end{pmatrix} = \frac{1}{\sqrt{\beta^2 + 1}} \begin{pmatrix} Ae^{j0} & q \\ Ae^{j0} & q \end{pmatrix} \begin{pmatrix} e^{j\theta_{11}(2Ni+1)} & \beta \times e^{j(\theta_{11}(2Ni+1) + \lambda)} \\ \beta \times e^{j\theta_{21}(2Ni+1)} & e^{j(\theta_{21}(2Ni+1) + \lambda + \delta)} \end{pmatrix} \begin{pmatrix} s1(2Ni + 1) \\ s2(2Ni + 1) \end{pmatrix}$$

[0378]

20 When generalized, this equation is as follows.

For symbol number $2Ni + k$ ($k = 0, 1, \dots, N - 1$):

[0379]

Math 127

Equation 120

$$\begin{pmatrix} r1(2Ni+k) \\ r2(2Ni+k) \end{pmatrix} = \frac{1}{\sqrt{\beta^2+1}} \begin{pmatrix} Ae^{j0} & q \\ Ae^{j0} & q \end{pmatrix} \begin{pmatrix} e^{j\theta_{11}(2Ni+k)} & \beta \times e^{j(\theta_{11}(2Ni+k)+\lambda)} \\ \beta \times e^{j\theta_{21}(2Ni+k)} & e^{j(\theta_{21}(2Ni+k)+\lambda+\delta)} \end{pmatrix} \begin{pmatrix} s1(2Ni+k) \\ s2(2Ni+k) \end{pmatrix}$$

[0380]

Furthermore, for symbol number $2Ni + N - 1$:

5 [0381]

Math 128

Equation 121

$$\begin{pmatrix} r1(2Ni+N-1) \\ r2(2Ni+N-1) \end{pmatrix} = \frac{1}{\sqrt{\beta^2+1}} \begin{pmatrix} Ae^{j0} & q \\ Ae^{j0} & q \end{pmatrix} \begin{pmatrix} e^{j\theta_{11}(2Ni+N-1)} & \beta \times e^{j(\theta_{11}(2Ni+N-1)+\lambda)} \\ \beta \times e^{j\theta_{21}(2Ni+N-1)} & e^{j(\theta_{21}(2Ni+N-1)+\lambda+\delta)} \end{pmatrix} \begin{pmatrix} s1(2Ni+N-1) \\ s2(2Ni+N-1) \end{pmatrix}$$

10 [0382]

For symbol number $2Ni + N$ (where i is an integer greater than or equal to zero):

[0383]

Math 129

Equation 122

$$15 \quad \begin{pmatrix} r1(2Ni+N) \\ r2(2Ni+N) \end{pmatrix} = \frac{1}{\sqrt{\alpha^2+1}} \begin{pmatrix} Ae^{j0} & q \\ Ae^{j0} & q \end{pmatrix} \begin{pmatrix} e^{j\theta_{11}(2Ni+N)} & \alpha \times e^{j(\theta_{11}(2Ni+N)+\lambda)} \\ \alpha \times e^{j\theta_{21}(2Ni+N)} & e^{j(\theta_{21}(2Ni+N)+\lambda+\delta)} \end{pmatrix} \begin{pmatrix} s1(2Ni+N) \\ s2(2Ni+N) \end{pmatrix}$$

[0384]

Here, j is an imaginary unit.

For symbol number $2Ni + N + 1$:

20 [0385]

Math 130

Equation 123

$$\begin{pmatrix} r1(2Ni+N+1) \\ r2(2Ni+N+1) \end{pmatrix} = \frac{1}{\sqrt{\alpha^2+1}} \begin{pmatrix} Ae^{j0} & q \\ Ae^{j0} & q \end{pmatrix} \begin{pmatrix} e^{j\theta_{11}(2Ni+N+1)} & \alpha \times e^{j(\theta_{11}(2Ni+N+1)+\lambda)} \\ \alpha \times e^{j\theta_{21}(2Ni+N+1)} & e^{j(\theta_{21}(2Ni+N+1)+\lambda+\delta)} \end{pmatrix} \begin{pmatrix} s1(2Ni+N+1) \\ s2(2Ni+N+1) \end{pmatrix}$$

[0386]

When generalized, this equation is as follows.

For symbol number $2Ni + N + k$ ($k = 0, 1, \dots, N - 1$):

5 [0387]

Math 131

Equation 124

$$\begin{pmatrix} r1(2Ni + N + k) \\ r2(2Ni + N + k) \end{pmatrix} = \frac{1}{\sqrt{\alpha^2 + 1}} \begin{pmatrix} A e^{j0} & q \\ A e^{j0} & q \end{pmatrix} \begin{pmatrix} e^{j\theta_{1i}(2Ni + N + k)} & \alpha \times e^{j(\theta_{1i}(2Ni + N + k) + \lambda)} \\ \alpha \times e^{j\theta_{2i}(2Ni + N + k)} & e^{j(\theta_{2i}(2Ni + N + k) + \lambda + \delta)} \end{pmatrix} \begin{pmatrix} s1(2Ni + N + k) \\ s2(2Ni + N + k) \end{pmatrix}$$

10 [0388]

Furthermore, for symbol number $2Ni + 2N - 1$:

[0389]

Math 132

Equation 125

$$15 \quad \begin{pmatrix} r1(2Ni + 2N - 1) \\ r2(2Ni + 2N - 1) \end{pmatrix} = \frac{1}{\sqrt{\alpha^2 + 1}} \begin{pmatrix} A e^{j0} & q \\ A e^{j0} & q \end{pmatrix} \begin{pmatrix} e^{j\theta_{1i}(2Ni + 2N - 1)} & \alpha \times e^{j(\theta_{1i}(2Ni + 2N - 1) + \lambda)} \\ \alpha \times e^{j\theta_{2i}(2Ni + 2N - 1)} & e^{j(\theta_{2i}(2Ni + 2N - 1) + \lambda + \delta)} \end{pmatrix} \begin{pmatrix} s1(2Ni + 2N - 1) \\ s2(2Ni + 2N - 1) \end{pmatrix}$$

[0390]

In Equations 118–125, let A be a real number and q be a complex number.

Equations 118–125 can be represented as follows.

20 For symbol number $2Ni$ (where i is an integer greater than or equal to zero):

[0391]

Math 133

Equation 126

$$\begin{pmatrix} r1(2Ni) \\ r2(2Ni) \end{pmatrix} = \frac{1}{\sqrt{\beta^2 + 1}} \begin{pmatrix} e^{j0} \\ e^{j0} \end{pmatrix} \begin{pmatrix} A e^{j0} & q \\ \beta \times e^{j\theta_{1i}(2Ni)} & \beta \times e^{j(\theta_{1i}(2Ni) + \lambda)} \\ \beta \times e^{j\theta_{2i}(2Ni)} & e^{j(\theta_{2i}(2Ni) + \lambda + \delta)} \end{pmatrix} \begin{pmatrix} s1(2Ni) \\ s2(2Ni) \end{pmatrix}$$

[0392]

Here, j is an imaginary unit.

For symbol number $2Ni + 1$:

5 [0393]

Math 134

Equation 127

$$\begin{pmatrix} r1(2Ni+1) \\ r2(2Ni+1) \end{pmatrix} = \frac{1}{\sqrt{\beta^2 + 1}} \begin{pmatrix} e^{j0} \\ e^{j0} \end{pmatrix} \left(A e^{j0} \quad q \begin{pmatrix} e^{j\theta_{11}(2Ni+1)} & \beta \times e^{j(\theta_{11}(2Ni+1)+\lambda)} \\ \beta \times e^{j\theta_{21}(2Ni+1)} & e^{j(\theta_{21}(2Ni+1)+\lambda+\delta)} \end{pmatrix} \right) \begin{pmatrix} s1(2Ni+1) \\ s2(2Ni+1) \end{pmatrix}$$

10 [0394]

When generalized, this equation is as follows.

For symbol number $2Ni + k$ ($k = 0, 1, \dots, N - 1$):

[0395]

Math 135

15 Equation 128

$$\begin{pmatrix} r1(2Ni+k) \\ r2(2Ni+k) \end{pmatrix} = \frac{1}{\sqrt{\beta^2 + 1}} \begin{pmatrix} e^{j0} \\ e^{j0} \end{pmatrix} \left(A e^{j0} \quad q \begin{pmatrix} e^{j\theta_{11}(2Ni+k)} & \beta \times e^{j(\theta_{11}(2Ni+k)+\lambda)} \\ \beta \times e^{j\theta_{21}(2Ni+k)} & e^{j(\theta_{21}(2Ni+k)+\lambda+\delta)} \end{pmatrix} \right) \begin{pmatrix} s1(2Ni+k) \\ s2(2Ni+k) \end{pmatrix}$$

[0396]

Furthermore, for symbol number $2Ni + N - 1$:

20 [0397]

Math 136

Equation 129

$$\begin{pmatrix} r1(2Ni+N-1) \\ r2(2Ni+N-1) \end{pmatrix} = \frac{1}{\sqrt{\beta^2 + 1}} \begin{pmatrix} e^{j0} \\ e^{j0} \end{pmatrix} \left(A e^{j0} \quad q \begin{pmatrix} e^{j\theta_{11}(2Ni+N-1)} & \beta \times e^{j(\theta_{11}(2Ni+N-1)+\lambda)} \\ \beta \times e^{j\theta_{21}(2Ni+N-1)} & e^{j(\theta_{21}(2Ni+N-1)+\lambda+\delta)} \end{pmatrix} \right) \begin{pmatrix} s1(2Ni+N-1) \\ s2(2Ni+N-1) \end{pmatrix}$$

[0398]

For symbol number $2Ni + N$ (where i is an integer greater than or equal to zero):

[0399]

5 Math 137

Equation 130

$$\begin{pmatrix} r1(2Ni + N) \\ r2(2Ni + N) \end{pmatrix} = \frac{1}{\sqrt{\alpha^2 + 1}} \begin{pmatrix} e^{j0} \\ e^{j0} \end{pmatrix} \left(A e^{j0} \quad q \begin{pmatrix} e^{j\theta_{11}(2Ni+N)} & \alpha \times e^{j(\theta_{11}(2Ni+N)+\lambda)} \\ \alpha \times e^{j\theta_{21}(2Ni+N)} & e^{j(\theta_{21}(2Ni+N)+\lambda+\delta)} \end{pmatrix} \right) \begin{pmatrix} s1(2Ni + N) \\ s2(2Ni + N) \end{pmatrix}$$

[0400]

10 Here, j is an imaginary unit.

For symbol number $2Ni + N + 1$:

[0401]

Math 138

Equation 131

$$15 \quad \begin{pmatrix} r1(2Ni + N + 1) \\ r2(2Ni + N + 1) \end{pmatrix} = \frac{1}{\sqrt{\alpha^2 + 1}} \begin{pmatrix} e^{j0} \\ e^{j0} \end{pmatrix} \left(A e^{j0} \quad q \begin{pmatrix} e^{j\theta_{11}(2Ni+N+1)} & \alpha \times e^{j(\theta_{11}(2Ni+N+1)+\lambda)} \\ \alpha \times e^{j\theta_{21}(2Ni+N+1)} & e^{j(\theta_{21}(2Ni+N+1)+\lambda+\delta)} \end{pmatrix} \right) \begin{pmatrix} s1(2Ni + N + 1) \\ s2(2Ni + N + 1) \end{pmatrix}$$

[0402]

When generalized, this equation is as follows.

For symbol number $2Ni + N + k$ ($k = 0, 1, \dots, N - 1$):

20 [0403]

Math 139

Equation 132

$$\begin{pmatrix} r1(2Ni + N + k) \\ r2(2Ni + N + k) \end{pmatrix} = \frac{1}{\sqrt{\alpha^2 + 1}} \begin{pmatrix} e^{j0} \\ e^{j0} \end{pmatrix} \left(A e^{j0} \quad q \begin{pmatrix} e^{j\theta_{11}(2Ni+N+k)} & \alpha \times e^{j(\theta_{11}(2Ni+N+k)+\lambda)} \\ \alpha \times e^{j\theta_{21}(2Ni+N+k)} & e^{j(\theta_{21}(2Ni+N+k)+\lambda+\delta)} \end{pmatrix} \right) \begin{pmatrix} s1(2Ni + N + k) \\ s2(2Ni + N + k) \end{pmatrix}$$

25 [0404]

Furthermore, for symbol number $2Ni + 2N - 1$:

[0405]

Math 140

Equation 133

$$5 \quad \begin{pmatrix} r1(2Ni + 2N - 1) \\ r2(2Ni + 2N - 1) \end{pmatrix} = \frac{1}{\sqrt{\alpha^2 + 1}} \begin{pmatrix} e^{j0} \\ e^{j0} \end{pmatrix} \left(A e^{j0} \quad q \begin{pmatrix} e^{j\theta_{11}(2Ni + 2N - 1)} & \alpha \times e^{j(\theta_{11}(2Ni + 2N - 1) - \lambda)} \\ \alpha \times e^{j\theta_{21}(2Ni + 2N - 1)} & e^{j(\theta_{21}(2Ni + 2N - 1) - \lambda + \delta)} \end{pmatrix} \begin{pmatrix} s1(2Ni + 2N - 1) \\ s2(2Ni + 2N - 1) \end{pmatrix} \right)$$

[0406]

As a result, when q is represented as follows, one of the signals $s1$ and $s2$ can no longer be obtained.

10 For symbol number $2Ni$ (where i is an integer greater than or equal to zero):

[0407]

Math 141

Equation 134

$$q = -\frac{A}{\beta} e^{j(\theta_{11}(2Ni) - \theta_{21}(2Ni))}, -A\beta e^{j(\theta_{11}(2Ni) - \theta_{21}(2Ni) - \delta)}$$

15

[0408]

For symbol number $2Ni + 1$:

[0409]

Math 142

20 Equation 135

$$q = -\frac{A}{\beta} e^{j(\theta_{11}(2Ni+1) - \theta_{21}(2Ni+1))}, -A\beta e^{j(\theta_{11}(2Ni+1) - \theta_{21}(2Ni+1) - \delta)}$$

[0410]

When generalized, this equation is as follows.

25 For symbol number $2Ni + k$ ($k = 0, 1, \dots, N - 1$):

[0411]

Math 143

Equation 136

$$q = -\frac{A}{\beta} e^{j(\theta_{11}(2Ni+k)-\theta_{21}(2Ni+k))}, -A\beta e^{j(\theta_{11}(2Ni+k)-\theta_{21}(2Ni+k)-\delta)}$$

5

[0412]

Furthermore, for symbol number $2Ni + N - 1$:

[0413]

Math 144

10 Equation 137

$$q = -\frac{A}{\beta} e^{j(\theta_{11}(2Ni+N-1)-\theta_{21}(2Ni+N-1))}, -A\beta e^{j(\theta_{11}(2Ni+N-1)-\theta_{21}(2Ni+N-1)-\delta)}$$

[0414]

For symbol number $2Ni + N$ (where i is an integer greater than or equal to zero):

15 [0415]

Math 145

Equation 138

$$q = -\frac{A}{\alpha} e^{j(\theta_{11}(2Ni+N)-\theta_{21}(2Ni+N))}, -A\alpha e^{j(\theta_{11}(2Ni+N)-\theta_{21}(2Ni+N)-\delta)}$$

20 [0416]

For symbol number $2Ni + N + 1$:

[0417]

Math 146

Equation 139

$$q = -\frac{A}{\alpha} e^{j(\theta_{11}(2Ni+N+1)-\theta_{21}(2Ni+N+1))}, -A\alpha e^{j(\theta_{11}(2Ni+N+1)-\theta_{21}(2Ni+N+1)-\delta)}$$

[0418]

When generalized, this equation is as follows.

5 For symbol number $2Ni + N + k$ ($k = 0, 1, \dots, N - 1$):

[0419]

Math 147

Equation 140

$$q = -\frac{A}{\alpha} e^{j(\theta_{11}(2Ni+N+k)-\theta_{21}(2Ni+N+k))}, -A\alpha e^{j(\theta_{11}(2Ni+N+k)-\theta_{21}(2Ni+N+k)-\delta)}$$

10

[0420]

Furthermore, for symbol number $2Ni + 2N - 1$:

[0421]

Math 148

15 Equation 141

$$q = -\frac{A}{\alpha} e^{j(\theta_{11}(2Ni+2N-1)-\theta_{21}(2Ni+2N-1))}, -A\alpha e^{j(\theta_{11}(2Ni+2N-1)-\theta_{21}(2Ni+2N-1)-\delta)}$$

[0422]

20 In this case, if q has the same solution in symbol numbers $2Ni$ through $2Ni + N - 1$, then since the channel elements of the direct waves do not greatly fluctuate, excellent reception quality can no longer be obtained for any of the symbol numbers. Therefore, it is difficult to achieve the ability to correct errors, even if error correction codes are introduced. Accordingly, for q not to have the same solution, Condition #7 or Condition #8 becomes necessary from Equations 134–141 and from

the fact that $\alpha \neq \beta$ when focusing on one of two solutions of q which does not include δ .

[0423]

Math 149

5 Condition #7

$$e^{j(\theta_{11}(2Ni+x)-\theta_{21}(2Ni+x))} \neq e^{j(\theta_{11}(2Ni+y)-\theta_{21}(2Ni+y))} \quad \text{for } \forall x, \forall y (x \neq y; x, y = 0, 1, 2, \dots, N-2, N-1)$$

(x is 0, 1, 2, ..., $N-2, N-1$; y is 0, 1, 2, ..., $N-2, N-1$; and $x \neq y$.)

and

$$e^{j(\theta_{11}(2Ni+N+x)-\theta_{21}(2Ni+N+x))} \neq e^{j(\theta_{11}(2Ni+N+y)-\theta_{21}(2Ni+N+y))} \quad \text{for } \forall x, \forall y (x \neq y; x, y = 0, 1, 2, \dots, N-2, N-1)$$

10 (x is 0, 1, 2, ..., $N-2, N-1$; y is 0, 1, 2, ..., $N-2, N-1$; and $x \neq y$.)

[0424]

Math 150

Condition #8

$$15 \quad e^{j(\theta_{11}(2Ni+x)-\theta_{21}(2Ni+x))} \neq e^{j(\theta_{11}(2Ni+y)-\theta_{21}(2Ni+y))} \quad \text{for } \forall x, \forall y (x \neq y; x, y = 0, 1, 2, \dots, 2N-2, 2N-1)$$

[0425]

In this case, Condition #8 is similar to the conditions described in Embodiment 1 through Embodiment 3. However, with regards to Condition #7, since $\alpha \neq \beta$, the solution not including δ among the two solutions of q is a different solution.

[0426]

Next, design requirements for not only θ_{11} and θ_{12} , but also for λ and δ are described. It suffices to set λ to a certain value; it is then necessary to establish requirements for δ . The following describes the design method for δ when λ is set to zero radians.

[0427]

In this case, similar to the method of changing the precoding weights in a four-slot period (cycle), by defining δ so that $\pi/2$ radians $\leq |\delta| \leq \pi$ radians, excellent reception quality is achieved, particularly in an LOS environment.

[0428]

5 In symbol numbers $2N_i$ through $2N_i + 2N - 1$, two points q exist where reception quality becomes poor, and therefore $4N$ such points exist. In an LOS environment, in order to achieve excellent characteristics, these $4N$ points should each have a different solution. In this case, focusing on amplitude, the following condition is necessary for Condition #7 or Condition #8, since $\alpha \neq \beta$.

10 [0429]

Math 151

Condition #9

$$\alpha \neq \frac{1}{\beta}$$

15 [0430]

As described above, when a transmission device transmits a plurality of modulated signals from a plurality of antennas in a MIMO system, the advantageous effect of improved transmission quality, as compared to conventional spatial multiplexing MIMO system, is achieved in an LOS environment in which direct waves dominate by hopping between precoding weights regularly over time.

[0431]

In the present embodiment, the structure of the reception device is as described in Embodiment 1, and in particular with regards to the structure of the reception device, operations have been described for a limited number of antennas, but the present invention may be embodied in the same way even if the number of antennas increases. In other words, the number of antennas in the reception device

does not affect the operations or advantageous effects of the present embodiment. Furthermore, in the present embodiment, similar to Embodiment 1, the error correction codes are not limited.

[0432]

5 In the present embodiment, in contrast with Embodiment 1, the method of changing the precoding weights in the time domain has been described. As described in Embodiment 1, however, the present invention may be similarly embodied by changing the precoding weights by using a multi-carrier transmission method and arranging symbols in the frequency domain and the frequency-time
10 domain. Furthermore, in the present embodiment, symbols other than data symbols, such as pilot symbols (preamble, unique word, and the like), symbols for control information, and the like, may be arranged in the frame in any way.

[0433]

(Embodiment 5)

15 In Embodiment 1 through Embodiment 4, the method of regularly hopping between precoding weights has been described. In the present embodiment, a modification of this method is described.

[0434]

In Embodiment 1 through Embodiment 4, the method of regularly hopping
20 between precoding weights as in Fig. 6 has been described. In the present embodiment, a method of regularly hopping between precoding weights that differs from Fig. 6 is described.

[0435]

As in Fig. 6, this method hops between four different precoding weights
25 (matrices). Fig. 22 shows the hopping method that differs from Fig. 6. In Fig. 22, four different precoding weights (matrices) are represented as W1, W2, W3, and W4. (For example, W1 is the precoding weight (matrix) in Equation 37, W2 is the precoding weight (matrix) in Equation 38, W3 is the precoding weight (matrix) in

Equation 39, and $W4$ is the precoding weight (matrix) in Equation 40.) In Fig. 3, elements that operate in a similar way to Fig. 3 and Fig. 6 bear the same reference signs.

The parts unique to Fig. 22 are as follows.

- 5 · The first period (cycle) 2201, the second period (cycle) 2202, the third period (cycle) 2203, ... are all four-slot periods (cycles).
- A different precoding weight matrix is used in each of the four slots, i.e. $W1$, $W2$, $W3$, and $W4$ are each used once.
- It is not necessary for $W1$, $W2$, $W3$, and $W4$ to be in the same order in the first
10 period (cycle) 2201, the second period (cycle) 2202, the third period (cycle) 2203, ...

In order to implement this method, a precoding weight generating unit 2200 receives, as an input, a signal regarding a weighting method and outputs information 2210 regarding precoding weights in order for each period (cycle). The weighting
15 unit 600 receives, as inputs, this information, $s1(t)$, and $s2(t)$, performs weighting, and outputs $z1(t)$ and $z2(t)$.

[0436]

Fig. 23 shows a different weighting method than Fig. 22 for the above precoding method. In Fig. 23, the difference from Fig. 22 is that a similar method to
20 Fig. 22 is achieved by providing a reordering unit after the weighting unit and by reordering signals.

[0437]

In Fig. 23, the precoding weight generating unit 2200 receives, as an input, information 315 regarding a weighting method and outputs information 2210 on
25 precoding weights in the order of precoding weights $W1$, $W2$, $W3$, $W4$, $W1$, $W2$, $W3$, $W4$, ... Accordingly, the weighting unit 600 uses the precoding weights in the order of precoding weights $W1$, $W2$, $W3$, $W4$, $W1$, $W2$, $W3$, $W4$, ... and outputs precoded signals 2300A and 2300B.

[0438]

A reordering unit 2300 receives, as inputs, the precoded signals 2300A and 2300B, reorders the precoded signals 2300A and 2300B in the order of the first period (cycle) 2201, the second period (cycle) 2202, and the third period (cycle) 2203 in Fig. 23, and outputs $z1(t)$ and $z2(t)$.

[0439]

Note that in the above description, the period (cycle) for hopping between precoding weights has been described as having four slots for the sake of comparison with Fig. 6. As in Embodiment 1 through Embodiment 4, however, the present invention may be similarly embodied with a period (cycle) having other than four slots.

[0440]

Furthermore, in Embodiment 1 through Embodiment 4, and in the above precoding method, within the period (cycle), the value of δ and β has been described as being the same for each slot, but the value of δ and β may change in each slot.

[0441]

As described above, when a transmission device transmits a plurality of modulated signals from a plurality of antennas in a MIMO system, the advantageous effect of improved transmission quality, as compared to conventional spatial multiplexing MIMO system, is achieved in an LOS environment in which direct waves dominate by hopping between precoding weights regularly over time.

[0442]

In the present embodiment, the structure of the reception device is as described in Embodiment 1, and in particular with regards to the structure of the reception device, operations have been described for a limited number of antennas, but the present invention may be embodied in the same way even if the number of antennas increases. In other words, the number of antennas in the reception device does not affect the operations or advantageous effects of the present embodiment.

Furthermore, in the present embodiment, similar to Embodiment 1, the error correction codes are not limited.

[0443]

In the present embodiment, in contrast with Embodiment 1, the method of changing the precoding weights in the time domain has been described. As described in Embodiment 1, however, the present invention may be similarly embodied by changing the precoding weights by using a multi-carrier transmission method and arranging symbols in the frequency domain and the frequency-time domain. Furthermore, in the present embodiment, symbols other than data symbols, such as pilot symbols (preamble, unique word, and the like), symbols for control information, and the like, may be arranged in the frame in any way.

[0444]

(Embodiment 6)

In Embodiments 1–4, a method for regularly hopping between precoding weights has been described. In the present embodiment, a method for regularly hopping between precoding weights is again described, including the content that has been described in Embodiments 1–4.

[0445]

First, out of consideration of an LOS environment, a method of designing a precoding matrix is described for a 2×2 spatial multiplexing MIMO system that adopts precoding in which feedback from a communication partner is not available.

[0446]

Fig. 30 shows a model of a 2×2 spatial multiplexing MIMO system that adopts precoding in which feedback from a communication partner is not available. An information vector z is encoded and interleaved. As output of the interleaving, an encoded bit vector $u(p) = (u_1(p), u_2(p))$ is acquired (where p is the slot time). Let $u_i(p) = (u_{i1}(p), \dots, u_{ih}(p))$ (where h is the number of transmission bits per symbol). Letting a signal after modulation (mapping) be $s(p) = (s1(p), s2(p))^T$ and a precoding

matrix be $\mathbf{F}(p)$, a precoded symbol $\mathbf{x}(p) = (x_1(p), x_2(p))^T$ is represented by the following equation.

[0447]

Math 152

5 Equation 142

$$\begin{aligned}\mathbf{x}(p) &= \left(x_1(p), x_2(p) \right)^T \\ &= \mathbf{F}(p)\mathbf{s}(p)\end{aligned}$$

[0448]

Accordingly, letting a received vector be $\mathbf{y}(p) = (y_1(p), y_2(p))^T$, the received
10 vector $\mathbf{y}(p)$ is represented by the following equation.

[0449]

Math 153

Equation 143

$$\begin{aligned}\mathbf{y}(p) &= \left(y_1(p), y_2(p) \right)^T \\ &= \mathbf{H}(p)\mathbf{F}(p)\mathbf{s}(p) + \mathbf{n}(p)\end{aligned}$$

15

[0450]

In this Equation, $\mathbf{H}(p)$ is the channel matrix, $\mathbf{n}(p) = (n_1(p), n_2(p))^T$ is the
noise vector, and $n_i(p)$ is the i.i.d. complex Gaussian random noise with an average
value 0 and variance σ^2 . Letting the Rician factor be K , the above equation can be
20 represented as follows.

[0451]

Math 154

Equation 144

$$\begin{aligned} \mathbf{y}(p) &= (y_1(p), y_2(p))^T \\ &= \left(\sqrt{\frac{K}{K+1}} \mathbf{H}_d(p) + \sqrt{\frac{1}{K+1}} \mathbf{H}_s(p) \right) \mathbf{F}(p) \mathbf{s}(p) + \mathbf{n}(p) \end{aligned}$$

[0452]

In this equation, $\mathbf{H}_d(p)$ is the channel matrix for the direct wave components,
 5 and $\mathbf{H}_s(p)$ is the channel matrix for the scattered wave components. Accordingly, the
 channel matrix $\mathbf{H}(p)$ is represented as follows.

[0453]

Math 155

Equation 145

$$\begin{aligned} \mathbf{H}(p) &= \sqrt{\frac{K}{K+1}} \mathbf{H}_d(p) + \sqrt{\frac{1}{K+1}} \mathbf{H}_s(p) \\ 10 \quad &= \sqrt{\frac{K}{K+1}} \begin{pmatrix} h_{11,d} & h_{12,d} \\ h_{21,d} & h_{22,d} \end{pmatrix} + \sqrt{\frac{1}{K+1}} \begin{pmatrix} h_{11,s}(p) & h_{12,s}(p) \\ h_{21,s}(p) & h_{22,s}(p) \end{pmatrix} \end{aligned}$$

[0454]

In Equation 145, it is assumed that the direct wave environment is uniquely
 determined by the positional relationship between transmitters, and that the channel
 15 matrix $\mathbf{H}_d(p)$ for the direct wave components does not fluctuate with time.
 Furthermore, in the channel matrix $\mathbf{H}_d(p)$ for the direct wave components, it is
 assumed that as compared to the interval between transmitting antennas, the
 probability of an environment with a sufficiently long distance between transmission
 and reception devices is high, and therefore that the channel matrix for the direct

wave components can be treated as a non-singular matrix. Accordingly, the channel matrix $\mathbf{H}_d(p)$ is represented as follows.

[0455]

Math 156

5 Equation 146

$$\begin{aligned}\mathbf{H}_d(p) &= \begin{pmatrix} h_{11,d} & h_{12,d} \\ h_{21,d} & h_{22,d} \end{pmatrix} \\ &= \begin{pmatrix} A e^{j\psi} & q \\ A e^{j\psi} & q \end{pmatrix}\end{aligned}$$

[0456]

In this equation, let A be a positive real number and q be a complex number.

10 Subsequently, out of consideration of an LOS environment, a method of designing a precoding matrix is described for a 2×2 spatial multiplexing MIMO system that adopts precoding in which feedback from a communication partner is not available.

[0457]

From Equations 144 and 145, it is difficult to seek a precoding matrix
15 without appropriate feedback in conditions including scattered waves, since it is difficult to perform analysis under conditions including scattered waves. Additionally, in a NLOS environment, little degradation in reception quality of data occurs as compared to an LOS environment. Therefore, the following describes a method of designing precoding matrices without appropriate feedback in an LOS
20 environment (precoding matrices for a precoding method that hops between precoding matrices over time).

[0458]

As described above, since it is difficult to perform analysis under conditions including scattered waves, an appropriate precoding matrix for a channel matrix including components of only direct waves is sought from Equations 144 and 145. Therefore, in Equation 144, the case when the channel matrix includes components
 5 of only direct waves is considered. It follows that from Equation 146, Equation 144 can be represented as follows.

[0459]

Math 157

Equation 147

$$\begin{pmatrix} y_1(p) \\ y_2(p) \end{pmatrix} = \mathbf{H}_d(p) \mathbf{F}(p) \mathbf{s}(p) + \mathbf{n}(p)$$

$$10 \quad = \begin{pmatrix} A e^{j\psi} & q \\ A e^{j\psi} & q \end{pmatrix} \mathbf{F}(p) \mathbf{s}(p) + \mathbf{n}(p)$$

[0460]

In this equation, a unitary matrix is used as the precoding matrix. Accordingly, the precoding matrix is represented as follows.

15 [0461]

Math 158

Equation 148

$$\mathbf{F}(p) = \frac{1}{\sqrt{\alpha^2 + 1}} \begin{pmatrix} e^{j\theta_{11}(p)} & \alpha \times e^{j(\theta_{11}(p) + \lambda)} \\ \alpha \times e^{j\theta_{21}(p)} & e^{j(\theta_{21}(p) + \lambda + \pi)} \end{pmatrix}$$

20 [0462]

In this equation, λ is a fixed value. Therefore, Equation 147 can be represented as follows.

[0463]

Math 159

5 Equation 149

$$\begin{pmatrix} y_1(p) \\ y_2(p) \end{pmatrix} = \frac{1}{\sqrt{\alpha^2 + 1}} \begin{pmatrix} A e^{j\psi} & q \\ A e^{j\psi} & q \end{pmatrix} \begin{pmatrix} e^{j\theta_{11}(p)} & \alpha \times e^{j(\theta_{11}(p) + \lambda)} \\ \alpha \times e^{j\theta_{21}(p)} & e^{j(\theta_{21}(p) + \lambda + \pi)} \end{pmatrix} \begin{pmatrix} s1(p) \\ s2(p) \end{pmatrix} + \mathbf{n}(p)$$

[0464]

As is clear from Equation 149, when the reception device performs linear
 10 operation of Zero Forcing (ZF) or the Minimum Mean Squared Error (MMSE), the
 transmitted bit cannot be determined by $s1(p)$, $s2(p)$. Therefore, the iterative APP (or
 iterative Max-log APP) or APP (or Max-log APP) described in Embodiment 1 is
 performed (hereafter referred to as Maximum Likelihood (ML) calculation), the
 log-likelihood ratio of each bit transmitted in $s1(p)$, $s2(p)$ is sought, and decoding
 15 with error correction codes is performed. Accordingly, the following describes a
 method of designing a precoding matrix without appropriate feedback in an LOS
 environment for a reception device that performs ML calculation.

[0465]

The precoding in Equation 149 is considered. The right-hand side and
 20 left-hand side of the first line are multiplied by $e^{-j\psi}$, and similarly the right-hand side
 and left-hand side of the second line are multiplied by $e^{-j\psi}$. The following equation
 represents the result.

[0466]

Math 160

25 Equation 150

$$\begin{aligned}
& \begin{pmatrix} e^{-j\psi} y_1(p) \\ e^{-j\psi} y_2(p) \end{pmatrix} \\
&= e^{-j\psi} \left\{ \frac{1}{\sqrt{\alpha^2 + 1}} \begin{pmatrix} A e^{j\psi} & q \\ A e^{j\psi} & q \end{pmatrix} \begin{pmatrix} e^{j\theta_{11}(p)} & \alpha \times e^{j(\theta_{11}(p) + \lambda)} \\ \alpha \times e^{j\theta_{21}(p)} & e^{j(\theta_{21}(p) + \lambda + \pi)} \end{pmatrix} \begin{pmatrix} s1(p) \\ s2(p) \end{pmatrix} + \mathbf{n}(p) \right\} \\
&= \frac{1}{\sqrt{\alpha^2 + 1}} \begin{pmatrix} A e^{j0} & e^{-j\psi} q \\ A e^{j0} & e^{-j\psi} q \end{pmatrix} \begin{pmatrix} e^{j\theta_{11}(p)} & \alpha \times e^{j(\theta_{11}(p) + \lambda)} \\ \alpha \times e^{j\theta_{21}(p)} & e^{j(\theta_{21}(p) + \lambda + \pi)} \end{pmatrix} \begin{pmatrix} s1(p) \\ s2(p) \end{pmatrix} + e^{-j\psi} \mathbf{n}(p)
\end{aligned}$$

[0467]

$e^{j\psi} y_1(p)$, $e^{j\psi} y_2(p)$, and $e^{j\psi} q$ are respectively redefined as $y_1(p)$, $y_2(p)$, and q .

- 5 Furthermore, since $e^{j\psi} \mathbf{n}(p) = (e^{j\psi} n_1(p), e^{j\psi} n_2(p))^T$, and $e^{j\psi} n_1(p)$, $e^{j\psi} n_2(p)$ are the independent identically distributed (i.i.d.) complex Gaussian random noise with an average value 0 and variance σ^2 , $e^{j\psi} \mathbf{n}(p)$ is redefined as $\mathbf{n}(p)$. As a result, generality is not lost by restating Equation 150 as Equation 151.

[0468]

10 Math 161

Equation 151

$$\begin{aligned}
& \begin{pmatrix} y_1(p) \\ y_2(p) \end{pmatrix} \\
&= \frac{1}{\sqrt{\alpha^2 + 1}} \begin{pmatrix} A e^{j0} & q \\ A e^{j0} & q \end{pmatrix} \begin{pmatrix} e^{j\theta_{11}(p)} & \alpha \times e^{j(\theta_{11}(p) + \lambda)} \\ \alpha \times e^{j\theta_{21}(p)} & e^{j(\theta_{21}(p) + \lambda + \pi)} \end{pmatrix} \begin{pmatrix} s1(p) \\ s2(p) \end{pmatrix} + \mathbf{n}(p)
\end{aligned}$$

[0469]

- 15 Next, Equation 151 is transformed into Equation 152 for the sake of clarity.

[0470]

Math 162

Equation 152

$$\begin{pmatrix} y_1(p) \\ y_2(p) \end{pmatrix} = \frac{1}{\sqrt{\alpha^2 + 1}} \begin{pmatrix} e^{j0} \\ e^{j0} \end{pmatrix} \left(A e^{j0} \quad q \begin{pmatrix} e^{j\theta_{11}(p)} & \alpha \times e^{j(\theta_{11}(p) + \lambda)} \\ \alpha \times e^{j\theta_{21}(p)} & e^{j(\theta_{21}(p) + \lambda + \pi)} \end{pmatrix} \begin{pmatrix} s1(p) \\ s2(p) \end{pmatrix} \right) + \mathbf{n}(p)$$

[0471]

In this case, letting the minimum Euclidian distance between a received
 5 signal point and a received candidate signal point be $d_{\min}^2$, then a poor point has a
 minimum value of zero for $d_{\min}^2$, and two values of q exist at which conditions are
 poor in that all of the bits transmitted by $s1(p)$ and all of the bits transmitted by $s2(p)$
 being eliminated.

[0472]

10 In Equation 152, when $s1(p)$ does not exist.

[0473]

Math 163

Equation 153

$$q = -\frac{A}{\alpha} e^{j(\theta_{11}(p) - \theta_{21}(p))}$$

15

[0474]

In Equation 152, when $s2(p)$ does not exist.

[0475]

Math 164

20 Equation 154

$$q = -A\alpha e^{j(\theta_{11}(p) - \theta_{21}(p) - \pi)}$$

[0476]

(Hereinafter, the values of q satisfying Equations 153 and 154 are respectively referred to as “poor reception points for $s1$ and $s2$ ”).

5 When Equation 153 is satisfied, since all of the bits transmitted by $s1(p)$ are eliminated, the received log-likelihood ratio cannot be sought for any of the bits transmitted by $s1(p)$. When Equation 154 is satisfied, since all of the bits transmitted by $s2(p)$ are eliminated, the received log-likelihood ratio cannot be sought for any of the bits transmitted by $s2(p)$.

10 [0477]

A broadcast/multicast transmission system that does not change the precoding matrix is now considered. In this case, a system model is considered in which a base station transmits modulated signals using a precoding method that does not hop between precoding matrices, and a plurality of terminals (Γ terminals)

15 receive the modulated signals transmitted by the base station.

[0478]

It is considered that the conditions of direct waves between the base station and the terminals change little over time. Therefore, from Equations 153 and 154, for a terminal that is in a position fitting the conditions of Equation 155 or Equation

20 156 and that is in an LOS environment where the Rician factor is large, the possibility of degradation in the reception quality of data exists. Accordingly, to resolve this problem, it is necessary to change the precoding matrix over time.

[0479]

Math 165

25 Equation 155

$$q \approx -\frac{A}{\alpha} e^{j(\theta_{11}(p) - \theta_{21}(p))}$$

[0480]

Math 166

Equation 156

$$5 \quad q \approx -A\alpha e^{j(\theta_{11}(p)-\theta_{21}(p)-\pi)}$$

[0481]

A method of regularly hopping between precoding matrices over a time period (cycle) with N slots (hereinafter referred to as a precoding hopping method) is considered.

[0482]

Since there are N slots in the time period (cycle), N varieties of precoding matrices $F[i]$ based on Equation 148 are prepared ($i = 0, 1, \dots, N - 1$). In this case, the precoding matrices $F[i]$ are represented as follows.

15 [0483]

Math 167

Equation 157

$$\mathbf{F}[i] = \frac{1}{\sqrt{\alpha^2 + 1}} \begin{pmatrix} e^{j\theta_{11}[i]} & \alpha \times e^{j(\theta_{11}[i] + \lambda)} \\ \alpha \times e^{j\theta_{21}[i]} & e^{j(\theta_{21}[i] + \lambda + \pi)} \end{pmatrix}$$

20 [0484]

In this equation, let α not change over time, and let λ also not change over time (though change over time may be allowed).

[0485]

As in Embodiment 1, $F[i]$ is the precoding matrix used to obtain a precoded signal x ($p = N \times k + i$) in Equation 142 for time $N \times k + i$ (where k is an integer equal to or greater than 0, and $i = 0, 1, \dots, N - 1$). The same is true below as well.

[0486]

- 5 At this point, based on Equations 153 and 154, design conditions such as the following are important for the precoding matrices for precoding hopping.

[0487]

Math 168

Condition #10

- 10 Equation 158

$$e^{j(\theta_{11}[x]-\theta_{21}[x])} \neq e^{j(\theta_{11}[y]-\theta_{21}[y])}$$

for $\forall x, \forall y \quad (x \neq y; \quad x, y = 0, 1, \dots, N - 1)$

[0488]

Math 169

- 15 Condition #11

Equation 159

$$e^{j(\theta_{11}[x]-\theta_{21}[x]-\pi)} \neq e^{j(\theta_{11}[y]-\theta_{21}[y]-\pi)}$$

for $\forall x, \forall y \quad (x \neq y; \quad x, y = 0, 1, \dots, N - 1)$

[0489]

- 20 From Condition #10, in all of the Γ terminals, there is one slot or less having poor reception points for $s1$ among the N slots in a time period (cycle). Accordingly, the log-likelihood ratio for bits transmitted by $s1(p)$ can be obtained for at least $N - 1$ slots. Similarly, from Condition #11, in all of the Γ terminals, there is one slot or less having poor reception points for $s2$ among the N slots in a time

period (cycle). Accordingly, the log-likelihood ratio for bits transmitted by $s_2(p)$ can be obtained for at least $N - 1$ slots.

[0490]

In this way, by providing the precoding matrix design model of Condition #10 and Condition #11, the number of bits for which the log-likelihood ratio is obtained among the bits transmitted by $s_1(p)$, and the number of bits for which the log-likelihood ratio is obtained among the bits transmitted by $s_2(p)$ is guaranteed to be equal to or greater than a fixed number in all of the Γ terminals. Therefore, in all of the Γ terminals, it is considered that degradation of data reception quality is moderated in an LOS environment where the Rician factor is large.

[0491]

The following shows an example of a precoding matrix in the precoding hopping method.

[0492]

The probability density distribution of the phase of a direct wave can be considered to be evenly distributed over $[0 \ 2\pi]$. Therefore, the probability density distribution of the phase of q in Equations 151 and 152 can also be considered to be evenly distributed over $[0 \ 2\pi]$. Accordingly, the following is established as a condition for providing fair data reception quality insofar as possible for Γ terminals in the same LOS environment in which only the phase of q differs.

Condition #12

When using a precoding hopping method with an N -slot time period (cycle), among the N slots in the time period (cycle), the poor reception points for s_1 are arranged to have an even distribution in terms of phase, and the poor reception points for s_2 are arranged to have an even distribution in terms of phase.

[0493]

The following describes an example of a precoding matrix in the precoding hopping method based on Condition #10 through Condition #12. Let $\alpha = 1.0$ in the precoding matrix in Equation 157.

(Example #5)

- 5 Let the number of slots N in the time period (cycle) be 8. In order to satisfy Condition #10 through Condition #12, precoding matrices for a precoding hopping method with an N = 8 time period (cycle) are provided as in the following equation.

[0494]

Math 170

10 Equation 160

$$\mathbf{F}[i] = \frac{1}{\sqrt{2}} \begin{pmatrix} e^{j0} & e^{j0} \\ e^{j\frac{i\pi}{4}} & e^{j\left(\frac{i\pi}{4} + \pi\right)} \end{pmatrix}$$

[0495]

- 15 Here, j is an imaginary unit, and i = 0, 1, ..., 7. Instead of Equation 160, Equation 161 may be provided (where λ and $\theta_{11}[i]$ do not change over time (though change may be allowed)).

[0496]

Math 171

Equation 161

20
$$\mathbf{F}[i] = \frac{1}{\sqrt{2}} \begin{pmatrix} e^{j\theta_{11}[i]} & e^{j(\theta_{11}[i] + \lambda)} \\ e^{j\left(\theta_{11}[i] + \frac{i\pi}{4}\right)} & e^{j\left(\theta_{11}[i] + \frac{i\pi}{4} + \lambda + \pi\right)} \end{pmatrix}$$

[0497]

Accordingly, the poor reception points for s1 and s2 become as in Figs. 31A and 31B. (In Figs. 31A and 31B, the horizontal axis is the real axis, and the vertical axis is the imaginary axis.) Instead of Equations 160 and 161, Equations 162 and 163 may be provided (where $i = 0, 1, \dots, 7$, and where λ and $\theta_{11}[i]$ do not change
 5 over time (though change may be allowed)).

[0498]

Math 172

Equation 162

$$\mathbf{F}[i] = \frac{1}{\sqrt{2}} \begin{pmatrix} e^{j0} & e^{j0} \\ e^{j\left(-\frac{i\pi}{4}\right)} & e^{j\left(-\frac{i\pi}{4} + \pi\right)} \end{pmatrix}$$

10

[0499]

Math 173

Equation 163

$$\mathbf{F}[i] = \frac{1}{\sqrt{2}} \begin{pmatrix} e^{j\theta_{11}[i]} & e^{j(\theta_{11}[i] + \lambda)} \\ e^{j\left(\theta_{11}[i] - \frac{i\pi}{4}\right)} & e^{j\left(\theta_{11}[i] - \frac{i\pi}{4} + \lambda + \pi\right)} \end{pmatrix}$$

15

[0500]

Next, the following is established as a condition, different from Condition #12, for providing fair data reception quality insofar as possible for Γ terminals in the same LOS environment in which only the phase of q differs.

20 Condition #13

When using a precoding hopping method with an N-slot time period (cycle), in addition to the condition

[0501]

Math 174

Equation 164

$$e^{j(\theta_{11}[x]-\theta_{21}[x])} \neq e^{j(\theta_{11}[y]-\theta_{21}[y]-\pi)} \quad \text{for } \forall x, \forall y \quad (x, y = 0, 1, \dots, N-1)$$

5 [0502]

the poor reception points for s1 and the poor reception points for s2 are arranged to be in an even distribution with respect to phase in the N slots in the time period (cycle).

[0503]

10 The following describes an example of a precoding matrix in the precoding hopping method based on Condition #10, Condition #11, and Condition #13. Let $\alpha = 1.0$ in the precoding matrix in Equation 157.

(Example #6)

Let the number of slots N in the time period (cycle) be 4. Precoding
15 matrices for a precoding hopping method with an N = 4 time period (cycle) are provided as in the following equation.

[0504]

Math 175

Equation 165

$$20 \quad \mathbf{F}[i] = \frac{1}{\sqrt{2}} \begin{pmatrix} e^{j0} & e^{j0} \\ e^{j\frac{i\pi}{4}} & e^{j\left(\frac{i\pi}{4}+\pi\right)} \end{pmatrix}$$

[0505]

Here, j is an imaginary unit, and i = 0, 1, 2, 3. Instead of Equation 165,
Equation 166 may be provided (where λ and $\theta_{11}[i]$ do not change over time (though
25 change may be allowed)).

[0506]

Math 176

Equation 166

$$\mathbf{F}[i] = \frac{1}{\sqrt{2}} \begin{pmatrix} e^{j\theta_{11}[i]} & e^{j(\theta_{11}[i]+\lambda)} \\ e^{j\left(\theta_{11}[i]+\frac{i\pi}{4}\right)} & e^{j\left(\theta_{11}[i]+\frac{i\pi}{4}+\lambda+\pi\right)} \end{pmatrix}$$

5

[0507]

Accordingly, the poor reception points for s1 and s2 become as in Fig. 32.

(In Fig. 32, the horizontal axis is the real axis, and the vertical axis is the imaginary axis.) Instead of Equations 165 and 166, Equations 167 and 168 may be provided

10 (where $i = 0, 1, 2, 3$, and where λ and $\theta_{11}[i]$ do not change over time (though change may be allowed)).

[0508]

Math 177

Equation 167

$$15 \quad \mathbf{F}[i] = \frac{1}{\sqrt{2}} \begin{pmatrix} e^{j0} & e^{j0} \\ e^{j\left(-\frac{i\pi}{4}\right)} & e^{j\left(-\frac{i\pi}{4}+\pi\right)} \end{pmatrix}$$

[0509]

Math 178

Equation 168

$$20 \quad \mathbf{F}[i] = \frac{1}{\sqrt{2}} \begin{pmatrix} e^{j\theta_{11}[i]} & e^{j(\theta_{11}[i]+\lambda)} \\ e^{j\left(\theta_{11}[i]-\frac{i\pi}{4}\right)} & e^{j\left(\theta_{11}[i]-\frac{i\pi}{4}+\lambda+\pi\right)} \end{pmatrix}$$

[0510]

Next, a precoding hopping method using a non-unitary matrix is described.

[0511]

Based on Equation 148, the precoding matrices presently under
5 consideration are represented as follows.

[0512]

Math 179

Equation 169

$$\mathbf{F}(p) = \frac{1}{\sqrt{\alpha^2 + 1}} \begin{pmatrix} e^{j\theta_{11}(p)} & \alpha \times e^{j(\theta_{11}(p) + \lambda)} \\ \alpha \times e^{j\theta_{21}(p)} & e^{j(\theta_{21}(p) + \lambda + \delta)} \end{pmatrix}$$

10

[0513]

Equations corresponding to Equations 151 and 152 are represented as
follows.

[0514]

15 Math 180

Equation 170

$$\begin{pmatrix} y_1(p) \\ y_2(p) \end{pmatrix} = \frac{1}{\sqrt{\alpha^2 + 1}} \begin{pmatrix} A e^{j0} & q \\ A e^{j0} & q \end{pmatrix} \begin{pmatrix} e^{j\theta_{11}(p)} & \alpha \times e^{j(\theta_{11}(p) + \lambda)} \\ \alpha \times e^{j\theta_{21}(p)} & e^{j(\theta_{21}(p) + \lambda + \delta)} \end{pmatrix} \begin{pmatrix} s1(p) \\ s2(p) \end{pmatrix} + \mathbf{n}(p)$$

[0515]

20 Math 181

Equation 171

$$\begin{pmatrix} y_1(p) \\ y_2(p) \end{pmatrix} = \frac{1}{\sqrt{\alpha^2 + 1}} \begin{pmatrix} e^{j0} \\ e^{j0} \end{pmatrix} \left(A e^{j0} + q \begin{pmatrix} e^{j\theta_{11}(p)} & \alpha \times e^{j(\theta_{11}(p) + \lambda)} \\ \alpha \times e^{j\theta_{21}(p)} & e^{j(\theta_{21}(p) + \lambda + \delta)} \end{pmatrix} \begin{pmatrix} s1(p) \\ s2(p) \end{pmatrix} \right) + \mathbf{n}(p)$$

[0516]

In this case, there are two q at which the minimum value $d_{\min}^2$ of the Euclidian distance between a received signal point and a received candidate signal point is zero.

[0517]

In Equation 171, when $s1(p)$ does not exist:

[0518]

10 Math 182

Equation 172

$$q = -\frac{A}{\alpha} e^{j(\theta_{11}(p) - \theta_{21}(p))}$$

[0519]

In Equation 171, when $s2(p)$ does not exist:

15 [0520]

Math 183

Equation 173

$$q = -A\alpha e^{j(\theta_{11}(p) - \theta_{21}(p) - \delta)}$$

20 [0521]

In the precoding hopping method for an N-slot time period (cycle), by referring to Equation 169, N varieties of the precoding matrix $F[i]$ are represented as follows.

[0522]

5 Math 184

Equation 174

$$F[i] = \frac{1}{\sqrt{\alpha^2 + 1}} \begin{pmatrix} e^{j\theta_{11}[i]} & \alpha \times e^{j(\theta_{11}[i] + \lambda)} \\ \alpha \times e^{j\theta_{21}[i]} & e^{j(\theta_{21}[i] + \lambda + \delta)} \end{pmatrix}$$

[0523]

10 In this equation, let α and δ not change over time. At this point, based on Equations 34 and 35, design conditions such as the following are provided for the precoding matrices for precoding hopping.

[0524]

Math 185

15 Condition #14

Equation 175

$$e^{j(\theta_{11}[x] - \theta_{21}[x])} \neq e^{j(\theta_{11}[y] - \theta_{21}[y])}$$

for $\forall x, \forall y \quad (x \neq y; x, y = 0, 1, \dots, N-1)$

[0525]

20 Math 186

Condition #15

Equation 176

$$e^{j(\theta_{11}[x]-\theta_{21}[x]-\delta)} \neq e^{j(\theta_{11}[y]-\theta_{21}[y]-\delta)}$$

for $\forall x, \forall y$ ($x \neq y$; $x, y = 0, 1, \dots, N-1$)

[0526]

(Example #7)

- 5 Let $\alpha = 1.0$ in the precoding matrix in Equation 174. Let the number of slots N in the time period (cycle) be 16. In order to satisfy Condition #12, Condition #14, and Condition #15, precoding matrices for a precoding hopping method with an $N = 16$ time period (cycle) are provided as in the following equations.

[0527]

- 10 For $i = 0, 1, \dots, 7$:

[0528]

Math 187

Equation 177

$$\mathbf{F}[i] = \frac{1}{\sqrt{2}} \begin{pmatrix} e^{j0} & e^{j0} \\ e^{j\frac{i\pi}{4}} & e^{j\left(\frac{i\pi}{4} + \frac{7\pi}{8}\right)} \end{pmatrix}$$

15

[0529]

For $i = 8, 9, \dots, 15$:

[0530]

Math 188

- 20 Equation 178

$$\mathbf{F}[i] = \frac{1}{\sqrt{2}} \begin{pmatrix} e^{j\frac{i\pi}{4}} & e^{j\left(\frac{i\pi}{4} + \frac{7\pi}{8}\right)} \\ e^{j0} & e^{j0} \end{pmatrix}$$

[0531]

Furthermore, a precoding matrix that differs from Equations 177 and 178 can be provided as follows.

[0532]

5 For $i = 0, 1, \dots, 7$:

[0533]

Math 189

Equation 179

$$\mathbf{F}[i] = \frac{1}{\sqrt{2}} \begin{pmatrix} e^{j\theta_{11}[i]} & e^{j(\theta_{11}[i]+\lambda)} \\ e^{j(\theta_{11}[i]+\frac{i\pi}{4})} & e^{j(\theta_{11}[i]+\frac{i\pi}{4}+\lambda+\frac{7\pi}{8})} \end{pmatrix}$$

10

[0534]

For $i = 8, 9, \dots, 15$:

[0535]

Math 190

15 Equation 180

$$\mathbf{F}[i] = \frac{1}{\sqrt{2}} \begin{pmatrix} e^{j(\theta_{11}[i]+\frac{i\pi}{4})} & e^{j(\theta_{11}[i]+\frac{i\pi}{4}+\lambda+\frac{7\pi}{8})} \\ e^{j\theta_{11}[i]} & e^{j(\theta_{11}[i]+\lambda)} \end{pmatrix}$$

[0536]

Accordingly, the poor reception points for s1 and s2 become as in Figs. 33A and 33B.

20

[0537]

(In Figs. 33A and 33B, the horizontal axis is the real axis, and the vertical axis is the imaginary axis.) Instead of Equations 177 and 178, and Equations 179 and 180, precoding matrices may be provided as below.

[0538]

5 For $i = 0, 1, \dots, 7$:

[0539]

Math 191

Equation 181

$$\mathbf{F}[i] = \frac{1}{\sqrt{2}} \begin{pmatrix} e^{j0} & e^{j0} \\ e^{j\left(-\frac{i\pi}{4}\right)} & e^{j\left(-\frac{i\pi}{4} + \frac{7\pi}{8}\right)} \end{pmatrix}$$

10

[0540]

For $i = 8, 9, \dots, 15$:

[0541]

Math 192

15 Equation 182

$$\mathbf{F}[i] = \frac{1}{\sqrt{2}} \begin{pmatrix} e^{j\left(-\frac{i\pi}{4}\right)} & e^{j\left(-\frac{i\pi}{4} + \frac{7\pi}{8}\right)} \\ e^{j0} & e^{j0} \end{pmatrix}$$

[0542]

or

20 For $i = 0, 1, \dots, 7$:

[0543]

Math 193

Equation 183

$$\mathbf{F}[i] = \frac{1}{\sqrt{2}} \begin{pmatrix} e^{j\theta_{11}[i]} & e^{j(\theta_{11}[i]+\lambda)} \\ e^{j(\theta_{11}[i]-\frac{i\pi}{4})} & e^{j(\theta_{11}[i]-\frac{i\pi}{4}+\lambda+\frac{7\pi}{8})} \end{pmatrix}$$

[0544]

5 For i = 8, 9, ..., 15:

[0545]

Math 194

Equation 184

$$\mathbf{F}[i] = \frac{1}{\sqrt{2}} \begin{pmatrix} e^{j(\theta_{11}[i]-\frac{i\pi}{4})} & e^{j(\theta_{11}[i]-\frac{i\pi}{4}+\lambda+\frac{7\pi}{8})} \\ e^{j\theta_{11}[i]} & e^{j(\theta_{11}[i]+\lambda)} \end{pmatrix}$$

10

[0546]

(In Equations 177–184, $7\pi/8$ may be changed to $-7\pi/8$.)

Next, the following is established as a condition, different from Condition #12, for providing fair data reception quality insofar as possible for Γ terminals in the same LOS environment in which only the phase of q differs.

15

Condition #16

When using a precoding hopping method with an N-slot time period (cycle), the following condition is set:

[0547]

20 Math 195

Equation 185

$$e^{j(\theta_{11}[x]-\theta_{21}[x])} \neq e^{j(\theta_{11}[y]-\theta_{21}[y]-\delta)} \quad \text{for } \forall x, \forall y \quad (x, y = 0, 1, \dots, N-1)$$

[0548]

and the poor reception points for s1 and the poor reception points for s2 are
 5 arranged to be in an even distribution with respect to phase in the N slots in the time period (cycle).

[0549]

The following describes an example of a precoding matrix in the precoding hopping method based on Condition #14, Condition #15, and Condition #16. Let $\alpha =$
 10 1.0 in the precoding matrix in Equation 174.

(Example #8)

Let the number of slots N in the time period (cycle) be 8. Precoding matrices for a precoding hopping method with an N = 8 time period (cycle) are provided as in the following equation.

15 [0550]

Math 196

Equation 186

$$\mathbf{F}[i] = \frac{1}{\sqrt{2}} \begin{pmatrix} e^{j0} & e^{j0} \\ e^{j\frac{i\pi}{4}} & e^{j\left(\frac{i\pi}{4} + \frac{7\pi}{8}\right)} \end{pmatrix}$$

20 [0551]

Here, $i = 0, 1, \dots, 7$.

[0552]

Furthermore, a precoding matrix that differs from Equation 186 can be provided as follows (where $i = 0, 1, \dots, 7$, and where λ and $\theta_{11}[i]$ do not change over
 25 time (though change may be allowed)).

[0553]

Math 197

Equation 187

$$\mathbf{F}[i] = \frac{1}{\sqrt{2}} \begin{pmatrix} e^{j\theta_{11}[i]} & e^{j(\theta_{11}[i]+\lambda)} \\ e^{j(\theta_{11}[i]+\frac{i\pi}{4})} & e^{j(\theta_{11}[i]+\frac{i\pi}{4}+\lambda+\frac{7\pi}{8})} \end{pmatrix}$$

5

[0554]

Accordingly, the poor reception points for s1 and s2 become as in Fig. 34.

Instead of Equations 186 and 187, precoding matrices may be provided as follows

(where $i = 0, 1, \dots, 7$, and where λ and $\theta_{11}[i]$ do not change over time (though

10 change may be allowed)).

[0555]

Math 198

Equation 188

$$\mathbf{F}[i] = \frac{1}{\sqrt{2}} \begin{pmatrix} e^{j0} & e^{j0} \\ e^{j(-\frac{i\pi}{4})} & e^{j(-\frac{i\pi}{4}+\frac{7\pi}{8})} \end{pmatrix}$$

15

[0556]

or

[0557]

Math 199

20 Equation 189

$$\mathbf{F}[i] = \frac{1}{\sqrt{2}} \begin{pmatrix} e^{j\theta_{11}[i]} & e^{j(\theta_{11}[i]+\lambda)} \\ e^{j(\theta_{11}[i]-\frac{i\pi}{4})} & e^{j(\theta_{11}[i]-\frac{i\pi}{4}+\lambda+\frac{7\pi}{8})} \end{pmatrix}$$

[0558]

(In Equations 186–189, $7\pi/8$ may be changed to $-7\pi/8$.)

Next, in the precoding matrix of Equation 174, a precoding hopping method
 5 that differs from Example #7 and Example #8 by letting $\alpha \neq 1$, and by taking into
 consideration the distance in the complex plane between poor reception points, is
 examined.

[0559]

In this case, the precoding hopping method for an N-slot time period (cycle)
 10 of Equation 174 is used, and from Condition #14, in all of the Γ terminals, there is
 one slot or less having poor reception points for s_1 among the N slots in a time
 period (cycle). Accordingly, the log-likelihood ratio for bits transmitted by $s_1(p)$ can
 be obtained for at least $N - 1$ slots. Similarly, from Condition #15, in all of the Γ
 terminals, there is one slot or less having poor reception points for s_2 among the N
 15 slots in a time period (cycle). Accordingly, the log-likelihood ratio for bits
 transmitted by $s_2(p)$ can be obtained for at least $N - 1$ slots.

[0560]

Therefore, it is clear that a larger value for N in the N-slot time period
 (cycle) increases the number of slots in which the log-likelihood ratio can be
 20 obtained.

[0561]

Incidentally, since the influence of scattered wave components is also
 present in an actual channel model, it is considered that when the number of slots N
 in the time period (cycle) is fixed, there is a possibility of improved data reception
 25 quality if the minimum distance in the complex plane between poor reception points
 is as large as possible. Accordingly, in the context of Example #7 and Example #8,
 precoding hopping methods in which $\alpha \neq 1$ and which improve on Example #7 and

Example #8 are considered. The precoding method that improves on Example #8 is easier to understand and is therefore described first.

(Example #9)

From Equation 186, the precoding matrices in an $N = 8$ time period (cycle) precoding hopping method that improves on Example #8 are provided in the following equation.

[0562]

Math 200

Equation 190

$$\mathbf{F}[i] = \frac{1}{\sqrt{\alpha^2 + 1}} \begin{pmatrix} e^{j0} & \alpha \times e^{j0} \\ \alpha \times e^{j\frac{i\pi}{4}} & e^{j\left(\frac{i\pi}{4} + \frac{7\pi}{8}\right)} \end{pmatrix}$$

[0563]

Here, $i = 0, 1, \dots, 7$. Furthermore, precoding matrices that differ from Equation 190 can be provided as follows (where $i = 0, 1, \dots, 7$, and where λ and $\theta_{11}[i]$ do not change over time (though change may be allowed)).

[0564]

Math 201

Equation 191

$$\mathbf{F}[i] = \frac{1}{\sqrt{\alpha^2 + 1}} \begin{pmatrix} e^{j\theta_{11}[i]} & \alpha \times e^{j(\theta_{11}[i] + \lambda)} \\ \alpha \times e^{j\left(\theta_{11}[i] + \frac{i\pi}{4}\right)} & e^{j\left(\theta_{11}[i] + \frac{i\pi}{4} + \lambda + \frac{7\pi}{8}\right)} \end{pmatrix}$$

20

[0565]

or

[0566]

Math 202

Equation 192

$$\mathbf{F}[i] = \frac{1}{\sqrt{\alpha^2 + 1}} \begin{pmatrix} e^{j0} & \alpha \times e^{j0} \\ \alpha \times e^{j\left(-\frac{i\pi}{4}\right)} & e^{j\left(-\frac{i\pi}{4} + \frac{7\pi}{8}\right)} \end{pmatrix}$$

5 [0567]

or

[0568]

Math 203

Equation 193

$$10 \quad \mathbf{F}[i] = \frac{1}{\sqrt{\alpha^2 + 1}} \begin{pmatrix} e^{j\theta_{11}[i]} & \alpha \times e^{j(\theta_{11}[i] + \lambda)} \\ \alpha \times e^{j\left(\theta_{11}[i] - \frac{i\pi}{4}\right)} & e^{j\left(\theta_{11}[i] - \frac{i\pi}{4} + \lambda + \frac{7\pi}{8}\right)} \end{pmatrix}$$

[0569]

or

[0570]

15 Math 204

Equation 194

$$\mathbf{F}[i] = \frac{1}{\sqrt{\alpha^2 + 1}} \begin{pmatrix} e^{j0} & \alpha \times e^{j0} \\ \alpha \times e^{j\frac{i\pi}{4}} & e^{j\left(\frac{i\pi}{4} - \frac{7\pi}{8}\right)} \end{pmatrix}$$

[0571]

20

or

[0572]

Math 205

Equation 195

$$\mathbf{F}[i] = \frac{1}{\sqrt{\alpha^2 + 1}} \begin{pmatrix} e^{j\theta_{11}[i]} & \alpha \times e^{j(\theta_{11}[i] + \lambda)} \\ \alpha \times e^{j(\theta_{11}[i] + \frac{i\pi}{4})} & e^{j(\theta_{11}[i] + \frac{i\pi}{4} + \lambda - \frac{7\pi}{8})} \end{pmatrix}$$

[0573]

5 or

[0574]

Math 206

Equation 196

$$\mathbf{F}[i] = \frac{1}{\sqrt{\alpha^2 + 1}} \begin{pmatrix} e^{j0} & \alpha \times e^{j0} \\ \alpha \times e^{j(-\frac{i\pi}{4})} & e^{j(-\frac{i\pi}{4} - \frac{7\pi}{8})} \end{pmatrix}$$

10

[0575]

or

[0576]

Math 207

15 Equation 197

$$\mathbf{F}[i] = \frac{1}{\sqrt{\alpha^2 + 1}} \begin{pmatrix} e^{j\theta_{11}[i]} & \alpha \times e^{j(\theta_{11}[i] + \lambda)} \\ \alpha \times e^{j(\theta_{11}[i] - \frac{i\pi}{4})} & e^{j(\theta_{11}[i] - \frac{i\pi}{4} + \lambda - \frac{7\pi}{8})} \end{pmatrix}$$

[0577]

Therefore, the poor reception points for s1 and s2 are represented as in Fig.

20 35A when $\alpha < 1.0$ and as in Fig. 35B when $\alpha > 1.0$.

[0578]

(i) When $\alpha < 1.0$

When $\alpha < 1.0$, the minimum distance in the complex plane between poor reception points is represented as $\min\{d_{\#1,\#2}, d_{\#1,\#3}\}$ when focusing on the distance ($d_{\#1,\#2}$) between poor reception points #1 and #2 and the distance ($d_{\#1,\#3}$) between poor reception points #1 and #3. In this case, the relationship between α and $d_{\#1,\#2}$ and between α and $d_{\#1,\#3}$ is shown in Fig. 36. The α which makes $\min\{d_{\#1,\#2}, d_{\#1,\#3}\}$ the largest is as follows.

[0579]

Math 208

Equation 198

$$\alpha = \frac{1}{\sqrt{\cos\left(\frac{\pi}{8}\right) + \sqrt{3} \sin\left(\frac{\pi}{8}\right)}} \approx 0.7938$$

[0580]

The $\min\{d_{\#1,\#2}, d_{\#1,\#3}\}$ in this case is as follows.

[0581]

15 Math 209

Equation 199

$$\min\{d_{\#1,\#2}, d_{\#1,\#3}\} = \frac{2A \sin\left(\frac{\pi}{8}\right)}{\sqrt{\cos\left(\frac{\pi}{8}\right) + \sqrt{3} \sin\left(\frac{\pi}{8}\right)}} \approx 0.6076A$$

[0582]

20 Therefore, the precoding method using the value of α in Equation 198 for Equations 190–197 is effective. Setting the value of α as in Equation 198 is one

appropriate method for obtaining excellent data reception quality. Setting α to be a value near Equation 198, however, may similarly allow for excellent data reception quality. Accordingly, the value to which α is set is not limited to Equation 198.

[0583]

5 (ii) When $\alpha > 1.0$

When $\alpha > 1.0$, the minimum distance in the complex plane between poor reception points is represented as $\min\{d_{\#4,\#5}, d_{\#4,\#6}\}$ when focusing on the distance ($d_{\#4,\#5}$) between poor reception points #4 and #5 and the distance ($d_{\#4,\#6}$) between poor reception points #4 and #6. In this case, the relationship between α and $d_{\#4,\#5}$ and between α and $d_{\#4,\#6}$ is shown in Fig. 37. The α which makes $\min\{d_{\#4,\#5}, d_{\#4,\#6}\}$ the largest is as follows.

[0584]

Math 210

Equation 200

$$15 \quad \alpha = \sqrt{\cos\left(\frac{\pi}{8}\right) + \sqrt{3} \sin\left(\frac{\pi}{8}\right)} \\ \approx 1.2596$$

[0585]

The $\min\{d_{\#4,\#5}, d_{\#4,\#6}\}$ in this case is as follows.

[0586]

20 Math 211

Equation 201

$$\min\{d_{\#4,\#5}, d_{\#4,\#6}\} = \frac{2A \sin\left(\frac{\pi}{8}\right)}{\sqrt{\cos\left(\frac{\pi}{8}\right) + \sqrt{3} \sin\left(\frac{\pi}{8}\right)}} \\ \approx 0.6076A$$

[0587]

Therefore, the precoding method using the value of α in Equation 200 for Equations 190–197 is effective. Setting the value of α as in Equation 200 is one
 5 appropriate method for obtaining excellent data reception quality. Setting α to be a value near Equation 200, however, may similarly allow for excellent data reception quality. Accordingly, the value to which α is set is not limited to Equation 200.

(Example #10)

Based on consideration of Example #9, the precoding matrices in an $N = 16$
 10 time period (cycle) precoding hopping method that improves on Example #7 are provided in the following equations (where λ and $\theta_{11}[i]$ do not change over time (though change may be allowed)).

[0588]

For $i = 0, 1, \dots, 7$:

15 [0589]

Math 212

Equation 202

$$\mathbf{F}[i] = \frac{1}{\sqrt{\alpha^2 + 1}} \begin{pmatrix} e^{j0} & \alpha \times e^{j0} \\ \alpha \times e^{j\frac{i\pi}{4}} & e^{j\left(\frac{i\pi}{4} + \frac{7\pi}{8}\right)} \end{pmatrix}$$

20 [0590]

For $i = 8, 9, \dots, 15$:

[0591]

Math 213

Equation 203

$$\mathbf{F}[i] = \frac{1}{\sqrt{\alpha^2 + 1}} \begin{pmatrix} \alpha \times e^{j\frac{i\pi}{4}} & e^{j\left(\frac{i\pi}{4} + \frac{7\pi}{8}\right)} \\ e^{j0} & \alpha \times e^{j0} \end{pmatrix}$$

[0592]

or

5 For i = 0, 1, ..., 7:

[0593]

Math 214

Equation 204

$$\mathbf{F}[i] = \frac{1}{\sqrt{\alpha^2 + 1}} \begin{pmatrix} e^{j\theta_{11}[i]} & \alpha \times e^{j(\theta_{11}[i] + \lambda)} \\ \alpha \times e^{j\left(\theta_{11}[i] + \frac{i\pi}{4}\right)} & e^{j\left(\theta_{11}[i] + \frac{i\pi}{4} + \lambda + \frac{7\pi}{8}\right)} \end{pmatrix}$$

10

[0594]

For i = 8, 9, ..., 15:

[0595]

Math 215

15 Equation 205

$$\mathbf{F}[i] = \frac{1}{\sqrt{\alpha^2 + 1}} \begin{pmatrix} \alpha \times e^{j\left(\theta_{11}[i] + \frac{i\pi}{4}\right)} & e^{j\left(\theta_{11}[i] + \frac{i\pi}{4} + \lambda + \frac{7\pi}{8}\right)} \\ e^{j\theta_{11}[i]} & \alpha \times e^{j(\theta_{11}[i] + \lambda)} \end{pmatrix}$$

[0596]

or

20 For i = 0, 1, ..., 7:

[0597]

Math 216

Equation 206

$$\mathbf{F}[i] = \frac{1}{\sqrt{\alpha^2 + 1}} \begin{pmatrix} e^{j0} & \alpha \times e^{j0} \\ \alpha \times e^{j\left(-\frac{i\pi}{4}\right)} & e^{j\left(-\frac{i\pi}{4} + \frac{7\pi}{8}\right)} \end{pmatrix}$$

[0598]

5 For i = 8, 9, ..., 15:

[0599]

Math 217

Equation 207

$$\mathbf{F}[i] = \frac{1}{\sqrt{\alpha^2 + 1}} \begin{pmatrix} \alpha \times e^{j\left(-\frac{i\pi}{4}\right)} & e^{j\left(-\frac{i\pi}{4} + \frac{7\pi}{8}\right)} \\ e^{j0} & \alpha \times e^{j0} \end{pmatrix}$$

10

[0600]

or

For i = 0, 1, ..., 7:

[0601]

15 Math 218

Equation 208

$$\mathbf{F}[i] = \frac{1}{\sqrt{\alpha^2 + 1}} \begin{pmatrix} e^{j\theta_{11}[i]} & \alpha \times e^{j(\theta_{11}[i] + \lambda)} \\ \alpha \times e^{j\left(\theta_{11}[i] - \frac{i\pi}{4}\right)} & e^{j\left(\theta_{11}[i] - \frac{i\pi}{4} + \lambda + \frac{7\pi}{8}\right)} \end{pmatrix}$$

[0602]

20 For i = 8, 9, ..., 15:

[0603]

Math 219

Equation 209

$$\mathbf{F}[i] = \frac{1}{\sqrt{\alpha^2 + 1}} \begin{pmatrix} \alpha \times e^{j\left(\theta_{11}[i] - \frac{i\pi}{4}\right)} & e^{j\left(\theta_{11}[i] - \frac{i\pi}{4} + \lambda + \frac{7\pi}{8}\right)} \\ e^{j\theta_{11}[i]} & \alpha \times e^{j(\theta_{11}[i] + \lambda)} \end{pmatrix}$$

[0604]

5 or

For i = 0, 1, ..., 7:

[0605]

Math 220

Equation 210

$$10 \quad \mathbf{F}[i] = \frac{1}{\sqrt{\alpha^2 + 1}} \begin{pmatrix} e^{j0} & \alpha \times e^{j0} \\ \alpha \times e^{j\frac{i\pi}{4}} & e^{j\left(\frac{i\pi}{4} - \frac{7\pi}{8}\right)} \end{pmatrix}$$

[0606]

For i = 8, 9, ..., 15:

[0607]

15 Math 221

Equation 211

$$\mathbf{F}[i] = \frac{1}{\sqrt{\alpha^2 + 1}} \begin{pmatrix} \alpha \times e^{j\frac{i\pi}{4}} & e^{j\left(\frac{i\pi}{4} - \frac{7\pi}{8}\right)} \\ e^{j0} & \alpha \times e^{j0} \end{pmatrix}$$

[0608]

20 or

For i = 0, 1, ..., 7:

[0609]

Math 222

Equation 212

$$\mathbf{F}[i] = \frac{1}{\sqrt{\alpha^2 + 1}} \begin{pmatrix} e^{j\theta_{11}[i]} & \alpha \times e^{j(\theta_{11}[i] + \lambda)} \\ \alpha \times e^{j(\theta_{11}[i] + \frac{i\pi}{4})} & e^{j(\theta_{11}[i] + \frac{i\pi}{4} + \lambda - \frac{7\pi}{8})} \end{pmatrix}$$

5 [0610]

For i = 8, 9, ..., 15:

[0611]

Math 223

Equation 213

$$10 \quad \mathbf{F}[i] = \frac{1}{\sqrt{\alpha^2 + 1}} \begin{pmatrix} \alpha \times e^{j(\theta_{11}[i] + \frac{i\pi}{4})} & e^{j(\theta_{11}[i] + \frac{i\pi}{4} + \lambda - \frac{7\pi}{8})} \\ e^{j\theta_{11}[i]} & \alpha \times e^{j(\theta_{11}[i] + \lambda)} \end{pmatrix}$$

[0612]

or

For i = 0, 1, ..., 7:

15 [0613]

Math 224

Equation 214

$$\mathbf{F}[i] = \frac{1}{\sqrt{\alpha^2 + 1}} \begin{pmatrix} e^{j0} & \alpha \times e^{j0} \\ \alpha \times e^{j(\frac{-i\pi}{4})} & e^{j(\frac{-i\pi}{4} - \frac{7\pi}{8})} \end{pmatrix}$$

20 [0614]

For i = 8, 9, ..., 15:

[0615]

Math 225

Equation 215

$$\mathbf{F}[i] = \frac{1}{\sqrt{\alpha^2 + 1}} \begin{pmatrix} \alpha \times e^{j\left(-\frac{i\pi}{4}\right)} & e^{j\left(-\frac{i\pi}{4} - \frac{7\pi}{8}\right)} \\ e^{j0} & \alpha \times e^{j0} \end{pmatrix}$$

5 [0616]

or

For i = 0, 1, ..., 7:

[0617]

Math 226

10 Equation 216

$$\mathbf{F}[i] = \frac{1}{\sqrt{\alpha^2 + 1}} \begin{pmatrix} e^{j\theta_{11}[i]} & \alpha \times e^{j(\theta_{11}[i] + \lambda)} \\ \alpha \times e^{j\left(\theta_{11}[i] - \frac{i\pi}{4}\right)} & e^{j\left(\theta_{11}[i] - \frac{i\pi}{4} + \lambda - \frac{7\pi}{8}\right)} \end{pmatrix}$$

[0618]

For i = 8, 9, ..., 15:

15 [0619]

Math 227

Equation 217

$$\mathbf{F}[i] = \frac{1}{\sqrt{\alpha^2 + 1}} \begin{pmatrix} \alpha \times e^{j\left(\theta_{11}[i] - \frac{i\pi}{4}\right)} & e^{j\left(\theta_{11}[i] - \frac{i\pi}{4} + \lambda - \frac{7\pi}{8}\right)} \\ e^{j\theta_{11}[i]} & \alpha \times e^{j(\theta_{11}[i] + \lambda)} \end{pmatrix}$$

20 [0620]

The value of α in Equation 198 and in Equation 200 is appropriate for obtaining excellent data reception quality. The poor reception points for s1 are

represented as in Figs. 38A and 38B when $\alpha < 1.0$ and as in Figs. 39A and 39B when $\alpha > 1.0$.

[0621]

In the present embodiment, the method of structuring N different precoding
 5 matrices for a precoding hopping method with an N-slot time period (cycle) has
 been described. In this case, as the N different precoding matrices, $F[0]$, $F[1]$, $F[2]$,
 ..., $F[N - 2]$, $F[N - 1]$ are prepared. In the present embodiment, an example of a
 single carrier transmission method has been described, and therefore the case of
 arranging symbols in the order $F[0]$, $F[1]$, $F[2]$, ..., $F[N - 2]$, $F[N - 1]$ in the time
 10 domain (or the frequency domain) has been described. The present invention is not,
 however, limited in this way, and the N different precoding matrices $F[0]$, $F[1]$, $F[2]$,
 ..., $F[N - 2]$, $F[N - 1]$ generated in the present embodiment may be adapted to a
 multi-carrier transmission method such as an OFDM transmission method or the like.
 As in Embodiment 1, as a method of adaption in this case, precoding weights may
 15 be changed by arranging symbols in the frequency domain and in the frequency-time
 domain. Note that a precoding hopping method with an N-slot time period (cycle)
 has been described, but the same advantageous effects may be obtained by randomly
 using N different precoding matrices. In other words, the N different precoding
 matrices do not necessarily need to be used in a regular period (cycle).

20 [0622]

Examples #5 through #10 have been shown based on Conditions #10
 through #16. However, in order to achieve a precoding matrix hopping method with
 a longer period (cycle), the period (cycle) for hopping between precoding matrices
 may be lengthened by, for example, selecting a plurality of examples from Examples
 25 #5 through #10 and using the precoding matrices indicated in the selected examples.
 For example, a precoding matrix hopping method with a longer period (cycle) may
 be achieved by using the precoding matrices indicated in Example #7 and the
 precoding matrices indicated in Example #10. In this case, Conditions #10 through

#16 are not necessarily observed. (In Equation 158 of Condition #10, Equation 159 of Condition #11, Equation 164 of Condition #13, Equation 175 of Condition #14, and Equation 176 of Condition #15, it becomes important for providing excellent reception quality for the conditions “all x and all y” to be “existing x and existing y”.) When viewed from a different perspective, in the precoding matrix hopping method over an N-slot period (cycle) (where N is a large natural number), the probability of providing excellent reception quality increases when the precoding matrices of one of Examples #5 through #10 are included.

(Embodiment 7)

10 The present embodiment describes the structure of a reception device for receiving modulated signals transmitted by a transmission method that regularly hops between precoding matrices as described in Embodiments 1–6.

[0623]

15 In Embodiment 1, the following method has been described. A transmission device that transmits modulated signals, using a transmission method that regularly hops between precoding matrices, transmits information regarding the precoding matrices. Based on this information, a reception device obtains information on the regular precoding matrix hopping used in the transmitted frames, decodes the precoding, performs detection, obtains the log-likelihood ratio for the transmitted bits, and subsequently performs error correction decoding.

[0624]

 The present embodiment describes the structure of a reception device, and a method of hopping between precoding matrices, that differ from the above structure and method.

25 [0625]

 Fig. 40 is an example of the structure of a transmission device in the present embodiment. Elements that operate in a similar way to Fig. 3 bear the same reference signs. An encoder group (4002) receives transmission bits (4001) as input.

The encoder group (4002), as described in Embodiment 1, includes a plurality of encoders for error correction coding, and based on the frame structure signal 313, a certain number of encoders operate, such as one encoder, two encoders, or four encoders.

5 [0626]

When one encoder operates, the transmission bits (4001) are encoded to yield encoded transmission bits. The encoded transmission bits are allocated into two parts, and the encoder group (4002) outputs allocated bits (4003A) and allocated bits (4003B).

10 [0627]

When two encoders operate, the transmission bits (4001) are divided in two (referred to as divided bits A and B). The first encoder receives the divided bits A as input, encodes the divided bits A, and outputs the encoded bits as allocated bits (4003A). The second encoder receives the divided bits B as input, encodes the divided bits B, and outputs the encoded bits as allocated bits (4003B).

[0628]

When four encoders operate, the transmission bits (4001) are divided in four (referred to as divided bits A, B, C, and D). The first encoder receives the divided bits A as input, encodes the divided bits A, and outputs the encoded bits A. The second encoder receives the divided bits B as input, encodes the divided bits B, and outputs the encoded bits B. The third encoder receives the divided bits C as input, encodes the divided bits C, and outputs the encoded bits C. The fourth encoder receives the divided bits D as input, encodes the divided bits D, and outputs the encoded bits D. The encoded bits A, B, C, and D are divided into allocated bits (4003A) and allocated bits (4003B).

[0629]

The transmission device supports a transmission method such as, for example, the following Table 1 (Table 1A and Table 1B).

[0630]

Table 1A

Number of modulated transmission signals (number of transmit antennas)	Modulation method	Number of encoders	Error correction coding method	Transmission information	Precoding matrix hopping method
1	QPSK	1	A	00000000	-
			B	00000001	-
			C	00000010	-
	16QAM	1	A	00000011	-
			B	00000100	-
			C	00000101	-
	64QAM	1	A	00000110	-
			B	00000111	-
			C	00001000	-
	256QAM	1	A	00001001	-
			B	00001010	-
			C	00001011	-
	1024QAM	1	A	00001100	-
			B	00001101	-
			C	00001110	-

[0631]

5

Table 1B

Number of modulated transmission signals (number of transmit antennas)	Modulation method	Number of encoders	Error correction coding method	Transmission information	Precoding matrix hopping method
2	#1: QPSK, #2: QPSK	1	A	00001111	D
			B	00010000	D
			C	00010001	D
		2	A	00010010	E
			B	00010011	E
			C	00010100	E
	#1: QPSK, #2: 16QAM	1	A	00010101	D
			B	00010110	D

	2	C	00010111	D
		A	00011000	E
		B	00011001	E
		C	00011010	E
#1: 16QAM, #2: 16QAM	1	A	00011011	D
		B	00011100	D
		C	00011101	D
	2	A	00011110	E
		B	00011111	E
		C	00100000	E
#1: 16QAM, #2: 64QAM	1	A	00100001	D
		B	00100010	D
		C	00100011	D
	2	A	00100100	E
		B	00100101	E
		C	00100110	E
#1: 64QAM, #2: 64QAM	1	A	00100111	F
		B	00101000	F
		C	00101001	F
	2	A	00101010	G
		B	00101011	G
		C	00101100	G
#1: 64QAM, #2: 256QAM	1	A	00101101	F
		B	00101110	F
		C	00101111	F
	2	A	00110000	G
		B	00110001	G
		C	00110010	G
#1: 256QAM, #2: 256QAM	1	A	00110011	F
		B	00110100	F
		C	00110101	F
	2	A	00110110	G
		B	00110111	G
		C	00111000	G
	4	A	00111001	H
		B	00111010	H
		C	00111011	H
#1: 256QAM, #2: 1024QAM	1	A	00111100	F
		B	00111101	F
		C	00111110	F
	2	A	00111111	G
		B	01000000	G
		C	01000001	G
	4	A	01000010	H
		B	01000011	H

	#1: 1024QAM, #2: 1024QAM	1	C	01000100	H
			A	01000101	F
			B	01000110	F
		2	C	01000111	F
			A	01001000	G
			B	01001001	G
			C	01001010	G
		4	A	01001011	H
			B	01001100	H
			C	01001101	H

[0632]

As shown in Table 1, transmission of a one-stream signal and transmission of a two-stream signal are supported as the number of transmission signals (number of transmit antennas). Furthermore, QPSK, 16QAM, 64QAM, 256QAM, and 1024QAM are supported as the modulation method. In particular, when the number of transmission signals is two, it is possible to set separate modulation methods for stream #1 and stream #2. For example, “#1: 256QAM, #2: 1024QAM” in Table 1 indicates that “the modulation method of stream #1 is 256QAM, and the modulation method of stream #2 is 1024QAM” (other entries in the table are similarly expressed). Three types of error correction coding methods, A, B, and C, are supported. In this case, A, B, and C may all be different coding methods. A, B, and C may also be different coding rates, and A, B, and C may be coding methods with different block sizes.

[0633]

The pieces of transmission information in Table 1 are allocated to modes that define a “number of transmission signals”, “modulation method”, “number of encoders”, and “error correction coding method”. Accordingly, in the case of “number of transmission signals: 2”, “modulation method: #1: 1024QAM, #2: 1024QAM”, “number of encoders: 4”, and “error correction coding method: C”, for example, the transmission information is set to 01001101. In the frame, the transmission device transmits the transmission information and the transmission data.

When transmitting the transmission data, in particular when the “number of transmission signals” is two, a “precoding matrix hopping method” is used in accordance with Table 1. In Table 1, five types of the “precoding matrix hopping method”, D, E, F, G, and H, are prepared. The precoding matrix hopping method is set to one of these five types in accordance with Table 1. The following, for example, are ways of implementing the five different types.

- Prepare five different precoding matrices.
- Use five different types of periods (cycles), for example a four-slot period (cycle) for D, an eight-slot period (cycle) for E, ...
- Use both different precoding matrices and different periods (cycles).

[0634]

Fig. 41 shows an example of a frame structure of a modulated signal transmitted by the transmission device in Fig. 40. The transmission device is assumed to support settings for both a mode to transmit two modulated signals, $z1(t)$ and $z2(t)$, and for a mode to transmit one modulated signal.

[0635]

In Fig. 41, the symbol (4100) is a symbol for transmitting the “transmission information” shown in Table 1. The symbols (4101_1) and (4101_2) are reference (pilot) symbols for channel estimation. The symbols (4102_1, 4103_1) are data transmission symbols for transmitting the modulated signal $z1(t)$. The symbols (4102_2, 4103_2) are data transmission symbols for transmitting the modulated signal $z2(t)$. The symbol (4102_1) and the symbol (4102_2) are transmitted at the same time along the same (shared/common) frequency, and the symbol (4103_1) and the symbol (4103_2) are transmitted at the same time along the same (shared/common) frequency. The symbols (4102_1, 4103_1) and the symbols (4102_2, 4103_2) are the symbols after precoding matrix calculation using the method of regularly hopping between precoding matrices described in Embodiments

1–4 and Embodiment 6 (therefore, as described in Embodiment 1, the structure of the streams $s_1(t)$ and $s_2(t)$ is as in Fig. 6).

Furthermore, in Fig. 41, the symbol (4104) is a symbol for transmitting the “transmission information” shown in Table 1. The symbol (4105) is a reference (pilot) symbol for channel estimation. The symbols (4106, 4107) are data transmission symbols for transmitting the modulated signal $z_1(t)$. The data transmission symbols for transmitting the modulated signal $z_1(t)$ are not precoded, since the number of transmission signals is one.

[0636]

Accordingly, the transmission device in Fig. 40 generates and transmits modulated signals in accordance with Table 1 and the frame structure in Fig. 41. In Fig. 40, the frame structure signal 313 includes information regarding the “number of transmission signals”, “modulation method”, “number of encoders”, and “error correction coding method” set based on Table 1. The encoder (4002), the mappers 306A, B, and the weighting units 308A, B receive the frame structure signal as an input and operate based on the “number of transmission signals”, “modulation method”, “number of encoders”, and “error correction coding method” that are set based on Table 1. “Transmission information” corresponding to the set “number of transmission signals”, “modulation method”, “number of encoders”, and “error correction coding method” is also transmitted to the reception device.

[0637]

The structure of the reception device may be represented similarly to Fig. 7 of Embodiment 1. The difference with Embodiment 1 is as follows: since the transmission device and the reception device store the information in Table 1 in advance, the transmission device does not need to transmit information for regularly hopping between precoding matrices, but rather transmits “transmission information” corresponding to the “number of transmission signals”, “modulation method”, “number of encoders”, and “error correction coding method”, and the

reception device obtains information for regularly hopping between precoding matrices from Table 1 by receiving the “transmission information”. Accordingly, by the control information decoding unit 709 obtaining the “transmission information” transmitted by the transmission device in Fig. 40, the reception device in Fig. 7
 5 obtains, from the information corresponding to Table 1, a signal 710 regarding information on the transmission method, as notified by the transmission device, which includes information for regularly hopping between precoding matrices. Therefore, when the number of transmission signals is two, the signal processing unit 711 can perform detection based on a precoding matrix hopping pattern to
 10 obtain received log-likelihood ratios.

[0638]

Note that in the above description, “transmission information” is set with respect to the “number of transmission signals”, “modulation method”, “number of encoders”, and “error correction coding method” as in Table 1, and the precoding
 15 matrix hopping method is set with respect to the “transmission information”. However, it is not necessary to set the “transmission information” with respect to the “number of transmission signals”, “modulation method”, “number of encoders”, and “error correction coding method”. For example, as in Table 2, the “transmission information” may be set with respect to the “number of transmission signals” and
 20 “modulation method”, and the precoding matrix hopping method may be set with respect to the “transmission information”.

[0639]

Table 2

Number of modulated transmission signals (number of transmit antennas)	Modulation method	Transmission information	Precoding matrix hopping method
1	QPSK	00000	-
	16QAM	00001	-

	64QAM	00010	-
	256QAM	00011	-
	1024QAM	00100	-
2	#1: QPSK, #2: QPSK	10000	D
	#1: QPSK, #2: 16QAM	10001	E
	#1: 16QAM, #2: 16QAM	10010	E
	#1: 16QAM, #2: 64QAM	10011	E
	#1: 64QAM, #2: 64QAM	10100	F
	#1: 64QAM, #2: 256QAM	10101	F
	#1: 256QAM, #2: 256QAM	10110	G
	#1: 256QAM, #2: 1024QAM	10111	G
	#1: 1024QAM, #2: 1024QAM	11000	H

[0640]

In this context, the “transmission information” and the method of setting the precoding matrix hopping method is not limited to Tables 1 and 2. As long as a rule is determined in advance for switching the precoding matrix hopping method based on transmission parameters, such as the “number of transmission signals”, “modulation method”, “number of encoders”, “error correction coding method”, or the like (as long as the transmission device and the reception device share a predetermined rule, or in other words, if the precoding matrix hopping method is switched based on any of the transmission parameters (or on any plurality of transmission parameters)), the transmission device does not need to transmit information regarding the precoding matrix hopping method. The reception device

can identify the precoding matrix hopping method used by the transmission device by identifying the information on the transmission parameters and can therefore accurately perform decoding and detection. Note that in Tables 1 and 2, a transmission method that regularly hops between precoding matrices is used when
5 the number of modulated transmission signals is two, but a transmission method that regularly hops between precoding matrices may be used when the number of modulated transmission signals is two or greater.

[0641]

Accordingly, if the transmission device and reception device share a table
10 regarding transmission patterns that includes information on precoding hopping methods, the transmission device need not transmit information regarding the precoding hopping method, transmitting instead control information that does not include information regarding the precoding hopping method, and the reception device can infer the precoding hopping method by acquiring this control
15 information.

[0642]

As described above, in the present embodiment, the transmission device does not transmit information directly related to the method of regularly hopping between precoding matrices. Rather, a method has been described wherein the
20 reception device infers information regarding precoding for the “method of regularly hopping between precoding matrices” used by the transmission device. This method yields the advantageous effect of improved transmission efficiency of data as a result of the transmission device not transmitting information directly related to the method of regularly hopping between precoding matrices.

25 [0643]

Note that the present embodiment has been described as changing precoding weights in the time domain, but as described in Embodiment 1, the present invention

may be similarly embodied when using a multi-carrier transmission method such as OFDM or the like.

[0644]

In particular, when the precoding hopping method only changes depending
5 on the number of transmission signals, the reception device can learn the precoding hopping method by acquiring information, transmitted by the transmission device, on the number of transmission signals.

[0645]

In the present description, it is considered that a
10 communications/broadcasting device such as a broadcast station, a base station, an access point, a terminal, a mobile phone, or the like is provided with the transmission device, and that a communications device such as a television, radio, terminal, personal computer, mobile phone, access point, base station, or the like is provided with the reception device. Additionally, it is considered that the
15 transmission device and the reception device in the present description have a communications function and are capable of being connected via some sort of interface to a device for executing applications for a television, radio, personal computer, mobile phone, or the like.

[0646]

20 Furthermore, in the present embodiment, symbols other than data symbols, such as pilot symbols (preamble, unique word, postamble, reference symbol, and the like), symbols for control information, and the like may be arranged in the frame in any way. While the terms “pilot symbol” and “symbols for control information” have been used here, any term may be used, since the function itself is what is
25 important.

[0647]

It suffices for a pilot symbol, for example, to be a known symbol modulated with PSK modulation in the transmission and reception devices (or for the reception

device to be able to synchronize in order to know the symbol transmitted by the transmission device). The reception device uses this symbol for frequency synchronization, time synchronization, channel estimation (estimation of Channel State Information (CSI) for each modulated signal), detection of signals, and the like.

[0648]

A symbol for control information is for transmitting information other than data (of applications or the like) that needs to be transmitted to the communication partner for achieving communication (for example, the modulation method, error correction coding method, coding ratio of the error correction coding method, setting information in the upper layer, and the like).

[0649]

Note that the present invention is not limited to the above Embodiments 1–5 and may be embodied with a variety of modifications. For example, the above embodiments describe communications devices, but the present invention is not limited to these devices and may be implemented as software for the corresponding communications method.

[0650]

Furthermore, a precoding hopping method used in a method of transmitting two modulated signals from two antennas has been described, but the present invention is not limited in this way. The present invention may be also embodied as a precoding hopping method for similarly changing precoding weights (matrices) in the context of a method whereby four mapped signals are precoded to generate four modulated signals that are transmitted from four antennas, or more generally, whereby N mapped signals are precoded to generate N modulated signals that are transmitted from N antennas.

[0651]

In the description, terms such as “precoding” and “precoding weight” are used, but any other terms may be used. What matters in the present invention is the actual signal processing.

[0652]

5 Different data may be transmitted in streams $s_1(t)$ and $s_2(t)$, or the same data may be transmitted.

[0653]

Each of the transmit antennas of the transmission device and the receive antennas of the reception device shown in the figures may be formed by a plurality
10 of antennas.

[0654]

Programs for executing the above transmission method may, for example, be stored in advance in Read Only Memory (ROM) and be caused to operate by a Central Processing Unit (CPU).

15 [0655]

Furthermore, the programs for executing the above transmission method may be stored in a computer-readable recording medium, the programs stored in the recording medium may be loaded in the Random Access Memory (RAM) of the computer, and the computer may be caused to operate in accordance with the
20 programs.

[0656]

The components in the above embodiments may be typically assembled as a Large Scale Integration (LSI), a type of integrated circuit. Individual components may respectively be made into discrete chips, or part or all of the components in
25 each embodiment may be made into one chip. While an LSI has been referred to, the terms Integrated Circuit (IC), system LSI, super LSI, or ultra LSI may be used depending on the degree of integration. Furthermore, the method for assembling integrated circuits is not limited to LSI, and a dedicated circuit or a general-purpose

processor may be used. A Field Programmable Gate Array (FPGA), which is programmable after the LSI is manufactured, or a reconfigurable processor, which allows reconfiguration of the connections and settings of circuit cells inside the LSI, may be used.

5 [0657]

Furthermore, if technology for forming integrated circuits that replaces LSIs emerges, owing to advances in semiconductor technology or to another derivative technology, the integration of functional blocks may naturally be accomplished using such technology. The application of biotechnology or the like is possible.

10 [0658]

(Embodiment 8)

The present embodiment describes an application of the method described in Embodiments 1–4 and Embodiment 6 for regularly hopping between precoding weights.

15 [0659]

Fig. 6 relates to the weighting method (precoding method) in the present embodiment. The weighting unit 600 integrates the weighting units 308A and 308B in Fig. 3. As shown in Fig. 6, the stream $s_1(t)$ and the stream $s_2(t)$ correspond to the baseband signals 307A and 307B in Fig. 3. In other words, the streams $s_1(t)$ and $s_2(t)$ are the baseband signal in-phase components I and quadrature components Q when mapped according to a modulation scheme such as QPSK, 16QAM, 64QAM, or the like. As indicated by the frame structure of Fig. 6, the stream $s_1(t)$ is represented as $s_1(u)$ at symbol number u , as $s_1(u + 1)$ at symbol number $u + 1$, and so forth. Similarly, the stream $s_2(t)$ is represented as $s_2(u)$ at symbol number u , as $s_2(u + 1)$ at symbol number $u + 1$, and so forth. The weighting unit 600 receives the baseband signals 307A ($s_1(t)$) and 307B ($s_2(t)$) and the information 315 regarding weighting information in Fig. 3 as inputs, performs weighting in accordance with the

information 315 regarding weighting, and outputs the signals 309A ($z1(t)$) and 309B ($z2(t)$) after weighting in Fig. 3.

[0660]

At this point, when for example a precoding matrix hopping method with an
 5 $N = 8$ period (cycle) as in Example #8 in Embodiment 6 is used, $z1(t)$ and $z2(t)$ are represented as follows.

For symbol number $8i$ (where i is an integer greater than or equal to zero):

[0661]

Math 228

10 Equation 218

$$\begin{pmatrix} z1(8i) \\ z2(8i) \end{pmatrix} = \frac{1}{\sqrt{\alpha^2 + 1}} \begin{pmatrix} e^{j0} & \alpha \times e^{j0} \\ \alpha \times e^{j\frac{i\pi}{4}} & e^{j\left(\frac{k\pi}{4} + \frac{7\pi}{8}\right)} \end{pmatrix} \begin{pmatrix} s1(8i) \\ s2(8i) \end{pmatrix}$$

[0662]

Here, j is an imaginary unit, and $k = 0$.

15 For symbol number $8i + 1$:

[0663]

Math 229

Equation 219

$$\begin{pmatrix} z1(8i + 1) \\ z2(8i + 1) \end{pmatrix} = \frac{1}{\sqrt{\alpha^2 + 1}} \begin{pmatrix} e^{j0} & \alpha \times e^{j0} \\ \alpha \times e^{j\frac{i\pi}{4}} & e^{j\left(\frac{k\pi}{4} + \frac{7\pi}{8}\right)} \end{pmatrix} \begin{pmatrix} s1(8i + 1) \\ s2(8i + 1) \end{pmatrix}$$

20

[0664]

Here, $k = 1$.

For symbol number $8i + 2$:

[0665]

Math 230

Equation 220

$$\begin{pmatrix} z1(8i+2) \\ z2(8i+2) \end{pmatrix} = \frac{1}{\sqrt{\alpha^2+1}} \begin{pmatrix} e^{j0} & \alpha \times e^{j0} \\ \alpha \times e^{j\frac{i\pi}{4}} & e^{j\left(\frac{k\pi}{4}+\frac{7\pi}{8}\right)} \end{pmatrix} \begin{pmatrix} s1(8i+2) \\ s2(8i+2) \end{pmatrix}$$

5 [0666]

Here, k = 2.

For symbol number 8i + 3:

[0667]

Math 231

10 Equation 221

$$\begin{pmatrix} z1(8i+3) \\ z2(8i+3) \end{pmatrix} = \frac{1}{\sqrt{\alpha^2+1}} \begin{pmatrix} e^{j0} & \alpha \times e^{j0} \\ \alpha \times e^{j\frac{i\pi}{4}} & e^{j\left(\frac{k\pi}{4}+\frac{7\pi}{8}\right)} \end{pmatrix} \begin{pmatrix} s1(8i+3) \\ s2(8i+3) \end{pmatrix}$$

[0668]

Here, k = 3.

15 For symbol number 8i + 4:

[0669]

Math 232

Equation 222

$$\begin{pmatrix} z1(8i+4) \\ z2(8i+4) \end{pmatrix} = \frac{1}{\sqrt{\alpha^2+1}} \begin{pmatrix} e^{j0} & \alpha \times e^{j0} \\ \alpha \times e^{j\frac{i\pi}{4}} & e^{j\left(\frac{k\pi}{4}+\frac{7\pi}{8}\right)} \end{pmatrix} \begin{pmatrix} s1(8i+4) \\ s2(8i+4) \end{pmatrix}$$

20

[0670]

Here, k = 4.

For symbol number $8i + 5$:

[0671]

Math 233

Equation 223

$$5 \quad \begin{pmatrix} z1(8i + 5) \\ z2(8i + 5) \end{pmatrix} = \frac{1}{\sqrt{\alpha^2 + 1}} \begin{pmatrix} e^{j0} & \alpha \times e^{j0} \\ \alpha \times e^{j\frac{i\pi}{4}} & e^{j\left(\frac{k\pi}{4} + \frac{7\pi}{8}\right)} \end{pmatrix} \begin{pmatrix} s1(8i + 5) \\ s2(8i + 5) \end{pmatrix}$$

[0672]

Here, $k = 5$.

For symbol number $8i + 6$:

10 [0673]

Math 234

Equation 224

$$\begin{pmatrix} z1(8i + 6) \\ z2(8i + 6) \end{pmatrix} = \frac{1}{\sqrt{\alpha^2 + 1}} \begin{pmatrix} e^{j0} & \alpha \times e^{j0} \\ \alpha \times e^{j\frac{i\pi}{4}} & e^{j\left(\frac{k\pi}{4} + \frac{7\pi}{8}\right)} \end{pmatrix} \begin{pmatrix} s1(8i + 6) \\ s2(8i + 6) \end{pmatrix}$$

15 [0674]

Here, $k = 6$.

For symbol number $8i + 7$:

[0675]

Math 235

20 Equation 225

$$\begin{pmatrix} z1(8i + 7) \\ z2(8i + 7) \end{pmatrix} = \frac{1}{\sqrt{\alpha^2 + 1}} \begin{pmatrix} e^{j0} & \alpha \times e^{j0} \\ \alpha \times e^{j\frac{i\pi}{4}} & e^{j\left(\frac{k\pi}{4} + \frac{7\pi}{8}\right)} \end{pmatrix} \begin{pmatrix} s1(8i + 7) \\ s2(8i + 7) \end{pmatrix}$$

[0676]

Here, $k = 7$.

[0677]

The symbol numbers shown here can be considered to indicate time. As described in other embodiments, in Equation 225, for example, $z1(8i + 7)$ and $z2(8i + 7)$ at time $8i + 7$ are signals at the same time, and the transmission device transmits $z1(8i + 7)$ and $z2(8i + 7)$ over the same (shared/common) frequency. In other words, letting the signals at time T be $s1(T)$, $s2(T)$, $z1(T)$, and $z2(T)$, then $z1(T)$ and $z2(T)$ are sought from some sort of precoding matrices and from $s1(T)$ and $s2(T)$, and the transmission device transmits $z1(T)$ and $z2(T)$ over the same (shared) frequency (at the same time). Furthermore, in the case of using a multi-carrier transmission method such as OFDM or the like, and letting signals corresponding to $s1$, $s2$, $z1$, and $z2$ for (sub)carrier L and time T be $s1(T, L)$, $s2(T, L)$, $z1(T, L)$, and $z2(T, L)$, then $z1(T, L)$ and $z2(T, L)$ are sought from some sort of precoding matrices and from $s1(T, L)$ and $s2(T, L)$, and the transmission device transmits $z1(T, L)$ and $z2(T, L)$ over the same (shared/common) frequency (at the same time).

[0678]

In this case, the appropriate value of α is given by Equation 198 or Equation 200.

20 [0679]

The present embodiment describes a precoding hopping method that increases period (cycle) size, based on the above-described precoding matrices of Equation 190.

[0680]

25 Letting the period (cycle) of the precoding hopping method be $8M$, $8M$ different precoding matrices are represented as follows.

[0681]

Math 236

Equation 226

$$\mathbf{F}[8 \times k + i] = \frac{1}{\sqrt{\alpha^2 + 1}} \begin{pmatrix} e^{j0} & \alpha \times e^{j0} \\ \alpha \times e^{j\left(\frac{i\pi}{4} + \frac{k\pi}{4M}\right)} & e^{j\left(\frac{i\pi}{4} + \frac{k\pi}{4M} + \frac{7\pi}{8}\right)} \end{pmatrix}$$

[0682]

5 In this case, $i = 0, 1, 2, 3, 4, 5, 6, 7$, and $k = 0, 1, \dots, M-2, M-1$.

[0683]

For example, letting $M = 2$ and $\alpha < 1$, the poor reception points for $s1$ ($\circ$) and for $s2$ ($\square$) at $k = 0$ are represented as in Fig. 42A. Similarly, the poor reception points for $s1$ ($\circ$) and for $s2$ ($\square$) at $k = 1$ are represented as in Fig. 42B. In this way, based on the precoding matrices in Equation 190, the poor reception points are as in Fig. 42A, and by using, as the precoding matrices, the matrices yielded by multiplying each term in the second line on the right-hand side of Equation 190 by e^{jX} (see Equation 226), the poor reception points are rotated with respect to Fig. 42A (see Fig. 42B). (Note that the poor reception points in Fig. 42A and Fig. 42B do not overlap. Even when multiplying by e^{jX} , the poor reception points should not overlap, as in this case. Furthermore, the matrices yielded by multiplying each term in the first line on the right-hand side of Equation 190, rather than in the second line on the right-hand side of Equation 190, by e^{jX} may be used as the precoding matrices.) In this case, the precoding matrices $\mathbf{F}[0]$ – $\mathbf{F}[15]$ are represented as follows.

20 [0684]

Math 237

Equation 227

$$\mathbf{F}[8 \times k + i] = \frac{1}{\sqrt{\alpha^2 + 1}} \begin{pmatrix} e^{j0} & \alpha \times e^{j0} \\ \alpha \times e^{j\left(\frac{i\pi}{4} + Xk\right)} & e^{j\left(\frac{i\pi}{4} + Xk + \frac{7\pi}{8}\right)} \end{pmatrix}$$

[0685]

Here, $i = 0, 1, 2, 3, 4, 5, 6, 7$, and $k = 0, 1$.

[0686]

In this case, when $M = 2$, precoding matrices $F[0]$ – $F[15]$ are generated (the
 5 precoding matrices $F[0]$ – $F[15]$ may be in any order, and the matrices $F[0]$ – $F[15]$
 may each be different). Symbol number $16i$ may be precoded using $F[0]$, symbol
 number $16i + 1$ may be precoded using $F[1]$, ..., and symbol number $16i + h$ may be
 precoded using $F[h]$, for example ($h = 0, 1, 2, \dots, 14, 15$). (In this case, as described
 in previous embodiments, precoding matrices need not be hopped between
 10 regularly.)

Summarizing the above considerations, with reference to Equations 82–85,
 N- period (cycle) precoding matrices are represented by the following equation.

[0687]

Math 238

15 Equation 228

$$F[i] = \frac{1}{\sqrt{\alpha^2 + 1}} \begin{pmatrix} e^{j\theta_{11}(i)} & \alpha \times e^{j(\theta_{11}(i) + \lambda)} \\ \alpha \times e^{j\theta_{21}(i)} & e^{j(\theta_{21}(i) + \lambda + \delta)} \end{pmatrix}$$

[0688]

Here, since the period (cycle) has N slots, $i = 0, 1, 2, \dots, N - 2, N - 1$.
 20 Furthermore, the $N \times M$ period (cycle) precoding matrices based on Equation 228
 are represented by the following equation.

[0689]

Math 239

Equation 229

$$F[N \times k + i] = \frac{1}{\sqrt{\alpha^2 + 1}} \begin{pmatrix} e^{j\theta_{11}(i)} & \alpha \times e^{j(\theta_{11}(i) + \lambda)} \\ \alpha \times e^{j(\theta_{21}(i) + X_k)} & e^{j(\theta_{21}(i) + X_k + \lambda + \delta)} \end{pmatrix}$$

[0690]

In this case, $i = 0, 1, 2, \dots, N - 2, N - 1$, and $k = 0, 1, \dots, M - 2, M - 1$.

5 [0691]

Precoding matrices $F[0]$ – $F[N \times M - 1]$ are thus generated (the precoding matrices $F[0]$ – $F[N \times M - 1]$ may be in any order for the $N \times M$ slots in the period (cycle)). Symbol number $N \times M \times i$ may be precoded using $F[0]$, symbol number $N \times M \times i + 1$ may be precoded using $F[1]$, ..., and symbol number $N \times M \times i + h$ may be precoded using $F[h]$, for example ($h = 0, 1, 2, \dots, N \times M - 2, N \times M - 1$). (In this case, as described in previous embodiments, precoding matrices need not be hopped between regularly.)

Generating the precoding matrices in this way achieves a precoding matrix hopping method with a large period (cycle), allowing for the position of poor reception points to be easily changed, which may lead to improved data reception quality. Note that while the $N \times M$ period (cycle) precoding matrices have been set to Equation 229, the $N \times M$ period (cycle) precoding matrices may be set to the following equation, as described above.

[0692]

20 Math 240

Equation 230

$$F[N \times k + i] = \frac{1}{\sqrt{\alpha^2 + 1}} \begin{pmatrix} e^{j(\theta_{11}(i) + X_k)} & \alpha \times e^{j(\theta_{11}(i) + X_k + \lambda)} \\ \alpha \times e^{j\theta_{21}(i)} & e^{j(\theta_{21}(i) + \lambda + \delta)} \end{pmatrix}$$

[0693]

In this case, $i = 0, 1, 2, \dots, N - 2, N - 1$, and $k = 0, 1, \dots, M - 2, M - 1$.

[0694]

In Equations 229 and 230, when $0 \text{ radians} \leq \delta < 2\pi \text{ radians}$, the matrices are a unitary matrix when $\delta = \pi \text{ radians}$ and are a non-unitary matrix when $\delta \neq \pi \text{ radians}$.

5 In the present method, use of a non-unitary matrix for $\pi/2 \text{ radians} \leq |\delta| < \pi \text{ radians}$ is one characteristic structure (the conditions for δ being similar to other embodiments), and excellent data reception quality is obtained. Use of a unitary matrix is another structure, and as described in detail in Embodiment 10 and Embodiment 16, if N is an odd number in Equations 229 and 230, the probability of obtaining excellent data
10 reception quality increases.

[0695]

(Embodiment 9)

The present embodiment describes a method for regularly hopping between precoding matrices using a unitary matrix.

15 [0696]

As described in Embodiment 8, in the method of regularly hopping between precoding matrices over a period (cycle) with N slots, the precoding matrices prepared for the N slots with reference to Equations 82–85 are represented as follows.

20 [0697]

Math 241

Equation 231

$$F[i] = \frac{1}{\sqrt{\alpha^2 + 1}} \begin{pmatrix} e^{j\theta_{11}(i)} & \alpha \times e^{j(\theta_{11}(i) + \lambda)} \\ \alpha \times e^{j\theta_{21}(i)} & e^{j(\theta_{21}(i) + \lambda + \delta)} \end{pmatrix}$$

25 [0698]

In this case, $i = 0, 1, 2, \dots, N - 2, N - 1$. (Let $\alpha > 0$.) Since a unitary matrix is used in the present embodiment, the precoding matrices in Equation 231 may be represented as follows.

[0699]

5 Math 242

Equation 232

$$F[i] = \frac{1}{\sqrt{\alpha^2 + 1}} \begin{pmatrix} e^{j\theta_{11}(i)} & \alpha \times e^{j(\theta_{11}(i) + \lambda)} \\ \alpha \times e^{j\theta_{21}(i)} & e^{j(\theta_{21}(i) + \lambda + \pi)} \end{pmatrix}$$

[0700]

10 In this case, $i = 0, 1, 2, \dots, N - 2, N - 1$. (Let $\alpha > 0$.) From Condition #5 (Math 106) and Condition #6 (Math 107) in Embodiment 3, the following condition is important for achieving excellent data reception quality.

[0701]

Math 243

15 Condition #17

$$e^{j(\theta_{11}(x) - \theta_{21}(x))} \neq e^{j(\theta_{11}(y) - \theta_{21}(y))} \quad \text{for } \forall x, \forall y (x \neq y; x, y = 0, 1, 2, \dots, N - 2, N - 1)$$

[0702]

(x is $0, 1, 2, \dots, N - 2, N - 1$; y is $0, 1, 2, \dots, N - 2, N - 1$; and $x \neq y$.)

20 [0703]

Math 244

Condition #18

$$e^{j(\theta_{11}(x) - \theta_{21}(x) - \pi)} \neq e^{j(\theta_{11}(y) - \theta_{21}(y) - \pi)} \quad \text{for } \forall x, \forall y (x \neq y; x, y = 0, 1, 2, \dots, N - 2, N - 1)$$

[0704]

(x is 0, 1, 2, ..., N - 2, N - 1; y is 0, 1, 2, ..., N - 2, N - 1; and x ≠ y.)

Embodiment 6 describes the distance between poor reception points. In order to increase the distance between poor reception points, it is important for the number of slots N to be an odd number three or greater. The following explains this point.

[0705]

In order to distribute the poor reception points evenly with regards to phase in the complex plane, as described in Embodiment 6, Condition #19 and Condition #20 are provided.

[0706]

Math 245

Condition #19

$$\frac{e^{j(\theta_{11}(x+1)-\theta_{21}(x+1))}}{e^{j(\theta_{11}(x)-\theta_{21}(x))}} = e^{j\left(\frac{2\pi}{N}\right)} \quad \text{for } \forall x (x = 0, 1, 2, \dots, N-2)$$

15

[0707]

Math 246

Condition #20

$$\frac{e^{j(\theta_{11}(x+1)-\theta_{21}(x+1))}}{e^{j(\theta_{11}(x)-\theta_{21}(x))}} = e^{j\left(-\frac{2\pi}{N}\right)} \quad \text{for } \forall x (x = 0, 1, 2, \dots, N-2)$$

20

[0708]

In other words, Condition #19 means that the difference in phase is $2\pi/N$ radians. On the other hand, Condition #20 means that the difference in phase is $-2\pi/N$ radians.

[0709]

Letting $\theta_{11}(0) - \theta_{21}(0) = 0$ radians, and letting $\alpha < 1$, the distribution of poor reception points for s1 and for s2 in the complex plane for an $N = 3$ period (cycle) is shown in Fig. 43A, and the distribution of poor reception points for s1 and for s2 in the complex plane for an $N = 4$ period (cycle) is shown in Fig. 43B. Letting $\theta_{11}(0) - \theta_{21}(0) = 0$ radians, and letting $\alpha > 1$, the distribution of poor reception points for s1 and for s2 in the complex plane for an $N = 3$ period (cycle) is shown in Fig. 44A, and the distribution of poor reception points for s1 and for s2 in the complex plane for an $N = 4$ period (cycle) is shown in Fig. 44B.

10 [0710]

In this case, when considering the phase between a line segment from the origin to a poor reception point and a half line along the real axis defined by $\text{real} \geq 0$ (see Fig. 43A), then for either $\alpha > 1$ or $\alpha < 1$, when $N = 4$, the case always occurs wherein the phase for the poor reception points for s1 and the phase for the poor reception points for s2 are the same value. (See 4301, 4302 in Fig. 43B, and 4401, 4402 in Fig. 44B.) In this case, in the complex plane, the distance between poor reception points becomes small. On the other hand, when $N = 3$, the phase for the poor reception points for s1 and the phase for the poor reception points for s2 are never the same value.

20 [0711]

Based on the above, considering how the case always occurs wherein the phase for the poor reception points for s1 and the phase for the poor reception points for s2 are the same value when the number of slots N in the period (cycle) is an even number, setting the number of slots N in the period (cycle) to an odd number increases the probability of a greater distance between poor reception points in the complex plane as compared to when the number of slots N in the period (cycle) is an even number. However, when the number of slots N in the period (cycle) is small, for example when $N \leq 16$, the minimum distance between poor reception points in

the complex plane can be guaranteed to be a certain length, since the number of poor reception points is small. Accordingly, when $N \leq 16$, even if N is an even number, cases do exist where data reception quality can be guaranteed.

[0712]

- 5 Therefore, in the method for regularly hopping between precoding matrices based on Equation 232, when the number of slots N in the period (cycle) is set to an odd number, the probability of improving data reception quality is high. Precoding matrices $F[0]$ – $F[N - 1]$ are generated based on Equation 232 (the precoding matrices $F[0]$ – $F[N - 1]$ may be in any order for the N slots in the period (cycle)).
- 10 Symbol number Ni may be precoded using $F[0]$, symbol number $Ni + 1$ may be precoded using $F[1]$, ..., and symbol number $N \times i + h$ may be precoded using $F[h]$, for example ($h = 0, 1, 2, \dots, N - 2, N - 1$). (In this case, as described in previous embodiments, precoding matrices need not be hopped between regularly.) Furthermore, when the modulation method for both $s1$ and $s2$ is 16QAM, if α is set
- 15 as follows,

[0713]

Math 247

Equation 233

$$\alpha = \frac{\sqrt{2} + 4}{\sqrt{2} + 2}$$

20

[0714]

the advantageous effect of increasing the minimum distance between $16 \times 16 = 256$ signal points in the IQ plane for a specific LOS environment may be achieved.

25 [0715]

In the present embodiment, the method of structuring N different precoding matrices for a precoding hopping method with an N -slot time period (cycle) has been described. In this case, as the N different precoding matrices, $F[0]$, $F[1]$, $F[2]$, ..., $F[N - 2]$, $F[N - 1]$ are prepared. In the present embodiment, an example of a single carrier transmission method has been described, and therefore the case of arranging symbols in the order $F[0]$, $F[1]$, $F[2]$, ..., $F[N - 2]$, $F[N - 1]$ in the time domain (or the frequency domain) has been described. The present invention is not, however, limited in this way, and the N different precoding matrices $F[0]$, $F[1]$, $F[2]$, ..., $F[N - 2]$, $F[N - 1]$ generated in the present embodiment may be adapted to a multi-carrier transmission method such as an OFDM transmission method or the like. As in Embodiment 1, as a method of adaption in this case, precoding weights may be changed by arranging symbols in the frequency domain and in the frequency-time domain. Note that a precoding hopping method with an N -slot time period (cycle) has been described, but the same advantageous effects may be obtained by randomly using N different precoding matrices. In other words, the N different precoding matrices do not necessarily need to be used in a regular period (cycle).

[0716]

Furthermore, in the precoding matrix hopping method over an H -slot period (cycle) (H being a natural number larger than the number of slots N in the period (cycle) of the above method of regularly hopping between precoding matrices), when the N different precoding matrices of the present embodiment are included, the probability of excellent reception quality increases. In this case, Condition #17 and Condition #18 can be replaced by the following conditions. (The number of slots in the period (cycle) is considered to be N .)

[0717]

Math 248

Condition #17'

$$e^{j(\theta_{11}(x)-\theta_{21}(x))} \neq e^{j(\theta_{11}(y)-\theta_{21}(y))} \quad \text{for } \exists x, \exists y (x \neq y; x, y = 0, 1, 2, \dots, N-2, N-1)$$

[0718]

(x is 0, 1, 2, ..., N-2, N-1; y is 0, 1, 2, ..., N-2, N-1; and x ≠ y.)

5 [0719]

Math 249

Condition #18'

$$e^{j(\theta_{11}(x)-\theta_{21}(x)-\pi)} \neq e^{j(\theta_{11}(y)-\theta_{21}(y)-\pi)} \quad \text{for } \exists x, \exists y (x \neq y; x, y = 0, 1, 2, \dots, N-2, N-1)$$

10 [0720]

(x is 0, 1, 2, ..., N-2, N-1; y is 0, 1, 2, ..., N-2, N-1; and x ≠ y.)

(Embodiment 10)

The present embodiment describes a method for regularly hopping between precoding matrices using a unitary matrix that differs from the example in

15 Embodiment 9.

[0721]

In the method of regularly hopping between precoding matrices over a period (cycle) with 2N slots, the precoding matrices prepared for the 2N slots are represented as follows.

20 [0722]

Math 250

Equation 234

for i = 0, 1, 2, ..., N-2, N-1:

$$F[i] = \frac{1}{\sqrt{\alpha^2 + 1}} \begin{pmatrix} e^{j\theta_{11}(i)} & \alpha \times e^{j(\theta_{11}(i)+\lambda)} \\ \alpha \times e^{j\theta_{21}(i)} & e^{j(\theta_{21}(i)+\lambda+\pi)} \end{pmatrix}$$

[0723]

Let α be a fixed value (not depending on i), where $\alpha > 0$.

[0724]

5 Math 251

Equation 235

for $i = N, N + 1, N + 2, \dots, 2N - 2, 2N - 1$:

$$F[i] = \frac{1}{\sqrt{\alpha^2 + 1}} \begin{pmatrix} \alpha \times e^{j\theta_{11}(i)} & e^{j(\theta_{11}(i) + \lambda)} \\ e^{j\theta_{21}(i)} & \alpha \times e^{j(\theta_{21}(i) + \lambda + \pi)} \end{pmatrix}$$

10 [0725]

Let α be a fixed value (not depending on i), where $\alpha > 0$. (Let the α in Equation 234 and the α in Equation 235 be the same value.)

From Condition #5 (Math 106) and Condition #6 (Math 107) in Embodiment 3, the following conditions are important in Equation 234 for achieving
15 excellent data reception quality.

[0726]

Math 252

Condition #21

$$e^{j(\theta_{11}(x) - \theta_{21}(x))} \neq e^{j(\theta_{11}(y) - \theta_{21}(y))} \quad \text{for } \forall x, \forall y (x \neq y; x, y = 0, 1, 2, \dots, N - 2, N - 1)$$

20

[0727]

(x is $0, 1, 2, \dots, N - 2, N - 1$; y is $0, 1, 2, \dots, N - 2, N - 1$; and $x \neq y$.)

[0728]

Math 253

25 Condition #22

$$e^{j(\theta_{11}(x)-\theta_{21}(x)-\pi)} \neq e^{j(\theta_{11}(y)-\theta_{21}(y)-\pi)} \quad \text{for } \forall x, \forall y (x \neq y; x, y = 0, 1, 2, \dots, N-2, N-1)$$

[0729]

(x is 0, 1, 2, ..., N-2, N-1; y is 0, 1, 2, ..., N-2, N-1; and x ≠ y.)

Addition of the following condition is considered.

5 [0730]

Math 254

Condition #23

$$\theta_{11}(x) = \theta_{11}(x + N) \quad \text{for } \forall x (x = 0, 1, 2, \dots, N-2, N-1)$$

and

$$10 \quad \theta_{21}(y) = \theta_{21}(y + N) \quad \text{for } \forall y (y = 0, 1, 2, \dots, N-2, N-1)$$

[0731]

Next, in order to distribute the poor reception points evenly with regards to phase in the complex plane, as described in Embodiment 6, Condition #24 and

15 Condition #25 are provided.

[0732]

Math 255

Condition #24

$$\frac{e^{j(\theta_{11}(x+1)-\theta_{21}(x+1))}}{e^{j(\theta_{11}(x)-\theta_{21}(x))}} = e^{j\left(\frac{2\pi}{N}\right)} \quad \text{for } \forall x (x = 0, 1, 2, \dots, N-2)$$

20

[0733]

Math 256

Condition #25

$$\frac{e^{j(\theta_{11}(x+1)-\theta_{21}(x+1))}}{e^{j(\theta_{11}(x)-\theta_{21}(x))}} = e^{j\left(-\frac{2\pi}{N}\right)} \quad \text{for } \forall x (x = 0, 1, 2, \dots, N-2)$$

[0734]

In other words, Condition #24 means that the difference in phase is $2\pi/N$ radians. On the other hand, Condition #25 means that the difference in phase is $-2\pi/N$ radians.

[0735]

Letting $\theta_{11}(0) - \theta_{21}(0) = 0$ radians, and letting $\alpha > 1$, the distribution of poor reception points for s1 and for s2 in the complex plane when $N = 4$ is shown in Figs. 45A and 45B. As is clear from Figs. 45A and 45B, in the complex plane, the minimum distance between poor reception points for s1 is kept large, and similarly, the minimum distance between poor reception points for s2 is also kept large. Similar conditions are created when $\alpha < 1$. Furthermore, making the same considerations as in Embodiment 9, the probability of a greater distance between poor reception points in the complex plane increases when N is an odd number as compared to when N is an even number. However, when N is small, for example when $N \leq 16$, the minimum distance between poor reception points in the complex plane can be guaranteed to be a certain length, since the number of poor reception points is small. Accordingly, when $N \leq 16$, even if N is an even number, cases do exist where data reception quality can be guaranteed.

[0736]

Therefore, in the method for regularly hopping between precoding matrices based on Equations 234 and 235, when N is set to an odd number, the probability of improving data reception quality is high. Precoding matrices $F[0]-F[2N-1]$ are generated based on Equations 234 and 235 (the precoding matrices $F[0]-F[2N-1]$ may be arranged in any order for the $2N$ slots in the period (cycle)). Symbol number

2Ni may be precoded using F[0], symbol number 2Ni + 1 may be precoded using F[1], ..., and symbol number 2N × i + h may be precoded using F[h], for example (h = 0, 1, 2, ..., 2N - 2, 2N - 1). (In this case, as described in previous embodiments, precoding matrices need not be hopped between regularly.) Furthermore, when the modulation method for both s1 and s2 is 16QAM, if α is set as in Equation 233, the advantageous effect of increasing the minimum distance between 16 × 16 = 256 signal points in the IQ plane for a specific LOS environment may be achieved.

[0737]

The following conditions are possible as conditions differing from

10 Condition #23:

[0738]

Math 257

Condition #26

$$e^{j(\theta_{11}(x)-\theta_{21}(x))} \neq e^{j(\theta_{11}(y)-\theta_{21}(y))} \quad \text{for } \forall x, \forall y (x \neq y; x, y = N, N+1, N+2, \dots, 2N-2, 2N-1)$$

15

[0739]

(where x is N, N - 1, N + 2, ..., 2N - 2, 2N - 1; y is N, N + 1, N + 2, ..., 2N - 2, 2N - 1; and x ≠ y.)

[0740]

20 Math 258

Condition #27

$$e^{j(\theta_{11}(x)-\theta_{21}(x)-\pi)} \neq e^{j(\theta_{11}(y)-\theta_{21}(y)-\pi)} \quad \text{for } \forall x, \forall y (x \neq y; x, y = N, N+1, N+2, \dots, 2N-2, 2N-1)$$

[0741]

25 (where x is N, N + 1, N + 2, ..., 2N - 2, 2N - 1; y is N, N + 1, N + 2, ..., 2N - 2, 2N - 1; and x ≠ y.)

In this case, by satisfying Condition #21, Condition #22, Condition #26, and Condition #27, the distance in the complex plane between poor reception points for

s1 is increased, as is the distance between poor reception points for s2, thereby achieving excellent data reception quality.

[0742]

In the present embodiment, the method of structuring $2N$ different precoding matrices for a precoding hopping method with a $2N$ -slot time period (cycle) has been described. In this case, as the $2N$ different precoding matrices, $F[0]$, $F[1]$, $F[2]$, ..., $F[2N - 2]$, $F[2N - 1]$ are prepared. In the present embodiment, an example of a single carrier transmission method has been described, and therefore the case of arranging symbols in the order $F[0]$, $F[1]$, $F[2]$, ..., $F[2N - 2]$, $F[2N - 1]$ in the time domain (or the frequency domain) has been described. The present invention is not, however, limited in this way, and the $2N$ different precoding matrices $F[0]$, $F[1]$, $F[2]$, ..., $F[2N - 2]$, $F[2N - 1]$ generated in the present embodiment may be adapted to a multi-carrier transmission method such as an OFDM transmission method or the like. As in Embodiment 1, as a method of adaption in this case, precoding weights may be changed by arranging symbols in the frequency domain and in the frequency-time domain. Note that a precoding hopping method with a $2N$ -slot time period (cycle) has been described, but the same advantageous effects may be obtained by randomly using $2N$ different precoding matrices. In other words, the $2N$ different precoding matrices do not necessarily need to be used in a regular period (cycle).

[0743]

Furthermore, in the precoding matrix hopping method over an H -slot period (cycle) (H being a natural number larger than the number of slots $2N$ in the period (cycle) of the above method of regularly hopping between precoding matrices), when the $2N$ different precoding matrices of the present embodiment are included, the probability of excellent reception quality increases.

(Embodiment 11)

The present embodiment describes a method for regularly hopping between precoding matrices using a non-unitary matrix.

[0744]

In the method of regularly hopping between precoding matrices over a period (cycle) with $2N$ slots, the precoding matrices prepared for the $2N$ slots are represented as follows.

[0745]

Math 259

Equation 236

10 for $i = 0, 1, 2, \dots, N-2, N-1$:

$$F[i] = \frac{1}{\sqrt{\alpha^2 + 1}} \begin{pmatrix} e^{j\theta_{11}(i)} & \alpha \times e^{j(\theta_{11}(i) + \lambda)} \\ \alpha \times e^{j\theta_{21}(i)} & e^{j(\theta_{21}(i) + \lambda + \delta)} \end{pmatrix}$$

[0746]

Let α be a fixed value (not depending on i), where $\alpha > 0$. Furthermore, let $\delta \neq \pi$ radians.

[0747]

Math 260

Equation 237

for $i = N, N+1, N+2, \dots, 2N-2, 2N-1$:

20

$$F[i] = \frac{1}{\sqrt{\alpha^2 + 1}} \begin{pmatrix} \alpha \times e^{j(\theta_{11}(i) + \lambda)} & e^{j\theta_{11}(i)} \\ e^{j(\theta_{21}(i) + \lambda + \delta)} & \alpha \times e^{j\theta_{21}(i)} \end{pmatrix}$$

[0748]

Let α be a fixed value (not depending on i), where $\alpha > 0$. (Let the α in Equation 236 and the α in Equation 237 be the same value.)

From Condition #5 (Math 106) and Condition #6 (Math 107) in Embodiment 3, the following conditions are important in Equation 236 for achieving excellent data reception quality.

[0749]

5 Math 261

Condition #28

$$e^{j(\theta_{11}(x)-\theta_{21}(x))} \neq e^{j(\theta_{11}(y)-\theta_{21}(y))} \quad \text{for } \forall x, \forall y (x \neq y; x, y = 0, 1, 2, \dots, N-2, N-1)$$

[0750]

10 (x is 0, 1, 2, ..., N-2, N-1; y is 0, 1, 2, ..., N-2, N-1; and x ≠ y.)

[0751]

Math 262

Condition #29

$$e^{j(\theta_{11}(x)-\theta_{21}(x)-\delta)} \neq e^{j(\theta_{11}(y)-\theta_{21}(y)-\delta)} \quad \text{for } \forall x, \forall y (x \neq y; x, y = 0, 1, 2, \dots, N-2, N-1)$$

15

[0752]

(x is 0, 1, 2, ..., N-2, N-1; y is 0, 1, 2, ..., N-2, N-1; and x ≠ y.)

Addition of the following condition is considered.

[0753]

20 Math 263

Condition #30

$$\theta_{11}(x) = \theta_{11}(x + N) \quad \text{for } \forall x (x = 0, 1, 2, \dots, N-2, N-1)$$

and

$$\theta_{21}(y) = \theta_{21}(y + N) \quad \text{for } \forall y (y = 0, 1, 2, \dots, N-2, N-1)$$

25

[0754]

Note that instead of Equation 237, the precoding matrices in the following Equation may be provided.

[0755]

Math 264

5 Equation 238

for $i = N, N + 1, N + 2, \dots, 2N - 2, 2N - 1$:

$$F[i] = \frac{1}{\sqrt{\alpha^2 + 1}} \begin{pmatrix} \alpha \times e^{j\theta_{11}(i)} & e^{j(\theta_{11}(i) + \lambda)} \\ e^{j\theta_{21}(i)} & \alpha \times e^{j(\theta_{21}(i) + \lambda - \delta)} \end{pmatrix}$$

[0756]

10 Let α be a fixed value (not depending on i), where $\alpha > 0$. (Let the α in Equation 236 and the α in Equation 238 be the same value.)

As an example, in order to distribute the poor reception points evenly with regards to phase in the complex plane, as described in Embodiment 6, Condition #31 and Condition #32 are provided.

15 [0757]

Math 265

Condition #31

$$\frac{e^{j(\theta_{11}(x+1) - \theta_{21}(x+1))}}{e^{j(\theta_{11}(x) - \theta_{21}(x))}} = e^{j\left(\frac{2\pi}{N}\right)} \quad \text{for } \forall x (x = 0, 1, 2, \dots, N - 2)$$

20 [0758]

Math 266

Condition #32

$$\frac{e^{j(\theta_{11}(x+1)-\theta_{21}(x+1))}}{e^{j(\theta_{11}(x)-\theta_{21}(x))}} = e^{j\left(-\frac{2\pi}{N}\right)} \quad \text{for } \forall x (x = 0, 1, 2, \dots, N-2)$$

[0759]

In other words, Condition #31 means that the difference in phase is $2\pi/N$ radians. On the other hand, Condition #32 means that the difference in phase is $-2\pi/N$ radians.

[0760]

Letting $\theta_{11}(0) - \theta_{21}(0) = 0$ radians, letting $\alpha > 1$, and letting $\delta = (3\pi)/4$ radians, the distribution of poor reception points for s1 and for s2 in the complex plane when $N = 4$ is shown in Figs. 46A and 46B. With these settings, the period (cycle) for hopping between precoding matrices is increased, and the minimum distance between poor reception points for s1, as well as the minimum distance between poor reception points for s2, in the complex plane is kept large, thereby achieving excellent reception quality. An example in which $\alpha > 1$, $\delta = (3\pi)/4$ radians, and $N = 4$ has been described, but the present invention is not limited in this way. Similar advantageous effects may be obtained for $\pi/2$ radians $\leq |\delta| < \pi$ radians, $\alpha > 0$, and $\alpha \neq 1$.

[0761]

The following conditions are possible as conditions differing from Condition #30:

[0762]

Math 267

Condition #33

$$e^{j(\theta_{11}(x)-\theta_{21}(x))} \neq e^{j(\theta_{11}(y)-\theta_{21}(y))} \quad \text{for } \forall x, \forall y (x \neq y; x, y = N, N+1, N+2, \dots, 2N-2, 2N-1)$$

25

[0763]

(where x is $N, N + 1, N + 2, \dots, 2N - 2, 2N - 1$; y is $N, N + 1, N + 2, \dots, 2N - 2, 2N - 1$; and $x \neq y$.)

[0764]

Math 268

5 Condition #34

$$e^{j(\theta_{11}(x)-\theta_{21}(x)-\pi)} \neq e^{j(\theta_{11}(y)-\theta_{21}(y)-\pi)} \quad \text{for } \forall x, \forall y (x \neq y; x, y = N, N+1, N+2, \dots, 2N-2, 2N-1)$$

[0765]

(where x is $N, N + 1, N + 2, \dots, 2N - 2, 2N - 1$; y is $N, N + 1, N + 2, \dots, 2N - 2,$

10 $2N - 1$; and $x \neq y$.)

In this case, by satisfying Condition #28, Condition #29, Condition #33, and Condition #34, the distance in the complex plane between poor reception points for s_1 is increased, as is the distance between poor reception points for s_2 , thereby achieving excellent data reception quality.

15 [0766]

In the present embodiment, the method of structuring $2N$ different precoding matrices for a precoding hopping method with a $2N$ -slot time period (cycle) has been described. In this case, as the $2N$ different precoding matrices, $F[0], F[1], F[2], \dots, F[2N - 2], F[2N - 1]$ are prepared. In the present embodiment, an example of a single carrier transmission method has been described, and therefore the case of arranging symbols in the order $F[0], F[1], F[2], \dots, F[2N - 2], F[2N - 1]$ in the time domain (or the frequency domain) has been described. The present invention is not, however, limited in this way, and the $2N$ different precoding matrices $F[0], F[1], F[2], \dots, F[2N - 2], F[2N - 1]$ generated in the present embodiment may be adapted to a multi-carrier transmission method such as an OFDM transmission method or the like. As in Embodiment 1, as a method of adaption in this case, precoding weights may be changed by arranging symbols in the frequency domain and in the frequency-time domain. Note that a precoding

hopping method with a $2N$ -slot time period (cycle) has been described, but the same advantageous effects may be obtained by randomly using $2N$ different precoding matrices. In other words, the $2N$ different precoding matrices do not necessarily need to be used in a regular period (cycle).

5 [0767]

Furthermore, in the precoding matrix hopping method over an H -slot period (cycle) (H being a natural number larger than the number of slots $2N$ in the period (cycle) of the above method of regularly hopping between precoding matrices), when the $2N$ different precoding matrices of the present embodiment are included,
10 the probability of excellent reception quality increases.

(Embodiment 12)

The present embodiment describes a method for regularly hopping between precoding matrices using a non-unitary matrix.

[0768]

15 In the method of regularly hopping between precoding matrices over a period (cycle) with N slots, the precoding matrices prepared for the N slots are represented as follows.

[0769]

Math 269

20 Equation 239

$$F[i] = \frac{1}{\sqrt{\alpha^2 + 1}} \begin{pmatrix} e^{j\theta_{11}(i)} & \alpha \times e^{j(\theta_{11}(i) + \lambda)} \\ \alpha \times e^{j\theta_{21}(i)} & e^{j(\theta_{21}(i) + \lambda + \delta)} \end{pmatrix}$$

[0770]

Let α be a fixed value (not depending on i), where $\alpha > 0$. Furthermore, let $\delta \neq \pi$
25 radians (a fixed value not depending on i), and $i = 0, 1, 2, \dots, N - 2, N - 1$.

[0771]

From Condition #5 (Math 106) and Condition #6 (Math 107) in Embodiment 3, the following conditions are important in Equation 239 for achieving excellent data reception quality.

[0772]

5 Math 270

Condition #35

$$e^{j(\theta_{11}(x)-\theta_{21}(x))} \neq e^{j(\theta_{11}(y)-\theta_{21}(y))} \quad \text{for } \forall x, \forall y (x \neq y; x, y = 0, 1, 2, \dots, N-2, N-1)$$

[0773]

10 (x is 0, 1, 2, ..., N-2, N-1; y is 0, 1, 2, ..., N-2, N-1; and x ≠ y.)

[0774]

Math 271

Condition #36

$$e^{j(\theta_{11}(x)-\theta_{21}(x)-\delta)} \neq e^{j(\theta_{11}(y)-\theta_{21}(y)-\delta)} \quad \text{for } \forall x, \forall y (x \neq y; x, y = 0, 1, 2, \dots, N-2, N-1)$$

15

[0775]

(x is 0, 1, 2, ..., N-2, N-1; y is 0, 1, 2, ..., N-2, N-1; and x ≠ y.)

As an example, in order to distribute the poor reception points evenly with regards to phase in the complex plane, as described in Embodiment 6, Condition #37

20 and Condition #38 are provided.

[0776]

Math 272

Condition #37

$$\frac{e^{j(\theta_{11}(x+1)-\theta_{21}(x+1))}}{e^{j(\theta_{11}(x)-\theta_{21}(x))}} = e^{j\left(\frac{2\pi}{N}\right)} \quad \text{for } \forall x (x = 0, 1, 2, \dots, N-2)$$

[0777]

Math 273

Condition #38

$$5 \quad \frac{e^{j(\theta_{11}(x+1)-\theta_{21}(x+1))}}{e^{j(\theta_{11}(x)-\theta_{21}(x))}} = e^{j\left(-\frac{2\pi}{N}\right)} \quad \text{for } \forall x (x = 0, 1, 2, \dots, N-2)$$

[0778]

In other words, Condition #37 means that the difference in phase is $2\pi/N$ radians. On the other hand, Condition #38 means that the difference in phase is
10 $-2\pi/N$ radians.

[0779]

In this case, if $\pi/2$ radians $\leq |\delta| < \pi$ radians, $\alpha > 0$, and $\alpha \neq 1$, the distance in the complex plane between poor reception points for s_1 is increased, as is the distance between poor reception points for s_2 , thereby achieving excellent data
15 reception quality. Note that Condition #37 and Condition #38 are not always necessary.

[0780]

In the present embodiment, the method of structuring N different precoding matrices for a precoding hopping method with an N -slot time period (cycle) has
20 been described. In this case, as the N different precoding matrices, $F[0]$, $F[1]$, $F[2]$, ..., $F[N-2]$, $F[N-1]$ are prepared. In the present embodiment, an example of a single carrier transmission method has been described, and therefore the case of arranging symbols in the order $F[0]$, $F[1]$, $F[2]$, ..., $F[N-2]$, $F[N-1]$ in the time domain (or the frequency domain) has been described. The present invention is not,
25 however, limited in this way, and the N different precoding matrices $F[0]$, $F[1]$, $F[2]$, ..., $F[N-2]$, $F[N-1]$ generated in the present embodiment may be adapted to a

multi-carrier transmission method such as an OFDM transmission method or the like. As in Embodiment 1, as a method of adaption in this case, precoding weights may be changed by arranging symbols in the frequency domain and in the frequency-time domain. Note that a precoding hopping method with an N-slot time period (cycle) has been described, but the same advantageous effects may be obtained by randomly using N different precoding matrices. In other words, the N different precoding matrices do not necessarily need to be used in a regular period (cycle).

[0781]

Furthermore, in the precoding matrix hopping method over an H-slot period (cycle) (H being a natural number larger than the number of slots N in the period (cycle) of the above method of regularly hopping between precoding matrices), when the N different precoding matrices of the present embodiment are included, the probability of excellent reception quality increases. In this case, Condition #35 and Condition #36 can be replaced by the following conditions. (The number of slots in the period (cycle) is considered to be N.)

[0782]

Math 274

Condition #35'

$$e^{j(\theta_{11}(x)-\theta_{21}(x))} \neq e^{j(\theta_{11}(y)-\theta_{21}(y))} \quad \text{for } \exists x, \exists y (x \neq y; x, y = 0, 1, 2, \dots, N-2, N-1)$$

20

[0783]

(x is 0, 1, 2, ..., N-2, N-1; y is 0, 1, 2, ..., N-2, N-1; and x ≠ y.)

[0784]

Math 275

25 Condition #36'

$$e^{j(\theta_{11}(x)-\theta_{21}(x)-\delta)} \neq e^{j(\theta_{11}(y)-\theta_{21}(y)-\delta)} \quad \text{for } \exists x, \exists y (x \neq y; x, y = 0, 1, 2, \dots, N-2, N-1)$$

[0785]

(x is 0, 1, 2, ..., N - 2, N - 1; y is 0, 1, 2, ..., N - 2, N - 1; and x ≠ y.)

(Embodiment 13)

5 The present embodiment describes a different example than Embodiment 8.

[0786]

In the method of regularly hopping between precoding matrices over a period (cycle) with 2N slots, the precoding matrices prepared for the 2N slots are represented as follows.

10 [0787]

Math 276

Equation 240

for i = 0, 1, 2, ..., N - 2, N - 1:

$$F[i] = \frac{1}{\sqrt{\alpha^2 + 1}} \begin{pmatrix} e^{j\theta_{11}(i)} & \alpha \times e^{j(\theta_{11}(i) + \lambda)} \\ \alpha \times e^{j\theta_{21}(i)} & e^{j(\theta_{21}(i) + \lambda + \delta)} \end{pmatrix}$$

15

[0788]

Let α be a fixed value (not depending on i), where α > 0. Furthermore, let δ ≠ π radians.

[0789]

20 Math 277

Equation 241

for i = N, N + 1, N + 2, ..., 2N - 2, 2N - 1:

$$F[i] = \frac{1}{\sqrt{\alpha^2 + 1}} \begin{pmatrix} \alpha \times e^{j(\theta_{11}(i) + \lambda)} & e^{j\theta_{11}(i)} \\ e^{j(\theta_{21}(i) + \lambda + \delta)} & \alpha \times e^{j\theta_{21}(i)} \end{pmatrix}$$

[0790]

Let α be a fixed value (not depending on i), where $\alpha > 0$. (Let the α in Equation 240 and the α in Equation 241 be the same value.)

Furthermore, the $2 \times N \times M$ period (cycle) precoding matrices based on
 5 Equations 240 and 241 are represented by the following equations.

[0791]

Math 278

Equation 242

for $i = 0, 1, 2, \dots, N-2, N-1$:

$$10 \quad F[2 \times N \times k + i] = \frac{1}{\sqrt{\alpha^2 + 1}} \begin{pmatrix} e^{j\theta_{11}(i)} & \alpha \times e^{j(\theta_{11}(i) + \lambda)} \\ \alpha \times e^{j(\theta_{21}(i) + X_k)} & e^{j(\theta_{21}(i) + X_k + \lambda + \delta)} \end{pmatrix}$$

[0792]

In this case, $k = 0, 1, \dots, M-2, M-1$.

[0793]

15 Math 279

Equation 243

for $i = N, N+1, N+2, \dots, 2N-2, 2N-1$:

$$F[2 \times N \times k + i] = \frac{1}{\sqrt{\alpha^2 + 1}} \begin{pmatrix} \alpha \times e^{j(\theta_{11}(i) + \lambda)} & e^{j\theta_{11}(i)} \\ e^{j(\theta_{21}(i) + \lambda + \delta + Y_k)} & \alpha \times e^{j\theta_{21}(i) + Y_k} \end{pmatrix}$$

20 [0794]

In this case, $k = 0, 1, \dots, M-2, M-1$. Furthermore, $X_k = Y_k$ may be true, or $X_k \neq Y_k$ may be true.

[0795]

Precoding matrices $F[0]$ – $F[2 \times N \times M - 1]$ are thus generated (the
 precoding matrices $F[0]$ – $F[2 \times N \times M - 1]$ may be in any order for the $2 \times N \times M$
 slots in the period (cycle)). Symbol number $2 \times N \times M \times i$ may be precoded using
 $F[0]$, symbol number $2 \times N \times M \times i + 1$ may be precoded using $F[1]$, ..., and symbol
 5 number $2 \times N \times M \times i + h$ may be precoded using $F[h]$, for example ($h = 0, 1, 2, \dots$
 $2 \times N \times M - 2, 2 \times N \times M - 1$). (In this case, as described in previous embodiments,
 precoding matrices need not be hopped between regularly.)

Generating the precoding matrices in this way achieves a precoding matrix
 hopping method with a large period (cycle), allowing for the position of poor
 10 reception points to be easily changed, which may lead to improved data reception
 quality.

[0796]

The $2 \times N \times M$ period (cycle) precoding matrices in Equation 242 may be
 changed to the following equation.

15 [0797]

Math 280

Equation 244

for $i = 0, 1, 2, \dots, N - 2, N - 1$:

$$F[2 \times N \times k + i] = \frac{1}{\sqrt{\alpha^2 + 1}} \begin{pmatrix} e^{j(\theta_{11}(i) + X_k)} & \alpha \times e^{j(\theta_{11}(i) + X_k + \lambda)} \\ \alpha \times e^{j\theta_{21}(i)} & e^{j(\theta_{21}(i) + \lambda + \delta)} \end{pmatrix}$$

20

[0798]

In this case, $k = 0, 1, \dots, M - 2, M - 1$.

[0799]

The $2 \times N \times M$ period (cycle) precoding matrices in Equation 243 may also
 25 be changed to any of Equations 245–247.

[0800]

Math 281

Equation 245

for $i = N, N + 1, N + 2, \dots, 2N - 2, 2N - 1$:

$$F[2 \times N \times k + i] = \frac{1}{\sqrt{\alpha^2 + 1}} \begin{pmatrix} \alpha \times e^{j(\theta_{11}(i) + \lambda + Y_k)} & e^{j\theta_{11}(i + Y_k)} \\ e^{j(\theta_{21}(i) + \lambda + \delta)} & \alpha \times e^{j\theta_{21}(i)} \end{pmatrix}$$

5

[0801]

In this case, $k = 0, 1, \dots, M - 2, M - 1$.

[0802]

Math 282

10 Equation 246

for $i = N, N + 1, N + 2, \dots, 2N - 2, 2N - 1$:

$$F[2 \times N \times k + i] = \frac{1}{\sqrt{\alpha^2 + 1}} \begin{pmatrix} \alpha \times e^{j\theta_{11}(i)} & e^{j(\theta_{11}(i) + \lambda)} \\ e^{j\theta_{21}(i + Y_k)} & \alpha \times e^{j(\theta_{21}(i) + \lambda - \delta + Y_k)} \end{pmatrix}$$

[0803]

15 In this case, $k = 0, 1, \dots, M - 2, M - 1$.

[0804]

Math 283

Equation 247

for $i = N, N + 1, N + 2, \dots, 2N - 2, 2N - 1$:

$$20 \quad F[2 \times N \times k + i] = \frac{1}{\sqrt{\alpha^2 + 1}} \begin{pmatrix} \alpha \times e^{j\theta_{11}(i + Y_k)} & e^{j(\theta_{11}(i) + \lambda + Y_k)} \\ e^{j\theta_{21}(i)} & \alpha \times e^{j(\theta_{21}(i) + \lambda - \delta)} \end{pmatrix}$$

[0805]

In this case, $k = 0, 1, \dots, M-2, M-1$.

[0806]

Focusing on poor reception points, if Equations 242 through 247 satisfy the following conditions,

5 [0807]

Math 284

Condition #39

$$e^{j(\theta_{11}(x)-\theta_{21}(x))} \neq e^{j(\theta_{11}(y)-\theta_{21}(y))} \quad \text{for } \forall x, \forall y (x \neq y; x, y = 0, 1, 2, \dots, N-2, N-1)$$

10 [0808]

(x is $0, 1, 2, \dots, N-2, N-1$; y is $0, 1, 2, \dots, N-2, N-1$; and $x \neq y$.)

[0809]

Math 285

Condition #40

15
$$e^{j(\theta_{11}(x)-\theta_{21}(x)-\delta)} \neq e^{j(\theta_{11}(y)-\theta_{21}(y)-\delta)} \quad \text{for } \forall x, \forall y (x \neq y; x, y = 0, 1, 2, \dots, N-2, N-1)$$

[0810]

(x is $0, 1, 2, \dots, N-2, N-1$; y is $0, 1, 2, \dots, N-2, N-1$; and $x \neq y$.)

[0811]

20 Math 286

Condition #41

$$\theta_{11}(x) = \theta_{11}(x+N) \quad \text{for } \forall x (x = 0, 1, 2, \dots, N-2, N-1)$$

and

$$\theta_{21}(y) = \theta_{21}(y+N) \quad \text{for } \forall y (y = 0, 1, 2, \dots, N-2, N-1)$$

25

[0812]

then excellent data reception quality is achieved. Note that in Embodiment 8, Condition #39 and Condition #40 should be satisfied.

[0813]

5 Focusing on X_k and Y_k , if Equations 242 through 247 satisfy the following conditions,

[0814]

Math 287

Condition #42

$$10 \quad X_a \neq X_b + 2 \times s \times \pi \quad \text{for } \forall a, \forall b (a \neq b; a, b = 0, 1, 2, \dots, M-2, M-1)$$

[0815]

(a is 0, 1, 2, ..., M-2, M-1; b is 0, 1, 2, ..., M-2, M-1; and $a \neq b$.)

(Here, s is an integer.)

15 [0816]

Math 288

Condition #43

$$Y_a \neq Y_b + 2 \times u \times \pi \quad \text{for } \forall a, \forall b (a \neq b; a, b = 0, 1, 2, \dots, M-2, M-1)$$

20 [0817]

(a is 0, 1, 2, ..., M-2, M-1; b is 0, 1, 2, ..., M-2, M-1; and $a \neq b$.)

(Here, u is an integer.)

then excellent data reception quality is achieved. Note that in Embodiment 8, Condition #42 should be satisfied.

25 [0818]

In Equations 242 and 247, when $0 \text{ radians} \leq \delta < 2\pi \text{ radians}$, the matrices are a unitary matrix when $\delta = \pi \text{ radians}$ and are a non-unitary matrix when $\delta \neq \pi \text{ radians}$. In the present method, use of a non-unitary matrix for $\pi/2 \text{ radians} \leq |\delta| < \pi \text{ radians}$ is

one characteristic structure, and excellent data reception quality is obtained. Use of a unitary matrix is another structure, and as described in detail in Embodiment 10 and Embodiment 16, if N is an odd number in Equations 242 through 247, the probability of obtaining excellent data reception quality increases.

5 [0819]

(Embodiment 14)

The present embodiment describes an example of differentiating between usage of a unitary matrix and a non-unitary matrix as the precoding matrix in the method for regularly hopping between precoding matrices.

10 [0820]

The following describes an example that uses a two-by-two precoding matrix (letting each element be a complex number), i.e. the case when two modulated signals ($s_1(t)$ and $s_2(t)$) that are based on a modulation method are precoded, and the two precoded signals are transmitted by two antennas.

15 When transmitting data using a method of regularly hopping between precoding matrices, the mappers 306A and 306B in the transmission device in Fig. 3 and Fig. 13 switch the modulation method in accordance with the frame structure signal 313. The relationship between the modulation level (the number of signal points for the modulation method in the IQ plane) of the modulation method and the precoding matrices is described.

[0821]

The advantage of the method of regularly hopping between precoding matrices is that, as described in Embodiment 6, excellent data reception quality is achieved in an LOS environment. In particular, when the reception device performs ML calculation or applies APP (or Max-log APP) based on ML calculation, the advantageous effect is considerable. Incidentally, ML calculation greatly impacts circuit scale (calculation scale) in accordance with the modulation level of the modulation method. For example, when two precoded signals are transmitted from

two antennas, and the same modulation method is used for two modulated signals (signals based on the modulation method before precoding), the number of candidate signal points in the IQ plane (received signal points 1101 in Fig. 11) is $4 \times 4 = 16$ when the modulation method is QPSK, $16 \times 16 = 256$ when the modulation method is 16QAM, $64 \times 64 = 4096$ when the modulation method is 64QAM, $256 \times 256 = 65,536$ when the modulation method is 256QAM, and $1024 \times 1024 = 1,048,576$ when the modulation method is 1024QAM. In order to keep the calculation scale of the reception device down to a certain circuit size, when the modulation method is QPSK, 16QAM, or 64QAM, ML calculation ((Max-log) APP based on ML calculation) is used, and when the modulation method is 256QAM or 1024QAM, linear operation such as MMSE or ZF is used in the reception device. (In some cases, ML calculation may be used for 256QAM.)

When such a reception device is assumed, consideration of the Signal-to-Noise power Ratio (SNR) after separation of multiple signals indicates that a unitary matrix is appropriate as the precoding matrix when the reception device performs linear operation such as MMSE or ZF, whereas either a unitary matrix or a non-unitary matrix may be used when the reception device performs ML calculation. Taking any of the above embodiments into consideration, when two precoded signals are transmitted from two antennas, the same modulation method is used for two modulated signals (signals based on the modulation method before precoding), a non-unitary matrix is used as the precoding matrix in the method for regularly hopping between precoding matrices, the modulation level of the modulation method is equal to or less than 64 (or equal to or less than 256), and a unitary matrix is used when the modulation level is greater than 64 (or greater than 256), then for all of the modulation methods supported by the transmission system, there is an increased probability of achieving the advantageous effect whereby excellent data reception quality is achieved for any of the modulation methods while reducing the circuit scale of the reception device.

[0822]

When the modulation level of the modulation method is equal to or less than 64 (or equal to or less than 256) as well, in some cases use of a unitary matrix may be preferable. Based on this consideration, when a plurality of modulation methods
 5 are supported in which the modulation level is equal to or less than 64 (or equal to or less than 256), it is important that in some cases, in some of the plurality of supported modulation methods where the modulation level is equal to or less than 64, a non-unitary matrix is used as the precoding matrix in the method for regularly hopping between precoding matrices.

10 [0823]

The case of transmitting two precoded signals from two antennas has been described above as an example, but the present invention is not limited in this way. In the case when N precoded signals are transmitted from N antennas, and the same modulation method is used for N modulated signals (signals based on the
 15 modulation method before precoding), a threshold β_N may be established for the modulation level of the modulation method. When a plurality of modulation methods for which the modulation level is equal to or less than β_N are supported, in some of the plurality of supported modulation methods where the modulation level is equal to or less than β_N , a non-unitary matrix is used as the precoding matrices in
 20 the method for regularly hopping between precoding matrices, whereas for modulation methods for which the modulation level is greater than β_N , a unitary matrix is used. In this way, for all of the modulation methods supported by the transmission system, there is an increased probability of achieving the advantageous effect whereby excellent data reception quality is achieved for any of the modulation
 25 methods while reducing the circuit scale of the reception device. (When the modulation level of the modulation method is equal to or less than β_N , a non-unitary matrix may always be used as the precoding matrix in the method for regularly hopping between precoding matrices.)

In the above description, the same modulation method has been described as being used in the modulation method for simultaneously transmitting N modulated signals. The following, however, describes the case in which two or more modulation methods are used for simultaneously transmitting N modulated signals.

5 [0824]

As an example, the case in which two precoded signals are transmitted by two antennas is described. The two modulated signals (signals based on the modulation method before precoding) are either modulated with the same modulation method, or when modulated with different modulation methods, are
 10 modulated with a modulation method having a modulation level of 2^{a1} or a modulation level of 2^{a2} . In this case, when the reception device uses ML calculation ((Max-log) APP based on ML calculation), the number of candidate signal points in the IQ plane (received signal points 1101 in Fig. 11) is $2^{a1} \times 2^{a2} = 2^{a1 + a2}$. As described above, in order to achieve excellent data reception quality while reducing
 15 the circuit scale of the reception device, a threshold 2^β may be provided for $2^{a1 + a2}$, and when $2^{a1 + a2} \leq 2^\beta$, a non-unitary matrix may be used as the precoding matrix in the method for regularly hopping between precoding matrices, whereas a unitary matrix may be used when $2^{a1 + a2} > 2^\beta$.

[0825]

20 Furthermore, when $2^{a1 + a2} \leq 2^\beta$, in some cases use of a unitary matrix may be preferable. Based on this consideration, when a plurality of combinations of modulation methods are supported for which $2^{a1 + a2} \leq 2^\beta$, it is important that in some of the supported combinations of modulation methods for which $2^{a1 + a2} \leq 2^\beta$, a non-unitary matrix is used as the precoding matrix in the method for regularly
 25 hopping between precoding matrices.

[0826]

As an example, the case in which two precoded signals are transmitted by two antennas has been described, but the present invention is not limited in this way.

For example, N modulated signals (signals based on the modulation method before precoding) may be either modulated with the same modulation method or, when modulated with different modulation methods, the modulation level of the modulation method for the i^{th} modulated signal may be 2^{a_i} (where $i = 1, 2, \dots, N - 1, N$).
[0827]

In this case, when the reception device uses ML calculation ((Max-log) APP based on ML calculation), the number of candidate signal points in the IQ plane (received signal points 1101 in Fig. 11) is $2^{a_1} \times 2^{a_2} \times \dots \times 2^{a_i} \times \dots \times 2^{a_N} = 2^{a_1 + a_2 + \dots + a_i + \dots + a_N}$. As described above, in order to achieve excellent data reception quality while reducing the circuit scale of the reception device, a threshold 2^β may be provided for $2^{a_1 + a_2 + \dots + a_i + \dots + a_N}$.
[0828]

Math 289

15 Condition #44

$$2^{a_1 + a_2 + \dots + a_i + \dots + a_N} = 2^Y \leq 2^\beta$$

where

$$Y = \sum_{i=1}^N a_i$$

20 [0829]

When a plurality of combinations of a modulation methods satisfying Condition #44 are supported, in some of the supported combinations of modulation methods satisfying Condition #44, a non-unitary matrix are used as the precoding matrix in the method for regularly hopping between precoding matrices.

25 [0830]

Math 290

Condition #45

$$2^{a_1+a_2+\dots+a_i+\dots+a_N} = 2^Y > 2^\beta$$

where

$$Y = \sum_{i=1}^N a_i$$

5

[0831]

By using a unitary matrix in all of the combinations of modulation methods satisfying Condition #45, then for all of the modulation methods supported by the transmission system, there is an increased probability of achieving the advantageous effect whereby excellent data reception quality is achieved while reducing the circuit scale of the reception device for any of the combinations of modulation methods. (A non-unitary matrix may be used as the precoding matrix in the method for regularly hopping between precoding matrices in all of the supported combinations of modulation methods satisfying Condition #44.)

15 (Embodiment 15)

The present embodiment describes an example of a system that adopts a method for regularly hopping between precoding matrices using a multi-carrier transmission method such as OFDM.

[0832]

20 Figs. 47A and 47B show an example according to the present embodiment of frame structure in the time and frequency domains for a signal transmitted by a broadcast station (base station) in a system that adopts a method for regularly hopping between precoding matrices using a multi-carrier transmission method such as OFDM. (The frame structure is set to extend from time \$1 to time \$T.) Fig. 47A shows the frame structure in the time and frequency domains for the stream s1

25 described in Embodiment 1, and Fig. 47B shows the frame structure in the time and

frequency domains for the stream s2 described in Embodiment 1. Symbols at the same time and the same (sub)carrier in stream s1 and stream s2 are transmitted by a plurality of antennas at the same time and the same frequency.

[0833]

5 In Figs. 47A and 47B, the (sub)carriers used when using OFDM are divided as follows: a carrier group #A composed of (sub)carrier $a - (sub)carrier a + N_a$, a carrier group #B composed of (sub)carrier $b - (sub)carrier b + N_b$, a carrier group #C composed of (sub)carrier $c - (sub)carrier c + N_c$, a carrier group #D composed of (sub)carrier $d - (sub)carrier d + N_d$, ... In each subcarrier group, a plurality of
10 transmission methods are assumed to be supported. By supporting a plurality of transmission methods, it is possible to effectively capitalize on the advantages of the transmission methods. For example, in Figs. 47A and 47B, a spatial multiplexing MIMO system, or a MIMO system with a fixed precoding matrix is used for carrier group #A, a MIMO system that regularly hops between precoding matrices is used
15 for carrier group #B, only stream s1 is transmitted in carrier group #C, and space-time block coding is used to transmit carrier group #D.

[0834]

Figs. 48A and 48B show an example according to the present embodiment of frame structure in the time and frequency domains for a signal transmitted by a
20 broadcast station (base station) in a system that adopts a method for regularly hopping between precoding matrices using a multi-carrier transmission method such as OFDM. Figs. 48A and 48B show a frame structure at a different time than Figs. 47A and 47B, from time X to time $X + T'$. In Figs. 48A and 48B, as in Figs. 47A and 47B, the (sub)carriers used when using OFDM are divided as follows: a carrier
25 group #A composed of (sub)carrier $a - (sub)carrier a + N_a$, a carrier group #B composed of (sub)carrier $b - (sub)carrier b + N_b$, a carrier group #C composed of (sub)carrier $c - (sub)carrier c + N_c$, a carrier group #D composed of (sub)carrier $d - (sub)carrier d + N_d$, ... The difference between Figs. 47A and 47B and Figs. 48A

and 48B is that in some carrier groups, the transmission method used in Figs. 47A and 47B differs from the transmission method used in Figs. 48A and 48B. In Figs. 48A and 48B, space-time block coding is used to transmit carrier group #A, a MIMO system that regularly hops between precoding matrices is used for carrier group #B, a MIMO system that regularly hops between precoding matrices is used for carrier group #C, and only stream s1 is transmitted in carrier group #D.

[0835]

Next, the supported transmission methods are described.

[0836]

Fig. 49 shows a signal processing method when using a spatial multiplexing MIMO system or a MIMO system with a fixed precoding matrix. Fig. 49 bears the same numbers as in Fig. 6.

[0837]

A weighting unit 600, which is a baseband signal in accordance with a certain modulation method, receives as inputs a stream s1(t) (307A), a stream s2(t) (307B), and information 315 regarding the weighting method, and outputs a modulated signal z1(t) (309A) after weighting and a modulated signal z2(t) (309B) after weighting. Here, when the information 315 regarding the weighting method indicates a spatial multiplexing MIMO system, the signal processing in method #1 of Fig. 49 is performed. Specifically, the following processing is performed.

[0838]

Math 291

Equation 250

$$\begin{aligned} \begin{pmatrix} z1(t) \\ z2(t) \end{pmatrix} &= \begin{pmatrix} e^{j0} & 0 \\ 0 & e^{j0} \end{pmatrix} \begin{pmatrix} s1(t) \\ s2(t) \end{pmatrix} \\ &= \begin{pmatrix} 1 & 0 \\ 0 & 1 \end{pmatrix} \begin{pmatrix} s1(t) \\ s2(t) \end{pmatrix} = \begin{pmatrix} s1(t) \\ s2(t) \end{pmatrix} \end{aligned}$$

[0839]

When a method for transmitting one modulated signal is supported, from the standpoint of transmission power, Equation 250 may be represented as Equation 251.

[0840]

Math 292

Equation 251

$$\begin{aligned} \begin{pmatrix} z1(t) \\ z2(t) \end{pmatrix} &= \frac{1}{\sqrt{2}} \begin{pmatrix} e^{j0} & 0 \\ 0 & e^{j0} \end{pmatrix} \begin{pmatrix} s1(t) \\ s2(t) \end{pmatrix} \\ &= \frac{1}{\sqrt{2}} \begin{pmatrix} 1 & 0 \\ 0 & 1 \end{pmatrix} \begin{pmatrix} s1(t) \\ s2(t) \end{pmatrix} = \begin{pmatrix} \frac{1}{\sqrt{2}} s1(t) \\ \frac{1}{\sqrt{2}} s2(t) \end{pmatrix} \end{aligned}$$

10

[0841]

When the information 315 regarding the weighting method indicates a MIMO system in which precoding matrices are regularly hopped between, signal processing in method #2, for example, of Fig. 49 is performed. Specifically, the following processing is performed.

[0842]

Math 293

Equation 252

$$\begin{pmatrix} z1(t) \\ z2(t) \end{pmatrix} = \frac{1}{\sqrt{\alpha^2 + 1}} \begin{pmatrix} e^{j\theta_{11}} & \alpha \times e^{j(\theta_{11} + \lambda)} \\ \alpha \times e^{j\theta_{21}} & e^{j(\theta_{21} + \lambda + \delta)} \end{pmatrix} \begin{pmatrix} s1(t) \\ s2(t) \end{pmatrix}$$

20

[0843]

Here, θ_{11} , θ_{12} , λ , and δ are fixed values.

[0844]

Fig. 50 shows the structure of modulated signals when using space-time
 5 block coding. A space-time block coding unit (5002) in Fig. 50 receives, as input, a
 baseband signal based on a certain modulation signal. For example, the space-time
 block coding unit (5002) receives symbol s_1 , symbol s_2 , ... as inputs. As shown in
 Fig. 50, space-time block coding is performed, z_1 (5003A) becomes “ s_1 as symbol
 #0”, “ $-s_2^*$ as symbol #0”, “ s_3 as symbol #2”, “ $-s_4^*$ as symbol #3”..., and z_2 (5003B)
 10 becomes “ s_2 as symbol #0”, “ s_1^* as symbol #1”, “ s_4 as symbol #2”, “ s_3^* as symbol
 #3”.... In this case, symbol #X in z_1 and symbol #X in z_2 are transmitted from the
 antennas at the same time, over the same frequency.

[0845]

In Figs. 47A, 47B, 48A, and 48B, only symbols transmitting data are shown.
 15 In practice, however, it is necessary to transmit information such as the transmission
 method, modulation method, error correction method, and the like. For example, as
 in Fig. 51, these pieces of information can be transmitted to a communication
 partner by regular transmission with only one modulated signal z_1 . It is also
 necessary to transmit symbols for estimation of channel fluctuation, i.e. for the
 20 reception device to estimate channel fluctuation (for example, a pilot symbol,
 reference symbol, preamble, a Phase Shift Keying (PSK) symbol known at the
 transmission and reception sides, and the like). In Figs. 47A, 47B, 48A, and 48B,
 these symbols are omitted. In practice, however, symbols for estimating channel
 fluctuation are included in the frame structure in the time and frequency domains.
 25 Accordingly, each carrier group is not composed only of symbols for transmitting
 data. (The same is true for Embodiment 1 as well.)

Fig. 52 is an example of the structure of a transmission device in a broadcast
 station (base station) according to the present embodiment. A transmission method

determining unit (5205) determines the number of carriers, modulation method, error correction method, coding ratio for error correction coding, transmission method, and the like for each carrier group and outputs a control signal (5206).

[0846]

5 A modulated signal generating unit #1 (5201_1) receives, as input, information (5200_1) and the control signal (5206) and, based on the information on the transmission method in the control signal (5206), outputs a modulated signal z1 (5202_1) and a modulated signal z2 (5203_1) in the carrier group #A of Figs. 47A, 47B, 48A, and 48B.

10 [0847]

 Similarly, a modulated signal generating unit #2 (5201_2) receives, as input, information (5200_2) and the control signal (5206) and, based on the information on the transmission method in the control signal (5206), outputs a modulated signal z1 (5202_2) and a modulated signal z2 (5203_2) in the carrier group #B of Figs. 47A, 15 47B, 48A, and 48B.

[0848]

 Similarly, a modulated signal generating unit #3 (5201_3) receives, as input, information (5200_3) and the control signal (5206) and, based on the information on the transmission method in the control signal (5206), outputs a modulated signal z1 20 (5202_3) and a modulated signal z2 (5203_3) in the carrier group #C of Figs. 47A, 47B, 48A, and 48B.

[0849]

 Similarly, a modulated signal generating unit #4 (5201_4) receives, as input, information (5200_4) and the control signal (5206) and, based on the information on 25 the transmission method in the control signal (5206), outputs a modulated signal z1 (5202_4) and a modulated signal z2 (5203_4) in the carrier group #D of Figs. 47A, 47B, 48A, and 48B.

[0850]

While not shown in the figures, the same is true for modulated signal generating unit #5 through modulated signal generating unit #M – 1.

[0851]

Similarly, a modulated signal generating unit #M (5201_M) receives, as
5 input, information (5200_M) and the control signal (5206) and, based on the information on the transmission method in the control signal (5206), outputs a modulated signal z1 (5202_M) and a modulated signal z2 (5203_M) in a certain carrier group.

[0852]

10 An OFDM related processor (5207_1) receives, as inputs, the modulated signal z1 (5202_1) in carrier group #A, the modulated signal z1 (5202_2) in carrier group #B, the modulated signal z1 (5202_3) in carrier group #C, the modulated signal z1 (5202_4) in carrier group #D, ..., the modulated signal z1 (5202_M) in a certain carrier group #M, and the control signal (5206), performs processing such as
15 reordering, inverse Fourier transform, frequency conversion, amplification, and the like, and outputs a transmission signal (5208_1). The transmission signal (5208_1) is output as a radio wave from an antenna (5209_1).

[0853]

Similarly, an OFDM related processor (5207_2) receives, as inputs, the
20 modulated signal z1 (5203_1) in carrier group #A, the modulated signal z1 (5203_2) in carrier group #B, the modulated signal z1 (5203_3) in carrier group #C, the modulated signal z1 (5203_4) in carrier group #D, ..., the modulated signal z1 (5203_M) in a certain carrier group #M, and the control signal (5206), performs processing such as reordering, inverse Fourier transform, frequency conversion,
25 amplification, and the like, and outputs a transmission signal (5208_2). The transmission signal (5208_2) is output as a radio wave from an antenna (5209_2).

[0854]

Fig. 53 shows an example of a structure of the modulated signal generating units #1—#M in Fig. 52. An error correction encoder (5302) receives, as inputs, information (5300) and a control signal (5301) and, in accordance with the control signal (5301), sets the error correction coding method and the coding ratio for error correction coding, performs error correction coding, and outputs data (5303) after error correction coding. (In accordance with the setting of the error correction coding method and the coding ratio for error correction coding, when using LDPC coding, turbo coding, or convolutional coding, for example, depending on the coding ratio, puncturing may be performed to achieve the coding ratio.)

10 An interleaver (5304) receives, as input, error correction coded data (5303) and the control signal (5301) and, in accordance with information on the interleaving method included in the control signal (5301), reorders the error correction coded data (5303) and outputs interleaved data (5305).

[0855]

15 A mapper (5306_1) receives, as input, the interleaved data (5305) and the control signal (5301) and, in accordance with the information on the modulation method included in the control signal (5301), performs mapping and outputs a baseband signal (5307_1).

[0856]

20 Similarly, a mapper (5306_2) receives, as input, the interleaved data (5305) and the control signal (5301) and, in accordance with the information on the modulation method included in the control signal (5301), performs mapping and outputs a baseband signal (5307_2).

[0857]

25 A signal processing unit (5308) receives, as input, the baseband signal (5307_1), the baseband signal (5307_2), and the control signal (5301) and, based on information on the transmission method (for example, in this embodiment, a spatial multiplexing MIMO system, a MIMO method using a fixed precoding matrix, a

MIMO method for regularly hopping between precoding matrices, space-time block coding, or a transmission method for transmitting only stream s1) included in the control signal (5301), performs signal processing. The signal processing unit (5308) outputs a processed signal z1 (5309_1) and a processed signal z2 (5309_2). Note
 5 that when the transmission method for transmitting only stream s1 is selected, the signal processing unit (5308) does not output the processed signal z2 (5309_2). Furthermore, in Fig. 53, one error correction encoder is shown, but the present invention is not limited in this way. For example, as shown in Fig. 3, a plurality of encoders may be provided.

10 [0858]

Fig. 54 shows an example of the structure of the OFDM related processors (5207_1 and 5207_2) in Fig. 52. Elements that operate in a similar way to Fig. 14 bear the same reference signs. A reordering unit (5402A) receives, as input, the modulated signal z1 (5400_1) in carrier group #A, the modulated signal z1 (5400_2) in carrier group #B, the modulated signal z1 (5400_3) in carrier group #C, the
 15 modulated signal z1 (5400_4) in carrier group #D, ..., the modulated signal z1 (5400_M) in a certain carrier group, and a control signal (5403), performs reordering, and output reordered signals 1405A and 1405B. Note that in Figs. 47A, 47B, 48A, 48B, and 51, an example of allocation of the carrier groups is described as being
 20 formed by groups of subcarriers, but the present invention is not limited in this way. Carrier groups may be formed by discrete subcarriers at each time interval. Furthermore, in Figs. 47A, 47B, 48A, 48B, and 51, an example has been described in which the number of carriers in each carrier group does not change over time, but the present invention is not limited in this way. This point will be described
 25 separately below.

[0859]

Figs. 55A and 55B show an example of frame structure in the time and frequency domains for a method of setting the transmission method for each carrier

group, as in Figs. 47A, 47B, 48A, 48B, and 51. In Figs. 55A and 55B, control information symbols are labeled 5500, individual control information symbols are labeled 5501, data symbols are labeled 5502, and pilot symbols are labeled 5503. Furthermore, Fig. 55A shows the frame structure in the time and frequency domains for stream s1, and Fig. 55B shows the frame structure in the time and frequency domains for stream s2.

[0860]

The control information symbols are for transmitting control information shared by the carrier group and are composed of symbols for the transmission and reception devices to perform frequency and time synchronization, information regarding the allocation of (sub)carriers, and the like. The control information symbols are set to be transmitted from only stream s1 at time \$1.

[0861]

The individual control information symbols are for transmitting control information on individual subcarrier groups and are composed of information on the transmission method, modulation method, error correction coding method, coding ratio for error correction coding, block size of error correction codes, and the like for the data symbols, information on the insertion method of pilot symbols, information on the transmission power of pilot symbols, and the like. The individual control information symbols are set to be transmitted from only stream s1 at time \$1.

[0862]

The data symbols are for transmitting data (information), and as described with reference to Figs. 47A through 50, are symbols of one of the following transmission methods, for example: a spatial multiplexing MIMO system, a MIMO method using a fixed precoding matrix, a MIMO method for regularly hopping between precoding matrices, space-time block coding, or a transmission method for transmitting only stream s1. Note that in carrier group #A, carrier group #B, carrier group #C, and carrier group #D, data symbols are shown in stream s2, but when the

transmission method for transmitting only stream s1 is used, in some cases there are no data symbols in stream s2.

[0863]

The pilot symbols are for the reception device to perform channel
 5 estimation, i.e. to estimate fluctuation corresponding to $h_{11}(t)$, $h_{12}(t)$, $h_{21}(t)$, and $h_{22}(t)$
 in Equation 36. (In this embodiment, since a multi-carrier transmission method such
 as an OFDM method is used, the pilot symbols are for estimating fluctuation
 corresponding to $h_{11}(t)$, $h_{12}(t)$, $h_{21}(t)$, and $h_{22}(t)$ in each subcarrier.) Accordingly, the
 PSK transmission method, for example, is used for the pilot symbols, which are
 10 structured to form a pattern known by the transmission and reception devices.
 Furthermore, the reception device may use the pilot symbols for estimation of
 frequency offset, estimation of phase distortion, and time synchronization.

[0864]

Fig. 56 shows an example of the structure of a reception device for
 15 receiving modulated signals transmitted by the transmission device in Fig. 52.
 Elements that operate in a similar way to Fig. 7 bear the same reference signs.

[0865]

In Fig. 56, an OFDM related processor (5600_X) receives, as input, a
 received signal 702_X, performs predetermined processing, and outputs a processed
 20 signal 704_X. Similarly, an OFDM related processor (5600_Y) receives, as input, a
 received signal 702_Y, performs predetermined processing, and outputs a processed
 signal 704_Y.

[0866]

The control information decoding unit 709 in Fig. 56 receives, as input, the
 25 processed signals 704_X and 704_Y, extracts the control information symbols and
 individual control information symbols in Figs. 55A and 55B to obtain the control
 information transmitted by these symbols, and outputs a control signal 710 that
 includes the obtained information.

[0867]

The channel fluctuation estimating unit 705_1 for the modulated signal z1 receives, as inputs, the processed signal 704_X and the control signal 710, performs channel estimation in the carrier group required by the reception device (the desired carrier group), and outputs a channel estimation signal 706_1.

[0868]

Similarly, the channel fluctuation estimating unit 705_2 for the modulated signal z2 receives, as inputs, the processed signal 704_X and the control signal 710, performs channel estimation in the carrier group required by the reception device (the desired carrier group), and outputs a channel estimation signal 706_2.

[0869]

Similarly, the channel fluctuation estimating unit 705_1 for the modulated signal z1 receives, as inputs, the processed signal 704_Y and the control signal 710, performs channel estimation in the carrier group required by the reception device (the desired carrier group), and outputs a channel estimation signal 708_1.

[0870]

Similarly, the channel fluctuation estimating unit 705_2 for the modulated signal z2 receives, as inputs, the processed signal 704_Y and the control signal 710, performs channel estimation in the carrier group required by the reception device (the desired carrier group), and outputs a channel estimation signal 708_2.

[0871]

The signal processing unit 711 receives, as inputs, the signals 706_1, 706_2, 708_1, 708_2, 704_X, 704_Y, and the control signal 710. Based on the information included in the control signal 710 on the transmission method, modulation method, error correction coding method, coding ratio for error correction coding, block size of error correction codes, and the like for the data symbols transmitted in the desired carrier group, the signal processing unit 711 demodulates and decodes the data symbols and outputs received data 712.

[0872]

Fig. 57 shows the structure of the OFDM related processors (5600_X, 5600_Y) in Fig. 56. A frequency converter (5701) receives, as input, a received signal (5700), performs frequency conversion, and outputs a frequency converted
5 signal (5702).

[0873]

A Fourier transformer (5703) receives, as input, the frequency converted signal (5702), performs a Fourier transform, and outputs a Fourier transformed signal (5704).

10 [0874]

As described above, when using a multi-carrier transmission method such as an OFDM method, carriers are divided into a plurality of carrier groups, and the transmission method is set for each carrier group, thereby allowing for the reception quality and transmission speed to be set for each carrier group, which yields the
15 advantageous effect of construction of a flexible system. In this case, as described in other embodiments, allowing for choice of a method of regularly hopping between precoding matrices offers the advantages of obtaining high reception quality, as well as high transmission speed, in an LOS environment. While in the present embodiment, the transmission methods to which a carrier group can be set are “a
20 spatial multiplexing MIMO system, a MIMO method using a fixed precoding matrix, a MIMO method for regularly hopping between precoding matrices, space-time block coding, or a transmission method for transmitting only stream s1”, but the transmission methods are not limited in this way. Furthermore, the space-time coding is not limited to the method described with reference to Fig. 50, nor is the
25 MIMO method using a fixed precoding matrix limited to method #2 in Fig. 49, as any structure with a fixed precoding matrix is acceptable. In the present embodiment, the case of two antennas in the transmission device has been described, but when the number of antennas is larger than two as well, the same advantageous effects may be

achieved by allowing for selection of a transmission method for each carrier group from among “a spatial multiplexing MIMO system, a MIMO method using a fixed precoding matrix, a MIMO method for regularly hopping between precoding matrices, space-time block coding, or a transmission method for transmitting only
5 stream s1”.

[0875]

Figs. 58A and 58B show a method of allocation into carrier groups that differs from Figs. 47A, 47B, 48A, 48B, and 51. In Figs. 47A, 47B, 48A, 48B, 51, 55A, and 55B, carrier groups have described as being formed by groups of
10 subcarriers. In Figs. 58A and 58B, on the other hand, the carriers in a carrier group are arranged discretely. Figs. 58A and 58B show an example of frame structure in the time and frequency domains that differs from Figs. 47A, 47B, 48A, 48B, 51, 55A, and 55B. Figs. 58A and 58B show the frame structure for carriers 1 through H, times \$1 through \$K. Elements that are similar to Figs. 55A and 55B bear the same
15 reference signs. Among the data symbols in Figs. 58A and 58B, the “A” symbols are symbols in carrier group A, the “B” symbols are symbols in carrier group B, the “C” symbols are symbols in carrier group C, and the “D” symbols are symbols in carrier group D. The carrier groups can thus be similarly implemented by discrete arrangement along (sub)carriers, and the same carrier need not always be used in the
20 time domain. This type of arrangement yields the advantageous effect of obtaining time and frequency diversity gain.

[0876]

In Figs. 47A, 47B, 48A, 48B, 51, 58A, and 58B, the control information symbols and the individual control information symbols are allocated to the same
25 time in each carrier group, but these symbols may be allocated to different times. Furthermore, the number of (sub)carriers used by a carrier group may change over time.

[0877]

(Embodiment 16)

Like Embodiment 10, the present embodiment describes a method for regularly hopping between precoding matrices using a unitary matrix when N is an odd number.

5 [0878]

In the method of regularly hopping between precoding matrices over a period (cycle) with 2N slots, the precoding matrices prepared for the 2N slots are represented as follows.

[0879]

10 Math 294

Equation 253

for $i = 0, 1, 2, \dots, N-2, N-1$:

$$F[i] = \frac{1}{\sqrt{\alpha^2 + 1}} \begin{pmatrix} e^{j\theta_{11}(i)} & \alpha \times e^{j(\theta_{11}(i) + \lambda)} \\ \alpha \times e^{j\theta_{21}(i)} & e^{j(\theta_{21}(i) + \lambda + \pi)} \end{pmatrix}$$

15 [0880]

Let α be a fixed value (not depending on i), where $\alpha > 0$.

[0881]

Math 295

Equation 254

20 for $i = N, N+1, N+2, \dots, 2N-2, 2N-1$:

$$F[i] = \frac{1}{\sqrt{\alpha^2 + 1}} \begin{pmatrix} \alpha \times e^{j\theta_{11}(i)} & e^{j(\theta_{11}(i) + \lambda)} \\ e^{j\theta_{21}(i)} & \alpha \times e^{j(\theta_{21}(i) + \lambda + \pi)} \end{pmatrix}$$

[0882]

Let α be a fixed value (not depending on i), where $\alpha > 0$. (Let the α in Equation 253 and the α in Equation 254 be the same value.)

From Condition #5 (Math 106) and Condition #6 (Math 107) in Embodiment 3, the following conditions are important in Equation 253 for achieving
5 excellent data reception quality.

[0883]

Math 296

Condition #46

$$e^{j(\theta_{11}(x)-\theta_{21}(x))} \neq e^{j(\theta_{11}(y)-\theta_{21}(y))} \quad \text{for } \forall x, \forall y (x \neq y; x, y = 0, 1, 2, \dots, N-2, N-1)$$

10

[0884]

(x is 0, 1, 2, ..., $N-2$, $N-1$; y is 0, 1, 2, ..., $N-2$, $N-1$; and $x \neq y$.)

[0885]

Math 297

15 Condition #47

$$e^{j(\theta_{11}(x)-\theta_{21}(x)-\pi)} \neq e^{j(\theta_{11}(y)-\theta_{21}(y)-\pi)} \quad \text{for } \forall x, \forall y (x \neq y; x, y = 0, 1, 2, \dots, N-2, N-1)$$

[0886]

(x is 0, 1, 2, ..., $N-2$, $N-1$; y is 0, 1, 2, ..., $N-2$, $N-1$; and $x \neq y$.)

20

Addition of the following condition is considered.

[0887]

Math 298

Condition #48

$$\theta_{11}(x) = \theta_{11}(x+N) \quad \text{for } \forall x (x = 0, 1, 2, \dots, N-2, N-1)$$

25 and

$$\theta_{21}(y) = \theta_{21}(y + N) \quad \text{for } \forall y (y = 0, 1, 2, \dots, N-2, N-1)$$

[0888]

Next, in order to distribute the poor reception points evenly with regards to
 5 phase in the complex plane, as described in Embodiment 6, Condition #49 and
 Condition #50 are provided.

[0889]

Math 299

Condition #49

$$10 \quad \frac{e^{j(\theta_{11}(x+1) - \theta_{21}(x+1))}}{e^{j(\theta_{11}(x) - \theta_{21}(x))}} = e^{j\left(\frac{2\pi}{N}\right)} \quad \text{for } \forall x (x = 0, 1, 2, \dots, N-2)$$

[0890]

Math 300

Condition #50

$$15 \quad \frac{e^{j(\theta_{11}(x+1) - \theta_{21}(x+1))}}{e^{j(\theta_{11}(x) - \theta_{21}(x))}} = e^{j\left(-\frac{2\pi}{N}\right)} \quad \text{for } \forall x (x = 0, 1, 2, \dots, N-2)$$

[0891]

In other words, Condition #49 means that the difference in phase is $2\pi/N$
 radians. On the other hand, Condition #50 means that the difference in phase is
 20 $-2\pi/N$ radians.

[0892]

Letting $\theta_{11}(0) - \theta_{21}(0) = 0$ radians, and letting $\alpha > 1$, the distribution of poor
 reception points for s1 and for s2 in the complex plane for $N = 3$ is shown in Figs.

60A and 60B. As is clear from Figs. 60A and 60B, in the complex plane, the minimum distance between poor reception points for s_1 is kept large, and similarly, the minimum distance between poor reception points for s_2 is also kept large. Similar conditions are created when $\alpha < 1$. Furthermore, upon comparison with Figs. 45A and 45B in Embodiment 10, making the same considerations as in Embodiment 9, the probability of a greater distance between poor reception points in the complex plane increases when N is an odd number as compared to when N is an even number. However, when N is small, for example when $N \leq 16$, the minimum distance between poor reception points in the complex plane can be guaranteed to be a certain length, since the number of poor reception points is small. Accordingly, when $N \leq 16$, even if N is an even number, cases do exist where data reception quality can be guaranteed.

[0893]

Therefore, in the method for regularly hopping between precoding matrices based on Equations 253 and 254, when N is set to an odd number, the probability of improving data reception quality is high. Precoding matrices $F[0]$ – $F[2N - 1]$ are generated based on Equations 253 and 254 (the precoding matrices $F[0]$ – $F[2N - 1]$ may be in any order for the $2N$ slots in the period (cycle)). Symbol number $2Ni$ may be precoded using $F[0]$, symbol number $2Ni + 1$ may be precoded using $F[1]$, ..., and symbol number $2N \times i + h$ may be precoded using $F[h]$, for example ($h = 0, 1, 2, \dots, 2N - 2, 2N - 1$). (In this case, as described in previous embodiments, precoding matrices need not be hopped between regularly.) Furthermore, when the modulation method for both s_1 and s_2 is 16QAM, if α is set as in Equation 233, the advantageous effect of increasing the minimum distance between $16 \times 16 = 256$ signal points in the IQ plane for a specific LOS environment may be achieved.

[0894]

The following conditions are possible as conditions differing from Condition #48:

[0895]

Math 301

Condition #51

$$e^{j(\theta_{11}(x)-\theta_{21}(x))} \neq e^{j(\theta_{11}(y)-\theta_{21}(y))} \quad \text{for } \forall x, \forall y (x \neq y; x, y = N, N+1, N+2, \dots, 2N-2, 2N-1)$$

5

[0896]

(where x is N, N + 1, N + 2, ..., 2N - 2, 2N - 1; y is N, N + 1, N + 2, ..., 2N - 2, 2N - 1; and x ≠ y.)

[0897]

10 Math 302

Condition #52

$$e^{j(\theta_{11}(x)-\theta_{21}(x)-\pi)} \neq e^{j(\theta_{11}(y)-\theta_{21}(y)-\pi)} \quad \text{for } \forall x, \forall y (x \neq y; x, y = N, N+1, N+2, \dots, 2N-2, 2N-1)$$

[0898]

15 (where x is N, N + 1, N + 2, ..., 2N - 2, 2N - 1; y is N, N + 1, N + 2, ..., 2N - 2, 2N - 1; and x ≠ y.)

In this case, by satisfying Condition #46, Condition #47, Condition #51, and Condition #52, the distance in the complex plane between poor reception points for s1 is increased, as is the distance between poor reception points for s2, thereby achieving excellent data reception quality.

20 [0899]

In the present embodiment, the method of structuring 2N different precoding matrices for a precoding hopping method with a 2N-slot time period (cycle) has been described. In this case, as the 2N different precoding matrices, F[0], F[1], F[2], ..., F[2N - 2], F[2N - 1] are prepared. In the present embodiment, an example of a single carrier transmission method has been described, and therefore the case of arranging symbols in the order F[0], F[1], F[2], ..., F[2N - 2], F[2N - 1] in the time domain (or the frequency domain) has been described. The present

invention is not, however, limited in this way, and the $2N$ different precoding matrices $F[0]$, $F[1]$, $F[2]$, ..., $F[2N - 2]$, $F[2N - 1]$ generated in the present embodiment may be adapted to a multi-carrier transmission method such as an OFDM transmission method or the like. As in Embodiment 1, as a method of
 5 adaption in this case, precoding weights may be changed by arranging symbols in the frequency domain and in the frequency-time domain. Note that a precoding hopping method with a $2N$ -slot time period (cycle) has been described, but the same advantageous effects may be obtained by randomly using $2N$ different precoding matrices. In other words, the $2N$ different precoding matrices do not necessarily
 10 need to be used in a regular period (cycle).

[0900]

Furthermore, in the precoding matrix hopping method over an H -slot period (cycle) (H being a natural number larger than the number of slots $2N$ in the period (cycle) of the above method of regularly hopping between precoding matrices),
 15 when the $2N$ different precoding matrices of the present embodiment are included, the probability of excellent reception quality increases.

[0901]

(Embodiment A1)

In the present Embodiment, data is transmitted hierarchically, and a
 20 transmission method adopting the method of regularly switching between precoding matrices described in Embodiments 1–16 is described in detail.

[0902]

Figs. 61 and 62 are an example, according to the present embodiment, of the structure of a transmission device in a broadcast station. An error correction encoder
 25 (6101_1) for a base stream (base layer) receives information (6100_1) of the base stream (base layer) as input, performs error correction coding, and outputs encoded information (6102_1) of the base stream (base layer).

[0903]

An error correction encoder (6101_2) for an enhancement stream (enhancement layer) receives information (6100_2) of the enhancement stream (enhancement layer) as input, performs error correction coding, and outputs encoded information (6102_2) of the enhancement stream (enhancement layer).

5 [0904]

An interleaver (6103_1) receives the encoded information (6102_1) of the base stream (base layer) as input, applies interleaving, and outputs interleaved, encoded data (6104_1).

[0905]

10 Similarly, an interleaver (6103_2) receives the encoded information (6102_2) on the enhancement stream (enhancement layer) as input, applies interleaving, and outputs interleaved, encoded data (6104_2).

[0906]

A mapper (6105_1) receives the interleaved, encoded data (6104_1) and an
 15 information signal regarding the transmission method (6111) as input, performs modulation in accordance with a predetermined modulation method based on the transmission method indicated by the information signal regarding the transmission method (6111), and outputs a baseband signal (6106_1) (corresponding to $s_1(t)$ (307A) in Fig. 3) and a baseband signal (6106_2) (corresponding to $s_2(t)$ (307B) in
 20 Fig. 3). The information (6111) regarding the transmission method is, for example, information such as the transmission system for hierarchical transmission (the modulation method, the transmission method, and information on precoding matrices used when adopting a transmission method that regularly switches between precoding matrices), the error correction coding method (type of coding, coding
 25 rate), and the like.

[0907]

Similarly, a mapper (6105_2) receives the interleaved, encoded data (6104_2) and the information signal regarding the transmission method (6111) as

input, performs modulation in accordance with a predetermined modulation method based on the transmission method indicated by the information signal regarding the transmission method (6111), and outputs a baseband signal (6107_1) (corresponding to $s_1(t)$ (307A) in Fig. 3) and a baseband signal (6107_2) (corresponding to $s_2(t)$ (307B) in Fig. 3).

[0908]

A precoder (6108_1) receives the baseband signal (6106_1) (corresponding to $s_1(t)$ (307A) in Fig. 3), the baseband signal (6106_2) (corresponding to $s_2(t)$ (307B) in Fig. 3), and the information signal regarding the transmission method (6111) as input, performs precoding based on the method of regularly switching between precoding matrices as indicated by the information signal regarding the transmission method (6111), and outputs a precoded baseband signal (6109_1) (corresponding to $z_1(t)$ (309A) in Fig. 3) and a precoded baseband signal (6109_2) (corresponding to $z_2(t)$ (309B) in Fig. 3).

15 [0909]

Similarly, a precoder (6108_2) receives the baseband signal (6107_1) (corresponding to $s_1(t)$ (307A) in Fig. 3), the baseband signal (6107_2) (corresponding to $s_2(t)$ (307B) in Fig. 3), and the information signal regarding the transmission method (6111) as input, performs precoding based on the method of regularly switching between precoding matrices as indicated by the information signal regarding the transmission method (6111), and outputs a precoded baseband signal (6110_1) (corresponding to $z_1(t)$ (309A) in Fig. 3) and a precoded baseband signal (6110_2) (corresponding to $z_2(t)$ (309B) in Fig. 3).

[0910]

25 In Fig. 62, a reordering unit (6200_1) receives the precoded baseband signal (6109_1) and the precoded baseband signal (6110_1) as input, performs reordering, and outputs a reordered, precoded baseband signal (6201_1).

[0911]

Similarly, a reordering unit (6200_2) receives the precoded baseband signal (6109_2) and the precoded baseband signal (6110_2) as input, performs reordering, and outputs a reordered, precoded baseband signal (6201_2).

[0912]

5 An OFDM related processor (6202_1) receives the reordered, precoded baseband signal (6201_1), applies the signal processing described in Embodiment 1, and outputs a transmission signal (6203_1). The transmission signal (6203_1) is output from an antenna (6204_1).

[0913]

10 Similarly, an OFDM related processor (6202_2) receives the reordered, precoded baseband signal (6201_2), applies the signal processing described in Embodiment 1, and outputs a transmission signal (6203_2). The transmission signal (6203_2) is output from an antenna (6204_2).

[0914]

15 Fig. 63 illustrates operations of the precoder (6108_1) in Fig. 61. The precoder (6108_1) regularly switches between precoding matrices, and the structure and operations of the precoder (6108_1) are similar to the structure and operations described in Figs. 3, 6, 22, and the like. Since Fig. 61 illustrates the precoder (6108_1), Fig. 63 shows operations for weighting of the base stream (base layer). As
20 shown in Fig. 63, when the precoder 6108_1 performs weighting, i.e. when the precoder 6108_1 generates a precoded baseband signal by performing precoding, $z_1(t)$ and $z_2(t)$ are generated as a result of precoding that regularly switches between precoding matrices. The precoding of the base stream (base layer) is set to an eight-slot period (cycle) over which the precoding matrix is switched. The precoding
25 matrices for weighting are represented as $F[0]$, $F[1]$, $F[2]$, $F[3]$, $F[4]$, $F[5]$, $F[6]$, and $F[7]$. The symbols in the precoded signals $z_1(t)$ and $z_2(t)$ are represented as 6301 and 6302. In Fig. 63, a symbol is represented as “B #X F[Y]”, which refers to the X^{th}

symbol in the base stream (base layer) being precoded with the $F[Y]$ precoding matrix (where Y is any integer from 0 to 7).

[0915]

Fig. 64 illustrates operations of the precoder (6108_2) in Fig. 61. The precoder (6108_2) regularly switches between precoding matrices, and the structure and operations of the precoder (6108_2) are similar to the structure and operations described in Figs. 3, 6, 22, and the like. Since Fig. 61 illustrates the precoder (6108_2), Fig. 64 shows operations for weighting of the enhancement stream (enhancement layer). As shown in Fig. 64, when the precoder 6108_2 performs weighting, i.e. when the precoder 6108_2 generates a precoded baseband signal by performing precoding, $z_1(t)$ and $z_2(t)$ are generated as a result of precoding that regularly switches between precoding matrices. The precoding of the enhancement stream (enhancement layer) is set to a four-slot period (cycle) over which the precoding matrix is switched. The precoding matrices for weighting are represented as $f[0]$, $f[1]$, $f[2]$, and $f[3]$. The symbols in the precoded signals $z_1(t)$ and $z_2(t)$ are represented as 6403 and 6404. In Fig. 64, a symbol is represented as “E #X f[Y]”, which refers to the X^{th} symbol in the enhancement stream (enhancement layer) being precoded with the $f[Y]$ precoding matrix (where Y is any integer from 0 to 4).

[0916]

Figs. 65A and 65B show the method of reordering symbols in the reordering unit (6200_1) and the reordering unit (6200_2) in Fig. 62. The reordering unit (6200_1) and the reordering unit (6200_2) arrange symbols shown in Figs. 63 and 64 in the frequency and time domain as shown in Figs. 65A and 65B. During transmission, symbols in the same (sub)carrier and at the same time are transmitted at the same frequency and at the same time from different antennas. Note that the arrangement of symbols in the frequency and the time domains as shown in Figs. 65A and 65B is only an example. Symbols may be arranged based on the method described in Embodiment 1.

[0917]

When the base stream (base layer) and the enhancement stream (enhancement layer) are transmitted, it is necessary for the reception quality of data in the base stream (base layer) to be made higher than the reception quality of data in the enhancement stream (enhancement layer), due to the nature of the streams (layers). Therefore, as in the present embodiment, when using a method of regularly switching between precoding matrices, the modulation method when transmitting the base stream (base layer) is set to differ from the modulation method when transmitting the enhancement stream (enhancement layer). For example, it is possible to use one of modes #1–#5 as in Table 3.

[0918]

Table 3

Mode	Modulation method for base stream (layer)	Modulation method for enhancement stream (layer)
Mode #1	QPSK	16QAM
Mode #2	QPSK	64QAM
Mode #3	QPSK	256QAM
Mode #4	16QAM	64QAM
Mode #5	16QAM	256QAM

[0919]

By correspondingly setting the method of regularly switching between precoding matrices used when transmitting the base stream (base layer) to differ from the method of regularly switching between precoding matrices used when transmitting the enhancement stream (enhancement layer), it is possible for the reception quality of data in the reception device to improve, or to simplify the structure of the transmission device and the reception device. As an example, as

shown in Figs. 63 and 64, when using a method of modulating by modulation level (the number of signal points in the IQ plane), it may be better for methods of regularly switching between precoding matrices to differ. Therefore, a method for setting the periods (cycles) in the method of regularly switching between precoding matrices used when transmitting the base stream (base layer) to differ from the periods (cycles) in the method of regularly switching between precoding matrices used when transmitting the enhancement stream (enhancement layer) is effective, since this method for setting improves reception quality of data in the reception device or simplifies the structure of the transmission device and the reception device.

Alternatively, the method of structuring the precoding matrices in the method of regularly switching between precoding matrices used when transmitting the base stream (base layer) may be made to differ from the method of regularly switching between precoding matrices used when transmitting the enhancement stream (enhancement layer). Accordingly, the method of switching between precoding matrices is set as shown in Table 4 for each of the modes that can be set for the modulation methods of the streams (layers) in Table 3. (In Table 4, A, B, C, and D indicate different methods of switching between precoding matrices.)

[0920]

Table 4

Mode	Base stream (layer)		Extension stream (layer)	
	modulation method	method of switching between precoding matrices	modulation method	method of switching between precoding matrices
Mode #1	QPSK	A	16QAM	B
Mode #2	QPSK	A	64QAM	C

Mode #3	QPSK	A	256QAM	D
Mode #4	16QAM	B	64QAM	C
Mode #5	16QAM	B	256QAM	D

[0921]

Accordingly, in the transmission device for the broadcast station in Figs. 61 and 62, when the modulation method is switched in the mappers (6105_1 and 6105_2), the precoding method is switched in the precoders (6108_1 and 6108_2). Note that Table 4 is no more than an example. The method of switching between precoding matrices may be the same even if the modulation method differs. For example, the method of switching between precoding matrices may be the same for 64QAM and for 256QAM. The important point is that there be at least two methods of switching between precoding matrices when a plurality of modulation methods are supported. This point is not limited to use of hierarchical transmission; by establishing the above relationship between the modulation method and the method of switching between precoding matrices even when not using hierarchical transmission, it is possible for the reception quality of data in the reception device to improve, or to simplify the structure of the transmission device and the reception device.

[0922]

It is possible for a system not only to support hierarchical transmission exclusively, but also to support transmission that is not hierarchical. In this case, when transmission is not hierarchical, in Figs. 61 and 62, operations of the functional units related to the enhancement stream (enhancement layer) are stopped, and only the base stream (base layer) is transmitted. Table 5 corresponds to Table 4

and shows, for this case, correspondence between the settable mode, modulation method, and method of switching between precoding matrices.

[0923]

Table 5

Mode	Base stream (layer)		Extension stream (layer)	
	modulation method	method of switching between precoding matrices	modulation method	method of switching between precoding matrices
Mode #1	QPSK	A	16QAM	B
Mode #2	QPSK	A	64QAM	C
Mode #3	QPSK	A	256QAM	D
Mode #4	16QAM	B	64QAM	C
Mode #5	16QAM	B	256QAM	D
Mode #6	QPSK	A		
Mode #7	16QAM	B		
Mode #8	64QAM	C		
Mode #9	256QAM	D		
Mode #10	1024QA M	E		

5

[0924]

In Table 5, modes #1–#5 are the modes used for hierarchical transmission, and modes #6–#10 are the modes when transmission is not hierarchical. In this case, the method of switching between precoding matrices is set appropriately for each mode.

10

[0925]

Next, operations of the reception device when supporting hierarchical transmission are described. The structure of the reception device in the present Embodiment may be the structure in Fig. 7 described in Embodiment 1. In this case, the structure of the signal processing unit 711 of Fig. 7 is shown in Fig. 66.

5 [0926]

In Fig. 66, 6601X is a channel estimation signal corresponding to the channel estimation signal 706_1 in Fig. 7. 6602X is a channel estimation signal corresponding to the channel estimation signal 706_2 in Fig. 7. 6603X is a baseband signal corresponding to the baseband signal 704_X in Fig. 7. 6604 is a signal regarding information on the transmission method indicated by the transmission device and corresponds to the signal 710 regarding information on the transmission method indicated by the transmission device.

[0927]

6601Y is a channel estimation signal corresponding to the channel estimation signal 708_1 in Fig. 7. 6602Y is a channel estimation signal corresponding to the channel estimation signal 708_2 in Fig. 7. 6603Y is a baseband signal corresponding to the baseband signal 704_Y in Fig. 7.

[0928]

A signal sorting unit (6605) receives the channel estimation signals (6601X, 6602X, 6601Y, 6602Y), the baseband signals (6603X, 6603Y), and the signal regarding information on the transmission method indicated by the transmission device (6604) as input, and based on the signal regarding information on the transmission method indicated by the transmission device (6604), sorts the input into signals related to the base stream (base layer) and information of the enhancement stream (enhancement layer), outputting channel estimation signals for the base stream (6606_1, 6607_1, 6609_1, and 6610_1), baseband signals for the base stream (6608_1, 6611_1), channel estimation signals for the enhancement stream (6606_2,

6607_2, 6609_2, and 6610_2), and baseband signals for the enhancement stream (6608_2, 6611_2).

[0929]

A detection and log-likelihood ratio calculation unit (6612_1) is a
 5 processing unit for the base stream (base layer) that receives the channel estimation signals for the base stream (6606_1, 6607_1, 6609_1, and 6610_1), baseband signals for the base stream (6608_1, 6611_1), and the signal regarding information on the transmission method indicated by the transmission device (6604) as input, estimates the modulation method and the method of switching between precoding matrices
 10 used for the base stream (base layer) from the signal regarding information on the transmission method indicated by the transmission device (6604), and based on the modulation method and the method of switching, decodes the precoding, calculates the log-likelihood ratio for each bit, and outputs a log-likelihood ratio signal (6613_1). Note that the detection and log-likelihood ratio calculation unit (6612_1)
 15 performs detection and decoding of precoding and outputs a log-likelihood ratio signal even for modes #6–#10 for which no enhancement stream (enhancement layer) exists in Table 5.

[0930]

A detection and log-likelihood ratio calculation unit (6612_2) is a
 20 processing unit for the enhancement stream (enhancement layer) that receives the channel estimation signals for the enhancement stream (6606_2, 6607_2, 6609_2, and 6610_2), baseband signals for the enhancement stream (6608_2, 6611_2), and the signal regarding information on the transmission method indicated by the transmission device (6604) as input, estimates the modulation method and the
 25 method of switching between precoding matrices used for the enhancement stream (enhancement layer) from the signal regarding information on the transmission method indicated by the transmission device (6604), and based on the modulation method and the method of switching, decodes the precoding, calculates the

log-likelihood ratio for each bit, and outputs a log-likelihood ratio signal (6613_2). Note that operations are stopped for modes #6–#10 for which no enhancement stream (enhancement layer) exists in Table 5.

[0931]

5 In the transmission device described with reference to Figs. 61 and 62, only the method of hierarchical transmission has been described, but in practice, in addition to information on the method for hierarchical transmission, it is also necessary to transmit, to the reception device, information regarding the transmission method for hierarchical transmission (the modulation method, the
10 transmission method, and information on precoding matrices used when adopting a transmission method that regularly switches between precoding matrices), the error correction coding method (type of coding, coding rate), and the like. Furthermore, in the reception device, pilot symbols, reference symbols, and preambles for channel estimation (estimation of fluctuations in the channel), frequency synchronization,
15 frequency offset estimation, and signal detection have a frame structure existing in a separately transmitted signal. Note that this is true not only for Embodiment A1, but also for Embodiment A2 and subsequent embodiments.

[0932]

 A deinterleaver (6614_1) receives the log-likelihood ratio signal (6613_1)
20 as input, reorders the signal, and outputs a deinterleaved log-likelihood ratio signal (6615_1).

[0933]

 Similarly, a deinterleaver (6614_2) receives the log-likelihood ratio signal (6613_2) as input, reorders the signal, and outputs a deinterleaved log-likelihood
25 ratio signal (6615_2).

[0934]

A decoder (6616_1) receives the deinterleaved log-likelihood ratio signal (6615_1) as input, performs error correction decoding, and outputs received information (6617_1).

[0935]

5 Similarly, a decoder (6616_2) receives the deinterleaved log-likelihood ratio signal (6615_2) as input, performs error correction decoding, and outputs received information (6617_2).

[0936]

10 When a transmission mode exists, as in Table 5, the following methods are possible.

- As described in Embodiment 1, the transmission device transmits information regarding the precoding matrices used in the method of switching between precoding matrices. The detection and log-likelihood ratio calculation units (6612_1 and 6612_2) obtain this information and decode the precoding.
- 15 · As described in Embodiment 7, the transmission and reception devices share the information in Table 5 beforehand, and the transmission device transmits information on the mode. Based on Table 5, the reception device estimates the precoding matrices used in the method of switching between precoding matrices and decodes the precoding.

20 [0937]

As described above, in the case of hierarchical transmission, using the above methods of switching between precoding matrices achieves the effect of improving reception quality of data.

[0938]

25 The present embodiment has described examples of four-slot and eight-slot periods (cycles) in the method of regularly switching between precoding matrices, but the periods (cycles) are not limited in this way. Accordingly, for a precoding hopping method with an N-slot period (cycle), N different precoding matrices are

necessary. In this case, $F[0]$, $F[1]$, $F[2]$, ..., $F[N - 2]$, $F[N - 1]$ are prepared as the N different precoding matrices. In the present embodiment, these have been described as being arranged in the frequency domain in the order of $F[0]$, $F[1]$, $F[2]$, ..., $F[N - 2]$, $F[N - 1]$, but arrangement is not limited in this way. With N different precoding
5 matrices $F[0]$, $F[1]$, $F[2]$, ..., $F[N - 2]$, $F[N - 1]$ generated in the present Embodiment, precoding weights may be changed by arranging symbols in the time domain or in the frequency/time domains as in Embodiment 1. Note that a precoding hopping method with an N -slot period (cycle) has been described, but the same advantageous effects may be obtained by randomly using N different precoding
10 matrices. In other words, the N different precoding matrices do not necessarily need to be used in a regular period (cycle).

[0939]

In Table 5, as an example of when transmission is not hierarchical, it has been described that for some modes, a hierarchical transmission method is not used
15 in the method of regularly switching between precoding matrices, but modes are not limited in this way. As described in Embodiment 15, a spatial multiplexing MIMO system, a MIMO system in which precoding matrices are fixed, a space-time block coding method, and a one-stream-only transmission mode may exist separately from the hierarchical transmission method described in the present embodiment, and the
20 transmission device (broadcast station, base station) may select the transmission method from among these modes. In this case, in the spatial multiplexing MIMO system, the MIMO system in which precoding matrices are fixed, the space-time block coding method, and the one-stream-only transmission mode, both transmission that is hierarchical and transmission that is not hierarchical may be
25 supported. Modes that use other transmission methods may also exist. The present embodiment may also be adapted to Embodiment 15 so that the hierarchical transmission method that uses the method of regularly switching between precoding

matrices, as described in the present Embodiment, is used in any of the (sub)carriers in Embodiment 15.

(Embodiment A2)

In Embodiment A1, a method of achieving hierarchical transmission with
5 methods of regularly switching between precoding matrices has been described. In the present embodiment, a different way of achieving hierarchical transmission is described.

[0940]

Figs. 67 and 68 show the structure of a transmission device when
10 performing the hierarchical transmission of the present embodiment. Constituent elements that are the same as in Figs. 61 and 62 are labeled with the same reference signs. The difference between Fig. 67 and Fig. 61 is that the precoder 6108_1 is not provided. The present embodiment differs from Embodiment A1 in that the base stream (layer) is not precoded.

15 [0941]

In Fig. 67, the mapper (6105_1) receives the interleaved, encoded data (6104_1) and the information signal regarding the transmission method (6111) as input, performs mapping according to a predetermined modulation method based on the information signal regarding the transmission method (6111), and outputs a
20 baseband signal (6700).

[0942]

In Fig. 68, the reordering unit (6200_1) receives the baseband signal (6700), the precoded baseband signal (6110_1), and the information signal regarding the transmission method (6111) as input, performs reordering based on the information
25 signal regarding the transmission method (6111), and outputs the reordered baseband signal (6201_1).

[0943]

The reordering unit (6200_2) receives the precoded baseband signal (6110_2) and the information signal regarding the transmission method (6111) as input, performs reordering based on the information signal regarding the transmission method (6111), and outputs the reordered baseband signal (6201_2).

5 [0944]

Fig. 69 shows an example of symbol structure in the baseband signal of Fig. 67. The symbol group is labeled 6901. In the symbol group (6901), symbols are represented as “B #X”, which refers to the “Xth symbol in the base stream (base layer)”. Note that the structure of symbols in the enhancement stream (enhancement
10 layer) is as shown in Fig. 64.

[0945]

Figs. 70A and 70B show the method of reordering in the reordering unit (6200_1) and the reordering unit (6200_2) in Fig. 68. Symbols shown in Figs. 64 and 69 are arranged in the frequency and time domain as shown in Figs. 70A and 70B. In Figs. 70A and 70B, a “-” indicates that no symbol exists. During
15 transmission, symbols in the same (sub)carrier and at the same time are transmitted at the same frequency and at the same time from different antennas. Note that the arrangement of symbols in the frequency and the time domains as shown in Figs. 70A and 70B is only an example. Symbols may be arranged based on the method
20 described in Embodiment 1.

[0946]

When the base stream (base layer) and the enhancement stream (enhancement layer) are transmitted, it is necessary for the reception quality of data in the base stream (base layer) to be made higher than the reception quality of data
25 in the enhancement stream (enhancement layer), due to the nature of the streams (layers). Therefore, as in the present embodiment, when transmitting the base stream, the reception quality of data is guaranteed by transmitting using only the modulated signal z_1 (i.e. without transmitting the modulated signal z_2). Conversely, when

transmitting the enhancement stream, hierarchical transmission is implemented by using a method of regularly switching between precoding matrices, since improvement of transmission speed is prioritized. For example, it is possible to use one of modes #1–#9 as in Table 6.

5 [0947]

Table 6

Mode	Modulation method for base stream (layer)	Modulation method for enhancement stream (layer)
Mode #1	QPSK	16QAM
Mode #2	QPSK	64QAM
Mode #3	QPSK	256QAM
Mode #4	16QAM	16QAM
Mode #5	16QAM	64QAM
Mode #6	16QAM	256QAM
Mode #7	64QAM	64QAM
Mode #8	64QAM	256QAM
Mode #9	256QAM	256QAM

[0948]

The characteristic feature of Table 6 is that the modulation method for the
 10 base stream (base layer) and the modulation method for the enhancement stream
 (enhancement layer) may be set the same. This is because even if the modulation
 method is the same, the transmission quality that can be guaranteed for the base
 stream (base layer) and the transmission quality that can be guaranteed for the
 enhancement stream (enhancement layer) differ, since different transmission
 15 methods are used for the two streams (layers).

[0949]

The structure of a transmission device according to the present embodiment is shown in Figs. 7 and 66. The difference from the operations in Embodiment A1 is that the detection and log-likelihood ratio calculation unit (6612_1) in Fig. 66 does not decode precoding.

5 [0950]

In the enhancement stream (enhancement layer), a method of regularly switching between precoding matrices is used. As long as information regarding the precoding method used by the transmission device is transmitted, the reception device can identify the precoding method used by acquiring this information. If the
10 transmission and reception devices share the information in Table 6, another method is for the reception device to identify the precoding method used for the enhancement stream (enhancement layer) by acquiring mode information transmitted by the transmission device. Accordingly, the reception device in Fig. 66 can acquire the log-likelihood ratio for each bit by having the detection and
15 log-likelihood ratio calculation unit change the signal processing method. Note that settable modes have been described with reference to Table 6, but modes are not limited in this way. The present embodiment may be similarly achieved using the modes for transmission methods described in Embodiment 8 or modes for transmission methods described in subsequent embodiments.

20 [0951]

As described above, in the case of hierarchical transmission, using the above methods of switching between precoding matrices achieves the effect of improving reception quality of data in the reception device.

[0952]

25 The periods (cycles) of switching between precoding matrices in the method of regularly switching between precoding matrices are not limited as above in the present embodiment. For a precoding hopping method with an N-slot period (cycle), N different precoding matrices are necessary. In this case, $F[0]$, $F[1]$, $F[2]$, ..., $F[N -$

2] , $F[N - 1]$ are prepared as the N different precoding matrices. In the present embodiment, these have been described as being arranged in the frequency domain in the order of $F[0]$, $F[1]$, $F[2]$, ..., $F[N - 2]$, $F[N - 1]$, but arrangement is not limited in this way. With N different precoding matrices $F[0]$, $F[1]$, $F[2]$, ..., $F[N - 2]$, $F[N - 1]$ generated in the present Embodiment, precoding weights may be changed by arranging symbols in the time domain or in the frequency/time domains as in Embodiment 1. Note that a precoding hopping method with an N -slot period (cycle) has been described, but the same advantageous effects may be obtained by randomly using N different precoding matrices. In other words, the N different precoding matrices do not necessarily need to be used in a regular period (cycle).

[0953]

Furthermore, Table 6 has been described as listing modes for methods of hierarchical transmission in the present embodiment, but modes are not limited in this way. As described in Embodiment 15, a spatial multiplexing MIMO system, a MIMO system in which precoding matrices are fixed, a space-time block coding method, a one-stream-only transmission mode, and modes for methods of regularly switching between precoding matrices may exist separately from the hierarchical transmission method described in the present embodiment, and the transmission device (broadcast station, base station) may select the transmission method from among these modes. In this case, in the spatial multiplexing MIMO system, the MIMO system in which precoding matrices are fixed, the space-time block coding method, the one-stream-only transmission mode, and the modes for methods of regularly switching between precoding matrices, both transmission that is hierarchical and transmission that is not hierarchical may be supported. Modes that use other transmission methods may also exist. The present embodiment may also be adapted to Embodiment 15 so that the hierarchical transmission method described in the present Embodiment is used in any of the (sub)carriers in Embodiment 15.

[0954]

(Embodiment A3)

The present embodiment describes hierarchical transmission that differs from Embodiments A1 and A2.

[0955]

5 Figs. 71 and 72 show the structure of a transmission device when performing the hierarchical transmission of the present embodiment. Constituent elements that are the same as in Figs. 61 and 62 are labeled with the same reference signs. The difference between Figs. 71 and 61 is that a space-time block coder 7101 is provided. The present embodiment differs from Embodiment A2 in that
10 space-time block coding is performed on the base stream (layer).

[0956]

The space-time block coder (7101) (which in some cases may be a frequency-space block coder) in Fig. 71 receives a mapped baseband signal (7100) and the information signal regarding the transmission method (6111) as input,
15 performs space-time block coding based on the information signal regarding the transmission method (6111), and outputs a space-time block coded baseband signal (7102_1) (represented as $z_1(t)$) and a space-time block coded baseband signal (7102_2) (represented as $z_2(t)$).

[0957]

20 While referred to here as space-time block coding, symbols that are space-time block coded are not limited to being arranged in order in the time domain. Space-time block coded symbols may be arranged in order in the frequency domain. Furthermore, blocks may be formed with a plurality of symbols in the time domain and a plurality of symbols in the frequency domain, and the blocks may be arranged
25 appropriately (i.e. arranged using both the time and the frequency axes).

[0958]

In Fig. 72, the reordering unit (6200_1) receives the space-time block coded baseband signal (7102_1), the precoded baseband signal (6110_1), and the

information signal regarding the transmission method (6111) as input, performs reordering based on the information signal regarding the transmission method (6111), and outputs the reordered baseband signal (6201_1).

[0959]

5 Similarly, the reordering unit (6200_2) receives the precoded baseband signal (7102_2), the precoded baseband signal (6110_2), and the information signal regarding the transmission method (6111) as input, performs reordering based on the information signal regarding the transmission method (6111), and outputs the reordered baseband signal (6201_2).

10 [0960]

Fig. 73 is an example of a structure of symbols in space-time block coded baseband signals (7102_1, 7102_2) output by the space-time block coder (7101) in Fig. 71. The symbol group (7301) corresponds to the space-time block coded baseband signal (7102_1) (represented as $z_1(t)$), and the symbol group (7302) corresponds to the space-time block coded baseband signal (7102_2) (represented as $z_2(t)$).

[0961]

The mapper (6105_1) in Fig. 71 represents signals as $s_1, s_2, s_3, s_4, s_5, s_6, s_7, s_8, s_9, s_{10}, s_{11}, s_{12}, \dots$ in the order in which signals are output. The space-time block coder (7101) in Fig. 71 then performs space-time block coding on s_1 and s_2 , yielding $s_1, s_2, s_1^*,$ and $-s_2^*$ (*: complex conjugate), which are output as in Fig. 73. Similarly, space-time block coding is performed on the sets $(s_3, s_4), (s_5, s_6), (s_7, s_8), (s_9, s_{10}), (s_{11}, s_{12}), \dots$, and symbols are arranged as in Fig. 73. Note that space-time block coding is not limited to the coding described in the present embodiment; the present embodiment may be similarly achieved using different space-time block coding.

[0962]

Figs. 74A and 74B show an example of the method of reordering in the reordering unit (6200_1) and the reordering unit (6200_2) in Fig. 72. Fig. 74A is an example of arranging symbols in the modulated signal z_1 in the time domain and the frequency domain. Fig. 74B is an example of arranging symbols in the modulated signal z_2 in the time domain and the frequency domain. During transmission, symbols in the same (sub)carrier and at the same time are transmitted at the same frequency and at the same time from different antennas. The characteristic feature of Figs. 74A and 74B is that space-time block coded symbols are arranged in the frequency domain in order.

10 [0963]

Figs. 75A and 75B show an example of the method of reordering in the reordering unit (6200_1) and the reordering unit (6200_2) in Fig. 72. Fig. 75A is an example of arranging symbols in the modulated signal z_1 in the time domain and the frequency domain. Fig. 75B is an example of arranging symbols in the modulated signal z_2 in the time domain and the frequency domain. During transmission, symbols in the same (sub)carrier and at the same time are transmitted at the same frequency and at the same time from different antennas. The characteristic feature of Figs. 75A and 75B is that space-time block coded symbols are arranged in the time domain in order.

20 [0964]

Space-time block coded symbols can thus be ordered in the frequency domain or in the time domain.

[0965]

When the base stream (base layer) and the enhancement stream (enhancement layer) are transmitted, it is necessary for the reception quality of data in the base stream (base layer) to be made higher than the reception quality of data in the enhancement stream (enhancement layer), due to the nature of the streams (layers). Therefore, as in the present embodiment, when transmitting the base stream,

the reception quality of data is guaranteed by using space-time block coding to achieve diversity gain. Conversely, when transmitting the enhancement stream, hierarchical transmission is implemented by using a method of regularly switching between precoding matrices, since improvement of transmission speed is prioritized.

- 5 For example, it is possible to use one of modes #1–#9 as in Table 7.

[0966]

Table 7

Mode	Modulation method for base stream (layer)	Modulation method for enhancement stream (layer)
Mode #1	QPSK	16QAM
Mode #2	QPSK	64QAM
Mode #3	QPSK	256QAM
Mode #4	16QAM	16QAM
Mode #5	16QAM	64QAM
Mode #6	16QAM	256QAM
Mode #7	64QAM	64QAM
Mode #8	64QAM	256QAM
Mode #9	256QAM	256QAM

[0967]

- 10 The characteristic feature of Table 7 is that the modulation method for the base stream (base layer) and the modulation method for the enhancement stream (enhancement layer) may be set the same. This is because even if the modulation method is the same, the transmission quality that can be guaranteed for the base stream (base layer) and the transmission quality that can be guaranteed for the
- 15 enhancement stream (enhancement layer) differ, since different transmission methods are used for the two streams (layers).

[0968]

Note that modes #1–#9 in Table 7 are modes for hierarchical transmission, but modes that are not for hierarchical transmission may also be supported. In the present embodiment, a single mode for space-time block coding and a single mode
5 for regularly switching between precoding matrices may exist as modes that are not for hierarchical transmission, and when supporting the modes for hierarchical transmission in Table 7, the transmission device and the reception device of the present embodiment may easily set the mode to the single mode for space-time block coding or the single mode for regularly switching between precoding matrices.

10 [0969]

Furthermore, in the enhancement stream (enhancement layer), a method of regularly switching between precoding matrices is used. As long as information regarding the precoding method used by the transmission device is transmitted, the reception device can identify the precoding method used by acquiring this
15 information. If the transmission and reception devices share the information in Table 7, another method is for the reception device to identify the precoding method used for the enhancement stream (enhancement layer) by acquiring mode information transmitted by the transmission device. Accordingly, the reception device in Fig. 66 can acquire the log-likelihood ratio for each bit by having the detection and
20 log-likelihood ratio calculation unit change the signal processing method. Note that settable modes have been described with reference to Table 7, but modes are not limited in this way. The present embodiment may be similarly achieved using the modes for transmission methods described in Embodiment 8 or modes for transmission methods described in subsequent embodiments.

25 [0970]

As described above, in the case of hierarchical transmission, using the above methods of switching between precoding matrices achieves the effect of improving reception quality of data in the reception device.

[0971]

The periods (cycles) of switching between precoding matrices in the method of regularly switching between precoding matrices are not limited as above in the present embodiment. For a precoding hopping method with an N-slot period (cycle),
 5 N different precoding matrices are necessary. In this case, $F[0]$, $F[1]$, $F[2]$, ..., $F[N - 2]$, $F[N - 1]$ are prepared as the N different precoding matrices. In the present embodiment, these have been described as being arranged in the frequency domain in the order of $F[0]$, $F[1]$, $F[2]$, ..., $F[N - 2]$, $F[N - 1]$, but arrangement is not limited in this way. With N different precoding matrices $F[0]$, $F[1]$, $F[2]$, ..., $F[N - 2]$, $F[N - 1]$ generated in the present Embodiment, precoding weights may be changed by
 10 arranging symbols in the time domain or in the frequency/time domains as in Embodiment 1. Note that a precoding hopping method with an N-slot period (cycle) has been described, but the same advantageous effects may be obtained by randomly using N different precoding matrices. In other words, the N different precoding
 15 matrices do not necessarily need to be used in a regular period (cycle).

[0972]

Furthermore, Table 7 has been described as listing modes for methods of hierarchical transmission in the present embodiment, but modes are not limited in this way. As described in Embodiment 15, a spatial multiplexing MIMO system, a
 20 MIMO system in which precoding matrices are fixed, a space-time block coding method, a one-stream-only transmission mode, and modes for methods of regularly switching between precoding matrices may exist separately from the hierarchical transmission method described in the present embodiment, and the transmission device (broadcast station, base station) may select the transmission method from
 25 among these modes. In this case, in the spatial multiplexing MIMO system, the MIMO system in which precoding matrices are fixed, the space-time block coding method, the one-stream-only transmission mode, and the modes for methods of regularly switching between precoding matrices, both transmission that is

hierarchical and transmission that is not hierarchical may be supported. Modes that use other transmission methods may also exist. The present embodiment may also be adapted to Embodiment 15 so that the hierarchical transmission method described in the present Embodiment is used in any of the (sub)carriers in Embodiment 15.

5 [0973]

(Embodiment A4)

The present embodiment describes, in detail, a method of regularly switching between precoding matrices when using block coding as shown in Non-Patent Literature 12 through Non-Patent Literature 15, such as a Quasi-Cyclic
 10 Low-Density Parity-Check (QC-LDPC) code (or an LDPC code other than a QC-LDPC code), a concatenated code consisting of an LDPC code and a Bose-Chaudhuri-Hocquenghem (BCH) code, or the like. This embodiment describes an example of transmitting two streams, s1 and s2. However, for the case of coding using block codes, when control information and the like is not necessary, the
 15 number of bits in an encoded block matches the number of bits composing the block code (the control information or the like listed below may, however, be included therein). For the case of coding using block codes, when control information or the like (such as a cyclic redundancy check (CRC), transmission parameters, or the like) is necessary, the number of bits in an encoded block is the sum of the number of bits
 20 composing the block code and the number of bits in the control information or the like.

[0974]

Fig. 76 shows a modification of the number of symbols and of slots necessary for one encoded block when using block coding. Fig. 76 “shows a
 25 modification of the number of symbols and of slots necessary for one encoded block when using block coding” for the case when, for example as shown in the transmission device in Fig. 4, two streams, s1 and s2, are transmitted, and the transmission device has one encoder. (In this case, the transmission method may be

either single carrier transmission, or multicarrier transmission such as OFDM.) As shown in Fig. 76, the number of bits constituting one block that has been encoded via block coding is set to 6,000. In order to transmit these 6,000 bits, 3,000 symbols are required when the modulation method is QPSK, 1,500 when the modulation
 5 method is 16QAM, and 1,000 when the modulation method is 64QAM.

[0975]

Since the transmission device in Fig. 4 simultaneously transmits two streams, 1,500 of the 3,000 symbols when the modulation method is QPSK are allocated to s1, and 1,500 to s2. Therefore, 1,500 slots (the term “slot” is used here)
 10 are required to transmit the 1,500 symbols transmitted in s1 and the 1,500 symbols transmitted in s2.

[0976]

By similar reasoning, when the modulation method is 16QAM, 750 slots are necessary to transmit all of the bits constituting one encoded block, and when the
 15 modulation method is 64QAM, 500 slots are necessary to transmit all of the bits constituting one block.

[0977]

The following describes the relationship between the slots defined above and the precoding matrices in the method of regularly switching between precoding
 20 matrices.

[0978]

Here, the number of precoding matrices prepared for the method of regularly switching between precoding matrices is set to five. In other words, five different precoding matrices are prepared for the weighting unit in the transmission
 25 device in Fig. 4. These five different precoding matrices are represented as $F[0]$, $F[1]$, $F[2]$, $F[3]$, and $F[4]$.

[0979]

When the modulation method is QPSK, among the 1,500 slots described above for transmitting the 6,000 bits constituting one encoded block, it is necessary for 300 slots to use the precoding matrix $F[0]$, 300 slots to use the precoding matrix $F[1]$, 300 slots to use the precoding matrix $F[2]$, 300 slots to use the precoding matrix $F[3]$, and 300 slots to use the precoding matrix $F[4]$. This is because if use of the precoding matrices is biased, the reception quality of data is greatly influenced by the precoding matrix that was used a greater number of times.

[0980]

When the modulation method is 16QAM, among the 750 slots described above for transmitting the 6,000 bits constituting one encoded block, it is necessary for 150 slots to use the precoding matrix $F[0]$, 150 slots to use the precoding matrix $F[1]$, 150 slots to use the precoding matrix $F[2]$, 150 slots to use the precoding matrix $F[3]$, and 150 slots to use the precoding matrix $F[4]$.

[0981]

When the modulation method is 64QAM, among the 500 slots described above for transmitting the 6,000 bits constituting one encoded block, it is necessary for 100 slots to use the precoding matrix $F[0]$, 100 slots to use the precoding matrix $F[1]$, 100 slots to use the precoding matrix $F[2]$, 100 slots to use the precoding matrix $F[3]$, and 100 slots to use the precoding matrix $F[4]$.

[0982]

As described above, in the method of regularly switching between precoding matrices, if there are N different precoding matrices (represented as $F[0]$, $F[1]$, $F[2]$, ..., $F[N - 2]$, and $F[N - 1]$), when transmitting all of the bits constituting one encoded block, condition #53 should be satisfied, wherein K_0 is the number of slots using the precoding matrix $F[0]$, K_1 is the number of slots using the precoding matrix $F[1]$, K_i is the number of slots using the precoding matrix $F[i]$ ($i = 0, 1, 2, \dots, N - 1$), and K_{N-1} is the number of slots using the precoding matrix $F[N - 1]$.

Condition #53

$K_0 = K_1 = \dots = K_i = \dots = K_{N-1}$, i.e. $K_a = K_b$ (for $\forall a, \forall b$, where $a, b, = 0, 1, 2, \dots, N - 1$, and $a \neq b$).

[0983]

- 5 If the communications system supports a plurality of modulation methods, and the modulation method that is used is selected from among the supported modulation methods, then a modulation method for which Condition #53 is satisfied should be selected.

[0984]

- 10 When a plurality of modulation methods are supported, it is typical for the number of bits that can be transmitted in one symbol to vary from modulation method to modulation method (although it is also possible for the number of bits to be the same), and therefore some modulation methods may not be capable of satisfying Condition #53. In such a case, instead of Condition #53, the following
15 condition should be satisfied.

[0985]

Condition #54

The difference between K_a and K_b is 0 or 1, i.e. $|K_a - K_b|$ is 0 or 1 (for $\forall a, \forall b$, where $a, b, = 0, 1, 2, \dots, N - 1$, and $a \neq b$).

20

- Fig. 77 shows a modification of the number of symbols and of slots necessary for one encoded block when using block coding. Fig. 77 “shows a modification of the number of symbols and of slots necessary for one encoded block when using block coding” for the case when, for example as shown in the
25 transmission device in Fig. 3 and in Fig. 13, two streams are transmitted, i.e. s_1 and s_2 , and the transmission device has two encoders. (In this case, the transmission method may be either single carrier transmission, or multicarrier transmission such as OFDM.)

As shown in Fig. 77, the number of bits constituting one block that has been encoded via block coding is set to 6,000. In order to transmit these 6,000 bits, 3,000 symbols are required when the modulation method is QPSK, 1,500 when the modulation method is 16QAM, and 1,000 when the modulation method is 64QAM.

5 [0986]

The transmission device in Fig. 3 or in Fig. 13 transmits two streams simultaneously, and since two encoders are provided, different encoded blocks are transmitted in the two streams. Accordingly, when the modulation method is QPSK, two encoded blocks are transmitted in s1 and s2 within the same interval. For example, a first encoded block is transmitted in s1, and a second encoded block is transmitted in s2, and therefore, 3,000 slots are required to transmit the first and second encoded blocks.

[0987]

By similar reasoning, when the modulation method is 16QAM, 1,500 slots are necessary to transmit all of the bits constituting two encoded blocks, and when the modulation method is 64QAM, 1,000 slots are necessary to transmit all of the bits constituting two blocks.

[0988]

The following describes the relationship between the slots defined above and the precoding matrices in the method of regularly switching between precoding matrices. Here, the number of precoding matrices prepared for the method of regularly switching between precoding matrices is set to five. In other words, five different precoding matrices are prepared for the weighting unit in the transmission device in Fig. 3 or in Fig. 13. These five different precoding matrices are represented as F[0], F[1], F[2], F[3], and F[4].

[0989]

When the modulation method is QPSK, among the 3,000 slots described above for transmitting the $6,000 \times 2$ bits constituting two encoded blocks, it is

necessary for 600 slots to use the precoding matrix $F[0]$, 600 slots to use the precoding matrix $F[1]$, 600 slots to use the precoding matrix $F[2]$, 600 slots to use the precoding matrix $F[3]$, and 600 slots to use the precoding matrix $F[4]$. This is because if use of the precoding matrices is biased, the reception quality of data is greatly influenced by the precoding matrix that was used a greater number of times.

[0990]

To transmit the first encoded block, it is necessary for the slot using the precoding matrix $F[0]$ to occur 600 times, the slot using the precoding matrix $F[1]$ to occur 600 times, the slot using the precoding matrix $F[2]$ to occur 600 times, the slot using the precoding matrix $F[3]$ to occur 600 times, and the slot using the precoding matrix $F[4]$ to occur 600 times. To transmit the second encoded block, the slot using the precoding matrix $F[0]$ should occur 600 times, the slot using the precoding matrix $F[1]$ should occur 600 times, the slot using the precoding matrix $F[2]$ should occur 600 times, the slot using the precoding matrix $F[3]$ should occur 600 times, and the slot using the precoding matrix $F[4]$ should occur 600 times.

[0991]

Similarly, when the modulation method is 16QAM, among the 1,500 slots described above for transmitting the $6,000 \times 2$ bits constituting two encoded blocks, it is necessary for 300 slots to use the precoding matrix $F[0]$, 300 slots to use the precoding matrix $F[1]$, 300 slots to use the precoding matrix $F[2]$, 300 slots to use the precoding matrix $F[3]$, and 300 slots to use the precoding matrix $F[4]$.

[0992]

To transmit the first encoded block, it is necessary for the slot using the precoding matrix $F[0]$ to occur 300 times, the slot using the precoding matrix $F[1]$ to occur 300 times, the slot using the precoding matrix $F[2]$ to occur 300 times, the slot using the precoding matrix $F[3]$ to occur 300 times, and the slot using the precoding matrix $F[4]$ to occur 300 times. To transmit the second encoded block, the slot using the precoding matrix $F[0]$ should occur 300 times, the slot using the precoding

matrix $F[1]$ should occur 300 times, the slot using the precoding matrix $F[2]$ should occur 300 times, the slot using the precoding matrix $F[3]$ should occur 300 times, and the slot using the precoding matrix $F[4]$ should occur 300 times.

[0993]

5 Similarly, when the modulation method is 64QAM, among the 1,000 slots described above for transmitting the $6,000 \times 2$ bits constituting two encoded blocks, it is necessary for 200 slots to use the precoding matrix $F[0]$, 200 slots to use the precoding matrix $F[1]$, 200 slots to use the precoding matrix $F[2]$, 200 slots to use the precoding matrix $F[3]$, and 200 slots to use the precoding matrix $F[4]$.

10 [0994]

To transmit the first encoded block, it is necessary for the slot using the precoding matrix $F[0]$ to occur 200 times, the slot using the precoding matrix $F[1]$ to occur 200 times, the slot using the precoding matrix $F[2]$ to occur 200 times, the slot using the precoding matrix $F[3]$ to occur 200 times, and the slot using the precoding
 15 matrix $F[4]$ to occur 200 times. To transmit the second encoded block, the slot using the precoding matrix $F[0]$ should occur 200 times, the slot using the precoding matrix $F[1]$ should occur 200 times, the slot using the precoding matrix $F[2]$ should occur 200 times, the slot using the precoding matrix $F[3]$ should occur 200 times, and the slot using the precoding matrix $F[4]$ should occur 200 times.

20 [0995]

As described above, in the method of regularly switching between precoding matrices, if there are N different precoding matrices (represented as $F[0]$, $F[1]$, $F[2]$, ..., $F[N - 2]$, and $F[N - 1]$), when transmitting all of the bits constituting two encoded blocks, Condition #55 should be satisfied, wherein K_0 is the number of
 25 slots using the precoding matrix $F[0]$, K_1 is the number of slots using the precoding matrix $F[1]$, K_i is the number of slots using the precoding matrix $F[i]$ ($i = 0, 1, 2, \dots, N - 1$), and K_{N-1} is the number of slots using the precoding matrix $F[N - 1]$.

Condition #55

$K_0 = K_1 = \dots = K_i = \dots = K_{N-1}$, i.e. $K_a = K_b$ (for $\forall a, \forall b$, where $a, b = 0, 1, 2, \dots, N - 1$, and $a \neq b$).

When transmitting all of the bits constituting the first encoded block, Condition #56
 5 should be satisfied, wherein $K_{0,1}$ is the number of times the precoding matrix $F[0]$ is used, $K_{1,1}$ is the number of times the precoding matrix $F[1]$ is used, $K_{i,1}$ is the number of times the precoding matrix $F[i]$ is used ($i = 0, 1, 2, \dots, N - 1$), and $K_{N-1,1}$ is the number of times the precoding matrix $F[N - 1]$ is used.

Condition #56

10 $K_{0,1} = K_{1,1} = \dots = K_{i,1} = \dots = K_{N-1,1}$, i.e. $K_{a,1} = K_{b,1}$ (for $\forall a, \forall b$, where $a, b = 0, 1, 2, \dots, N - 1$, and $a \neq b$).

When transmitting all of the bits constituting the second encoded block, Condition #57 should be satisfied, wherein $K_{0,2}$ is the number of times the precoding matrix
 15 $F[0]$ is used, $K_{1,2}$ is the number of times the precoding matrix $F[1]$ is used, $K_{i,2}$ is the number of times the precoding matrix $F[i]$ is used ($i = 0, 1, 2, \dots, N - 1$), and $K_{N-1,2}$ is the number of times the precoding matrix $F[N - 1]$ is used.

Condition #57

20 $K_{0,2} = K_{1,2} = \dots = K_{i,2} = \dots = K_{N-1,2}$, i.e. $K_{a,2} = K_{b,2}$ (for $\forall a, \forall b$, where $a, b = 0, 1, 2, \dots, N - 1$, and $a \neq b$).

[0996]

If the communications system supports a plurality of modulation methods, and the modulation method that is used is selected from among the supported
 25 modulation methods, and the selected modulation method preferably satisfies Conditions #55, #56, and #57.

[0997]

DEMANDES OU BREVETS VOLUMINEUX

**LA PRÉSENTE PARTIE DE CETTE DEMANDE OU CE BREVETS
COMPREND PLUS D'UN TOME.**

CECI EST LE TOME __1__ DE __2__

NOTE: Pour les tomes additionels, veuillez contacter le Bureau Canadien des Brevets.

JUMBO APPLICATIONS / PATENTS

**THIS SECTION OF THE APPLICATION / PATENT CONTAINS MORE
THAN ONE VOLUME.**

THIS IS VOLUME __1__ OF __2__

NOTE: For additional volumes please contact the Canadian Patent Office.

The embodiments of the invention in which an exclusive property or privilege is claimed are defined as follows:

1. A precoding method for generating, from a plurality of baseband signals, a plurality of precoded signals to be transmitted over the same frequency bandwidth at the same time, comprising the steps of:

selecting a matrix $F[i]$ from among N matrices while switching between the N matrices, the N matrices defining precoding performed on the plurality of baseband signals, i being an integer from 0 to $N - 1$, and N being an integer at least two; and

generating a first precoded signal $z1$ and a second precoded signal $z2$ by precoding, in accordance with the selected matrix $F[i]$, a first baseband signal $s1$ generated from a first plurality of bits and a second baseband signal $s2$ generated from a second plurality of bits, a first encoded block and a second encoded block being generated respectively as the first plurality of bits and the second plurality of bits using a predetermined error correction block encoding method, the first baseband signal $s1$ and the second baseband signal $s2$ being generated respectively from the first encoded block and the second encoded block to have M symbols each, the first precoded signal $z1$ and the second precoded signal $z2$ being generated to have M slots each by precoding a combination of the first baseband signal $s1$ and the second baseband signal $s2$, M being an integer at least two,

the first precoded signal $z1$ and the second precoded signal $z2$ satisfying the equation $(z1, z2)^T = F[i](s1, s2)^T$,

$(z1, z2)^T$ being a transposed matrix of $(z1, z2)$, and

$(s1, s2)^T$ being a transposed matrix of $(s1, s2)$.

2. A precoding apparatus for generating, from a plurality of baseband signals, a plurality of precoded signals to be transmitted over the same frequency bandwidth at the same time, comprising:

a weighting information generation unit configured to select a matrix $F[i]$ from among N matrices while switching between the N matrices, the N matrices defining precoding performed on the plurality of baseband signals, i being an integer from 0 to $N - 1$, and N being an integer at least two;

5 a weighting unit configured to generate a first precoded signal $z1$ and a second precoded signal $z2$ by precoding, in accordance with the selected matrix $F[i]$, a first baseband signal $s1$ generated from a first plurality of bits and a second baseband signal $s2$ generated from a second plurality of bits;

10 an error correction coding unit configured to generate a first encoded block as the first plurality of bits and a second encoded block as the second plurality of bits using a predetermined error correction block encoding method; and

a mapper configured to generate a baseband signal with M symbols from the first encoded block and a baseband signal with M symbols from the second encoded block, M being an integer at least two,

15 the first precoded signal $z1$ and the second precoded signal $z2$ satisfying the equation $(z1, z2)^T = F[i](s1, s2)^T$,

$(z1, z2)^T$ being a transposed matrix of $(z1, z2)$,

$(s1, s2)^T$ being a transposed matrix of $(s1, s2)$, and

20 the weighting unit generating precoded signals with M slots by precoding a combination of the baseband signal generated from the first encoded block and the baseband signal generated from the second encoded block.

FIG. 1

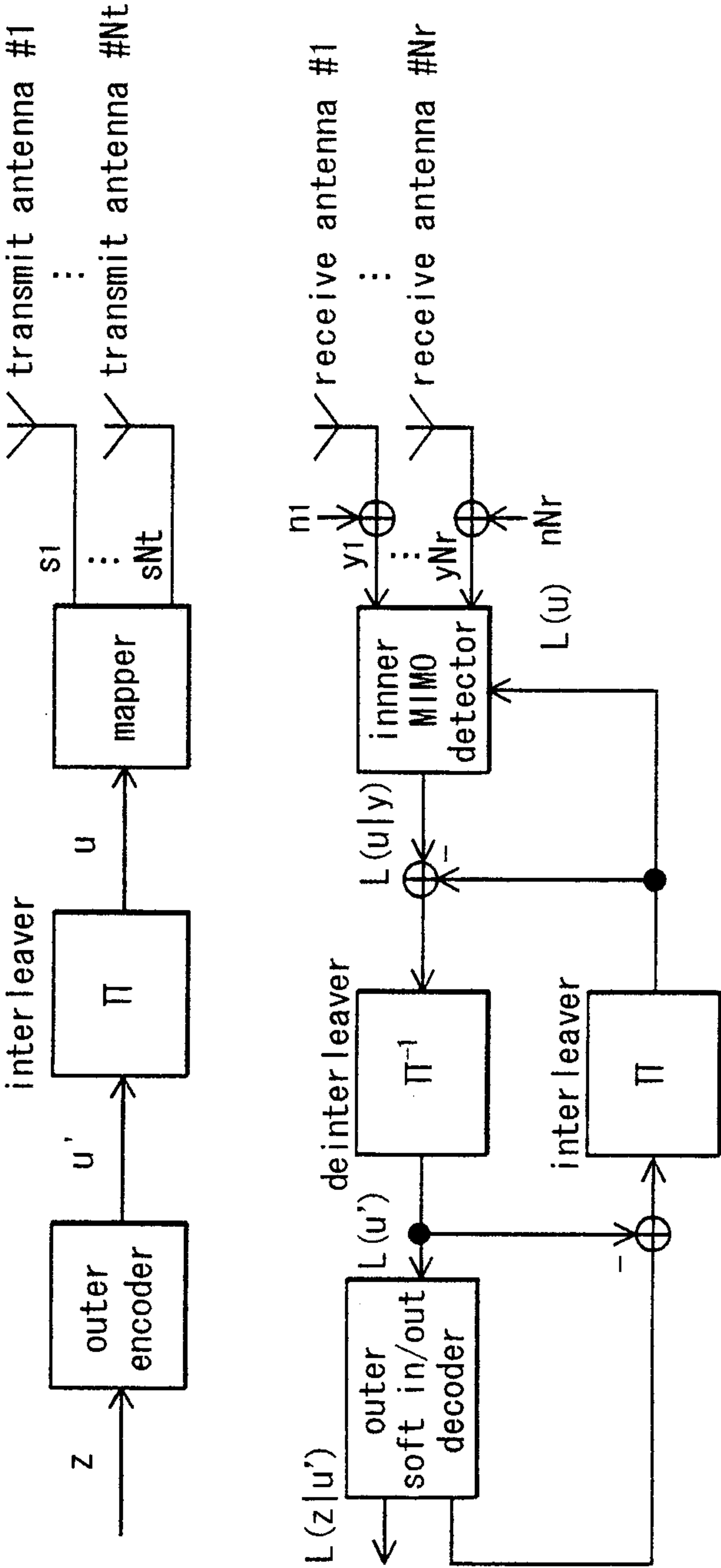
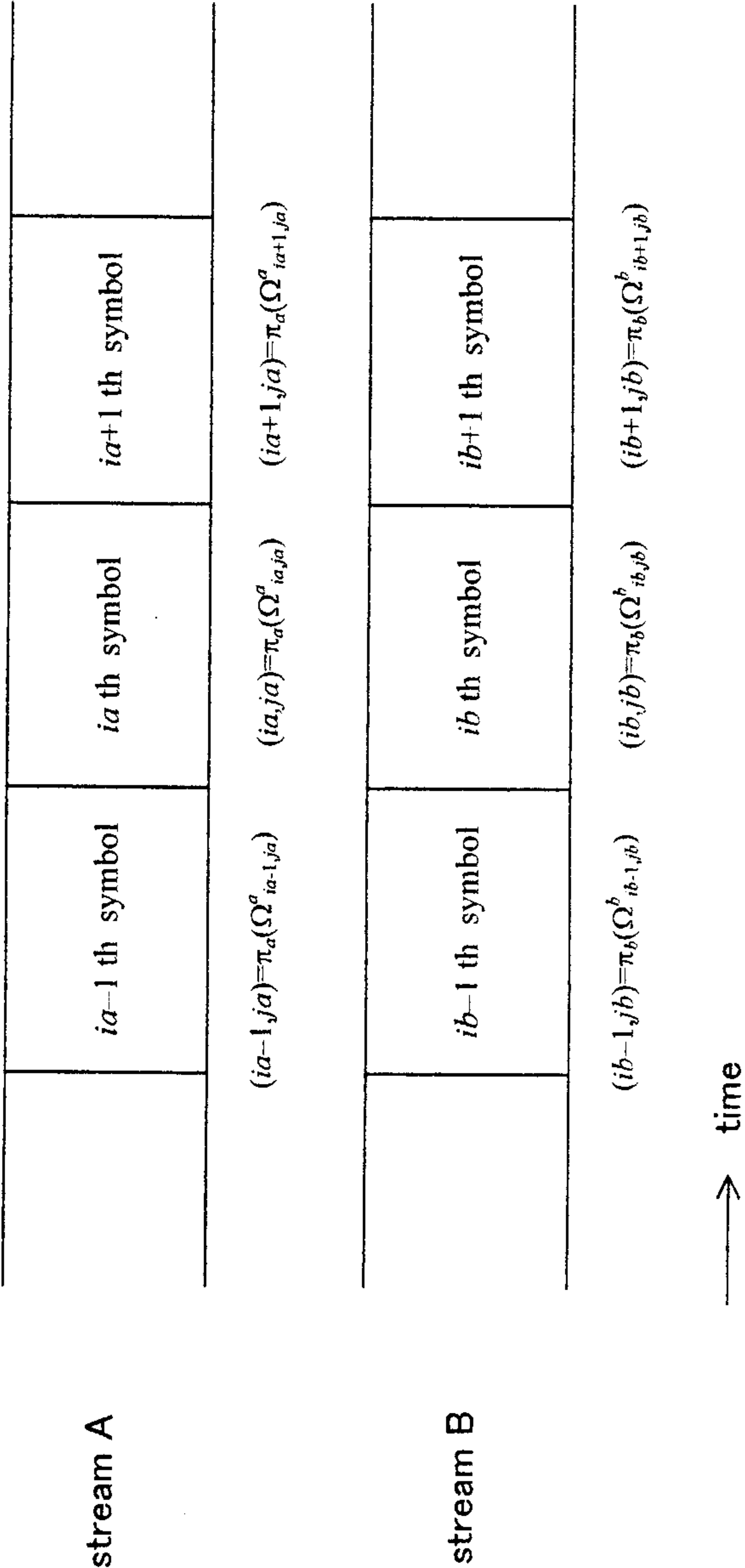
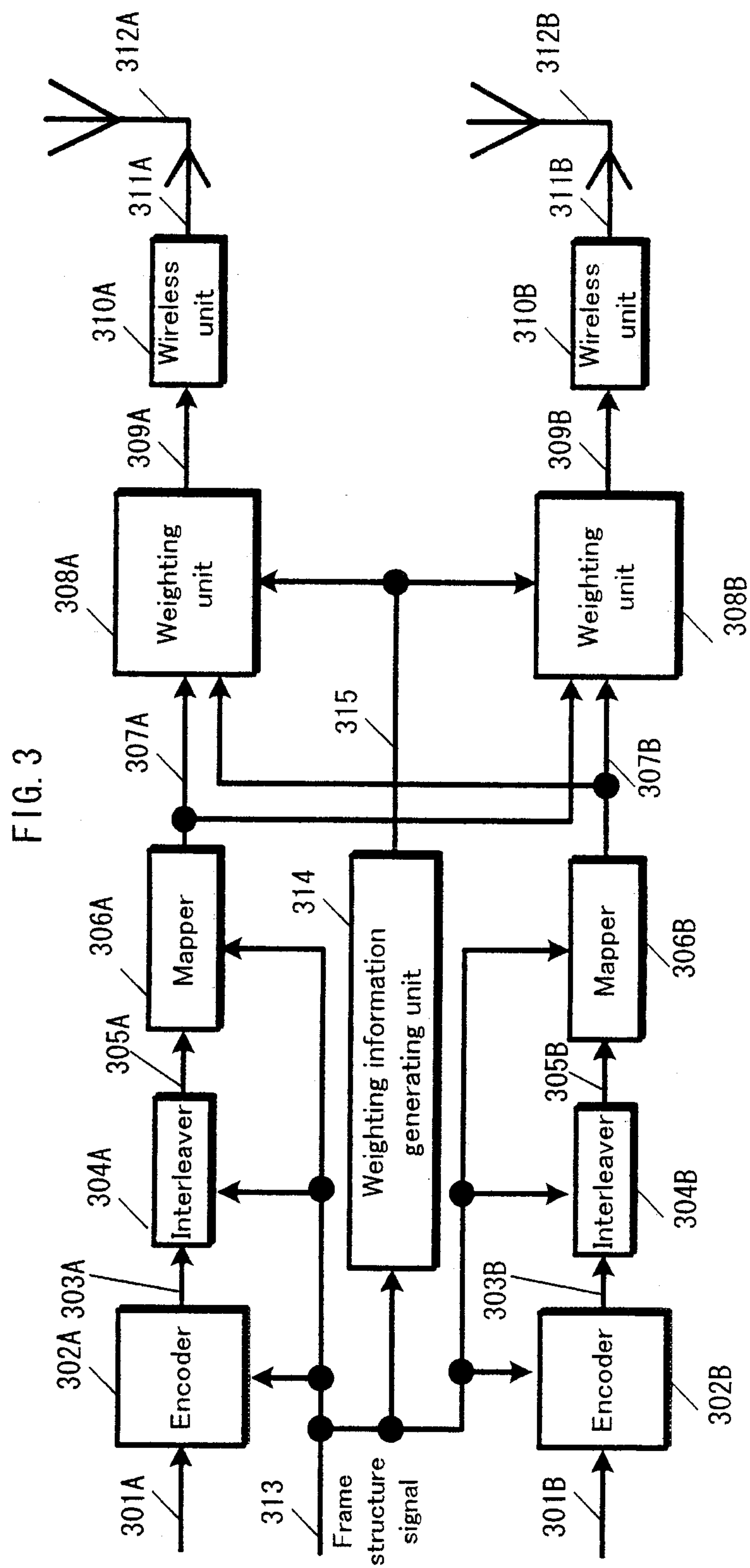


FIG. 2



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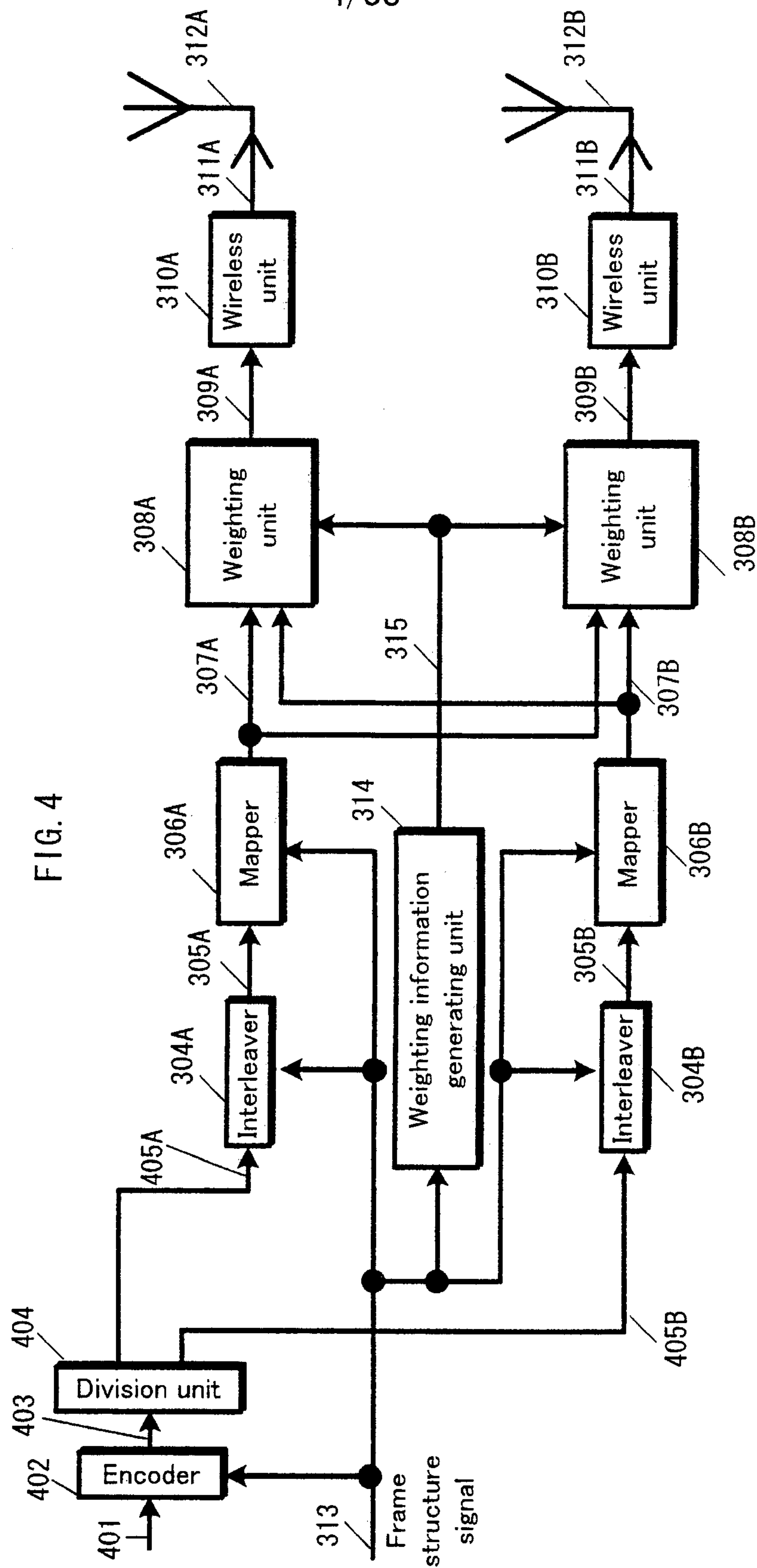


FIG. 5

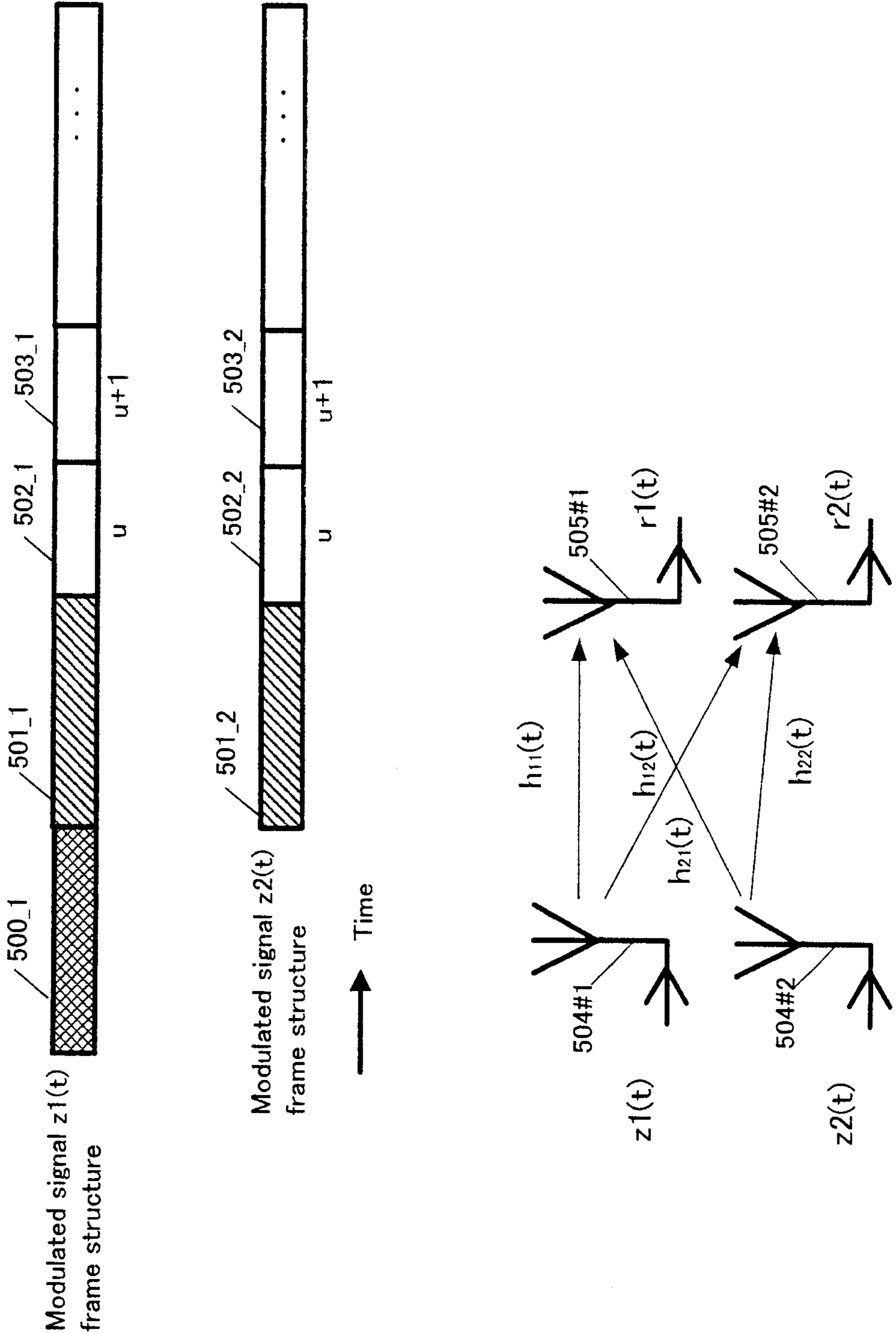
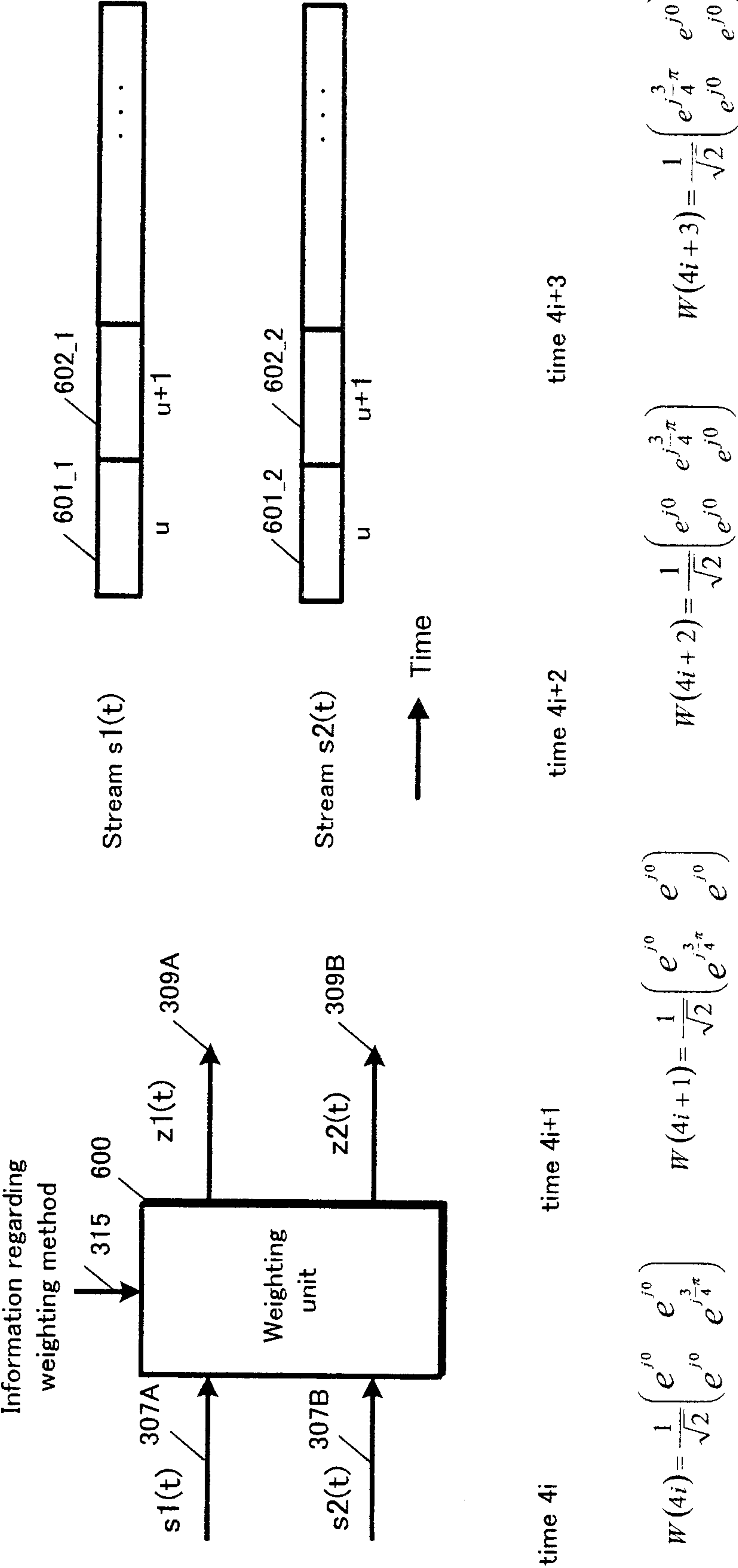
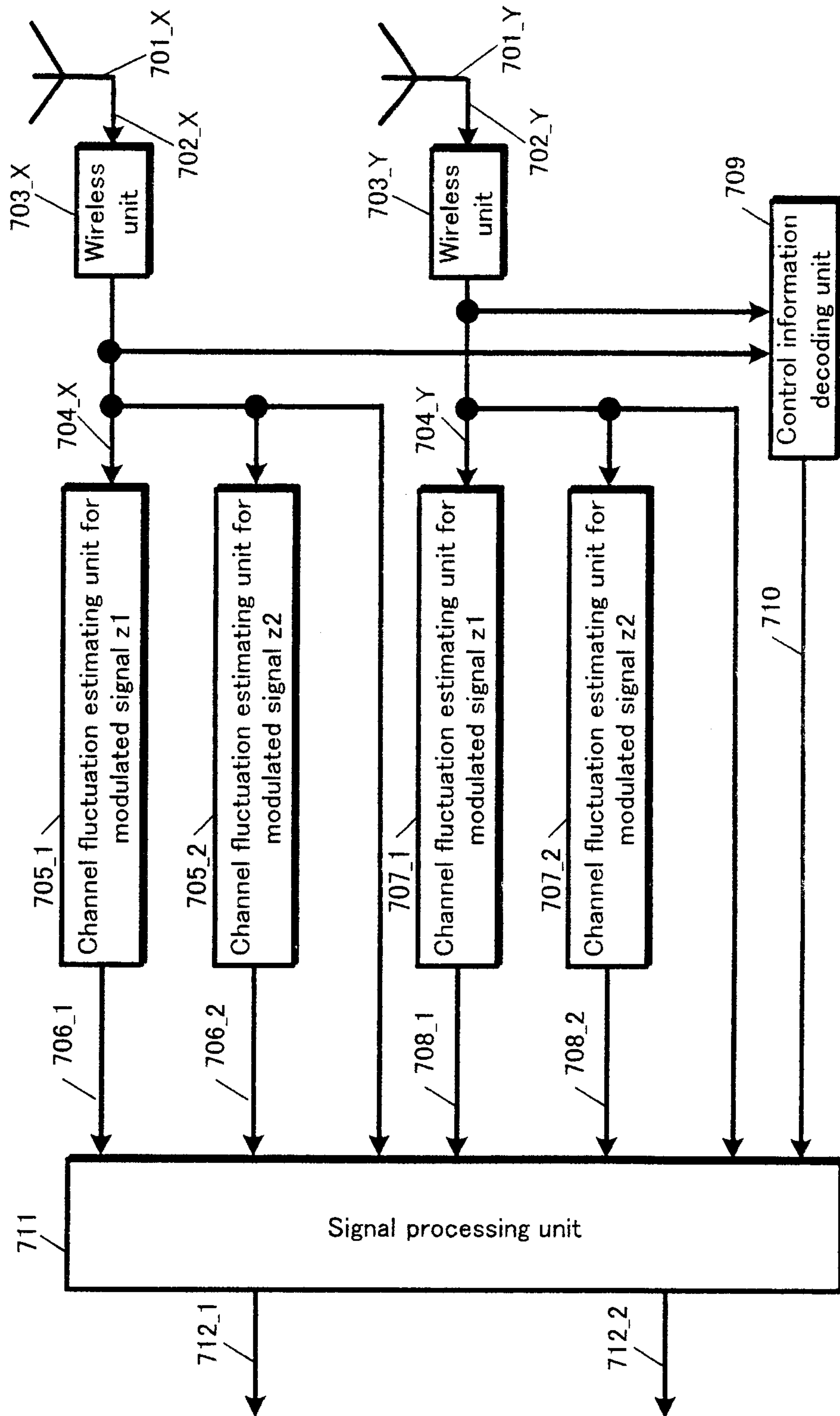


FIG. 6

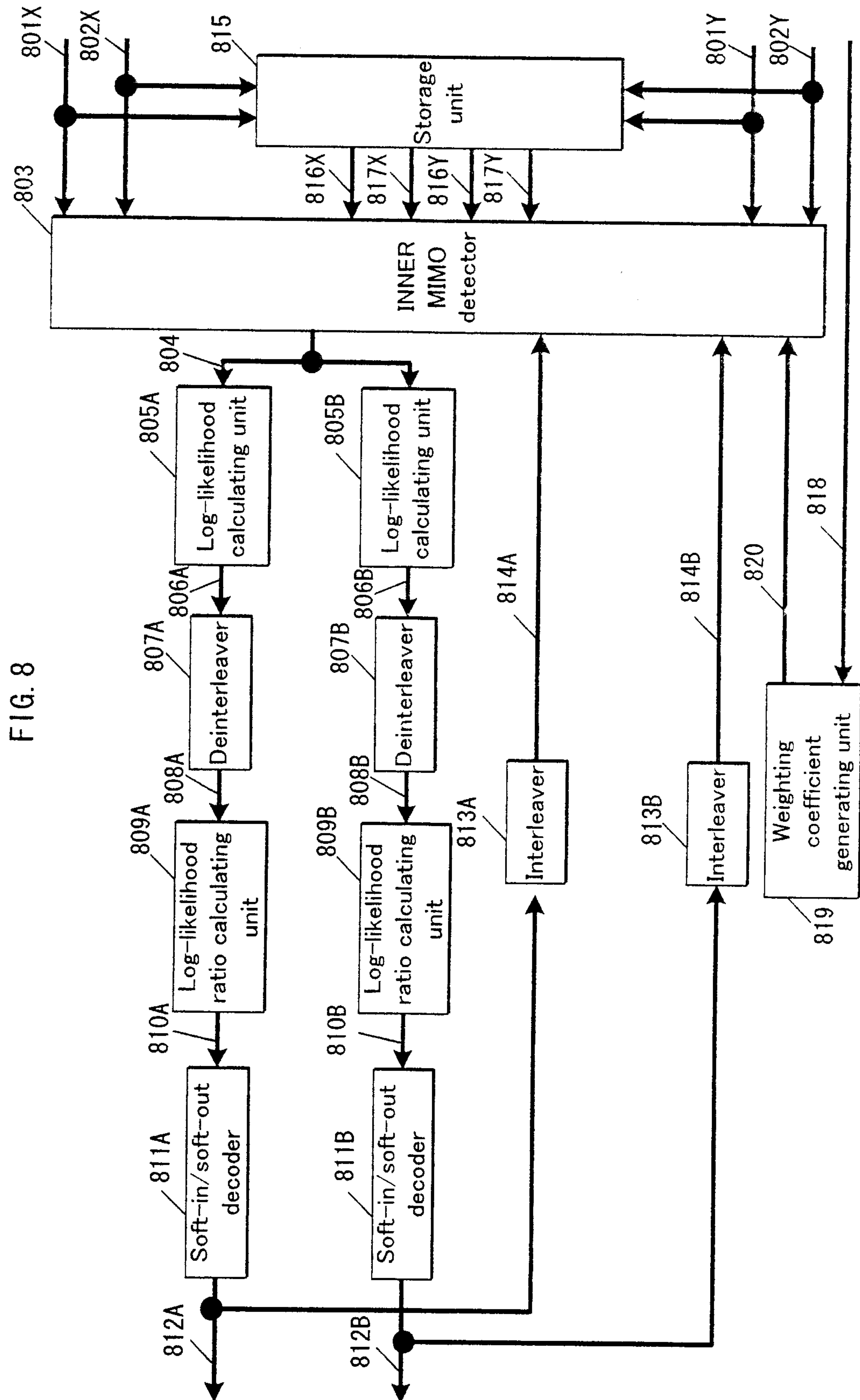


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FIG. 7



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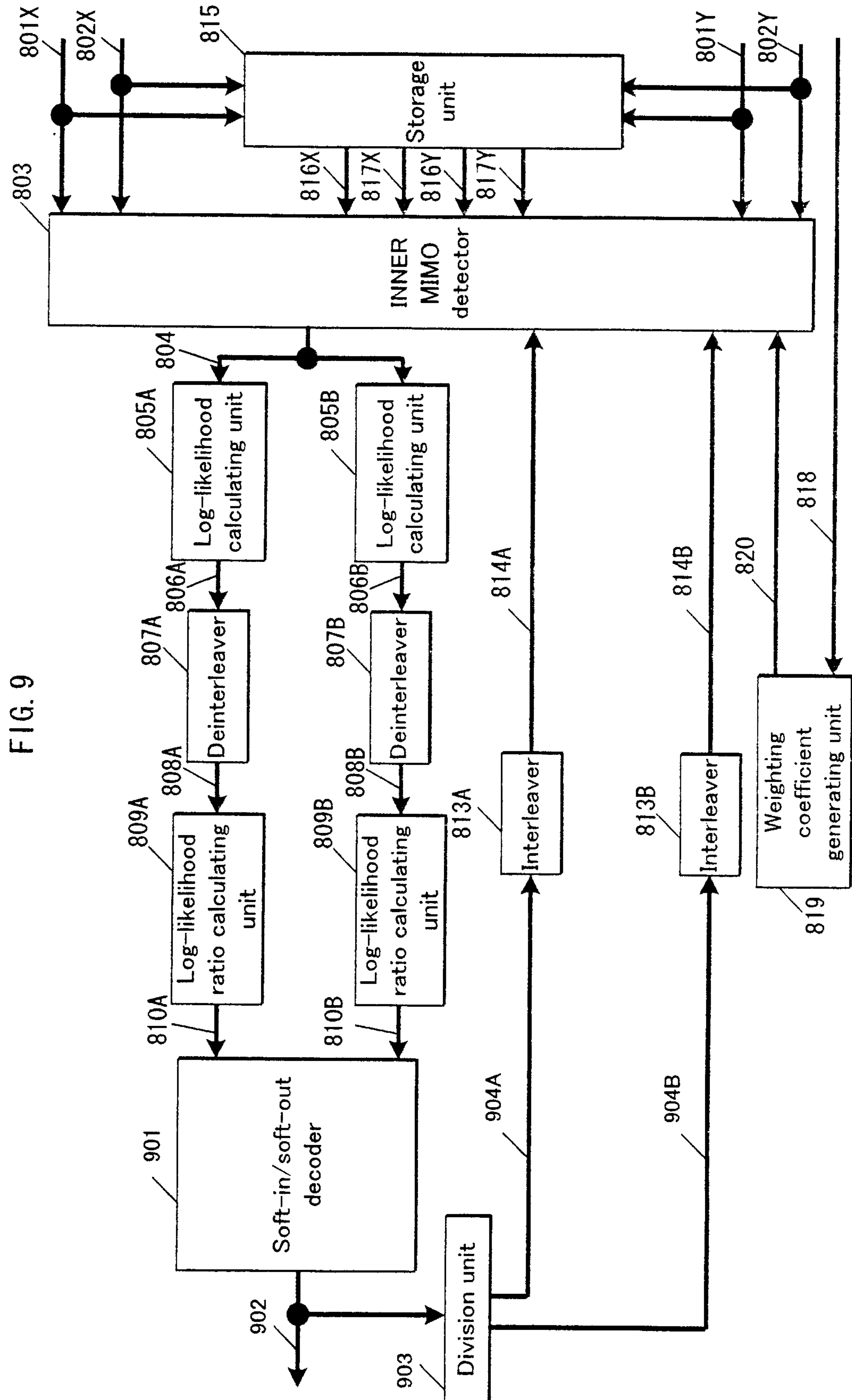
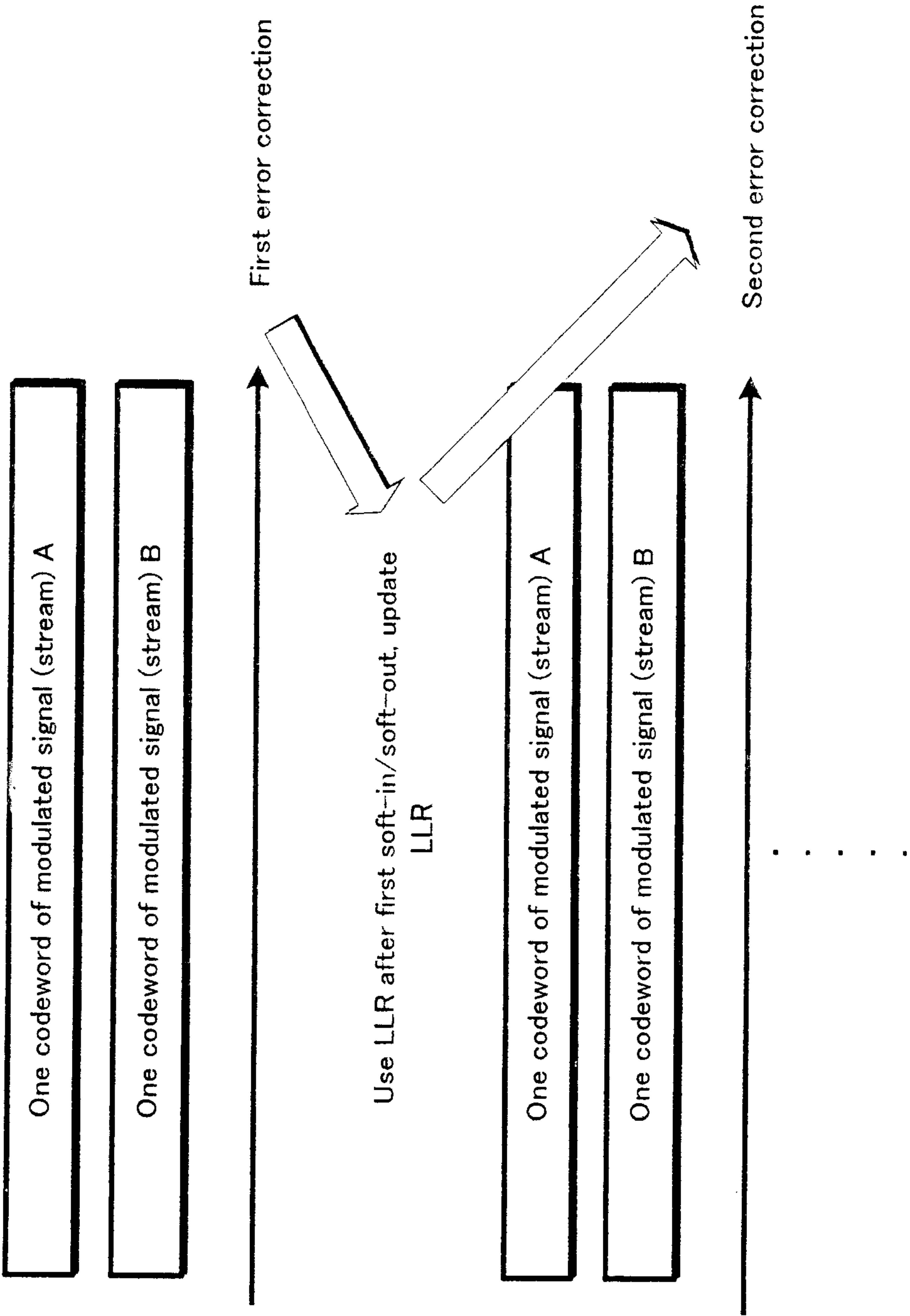
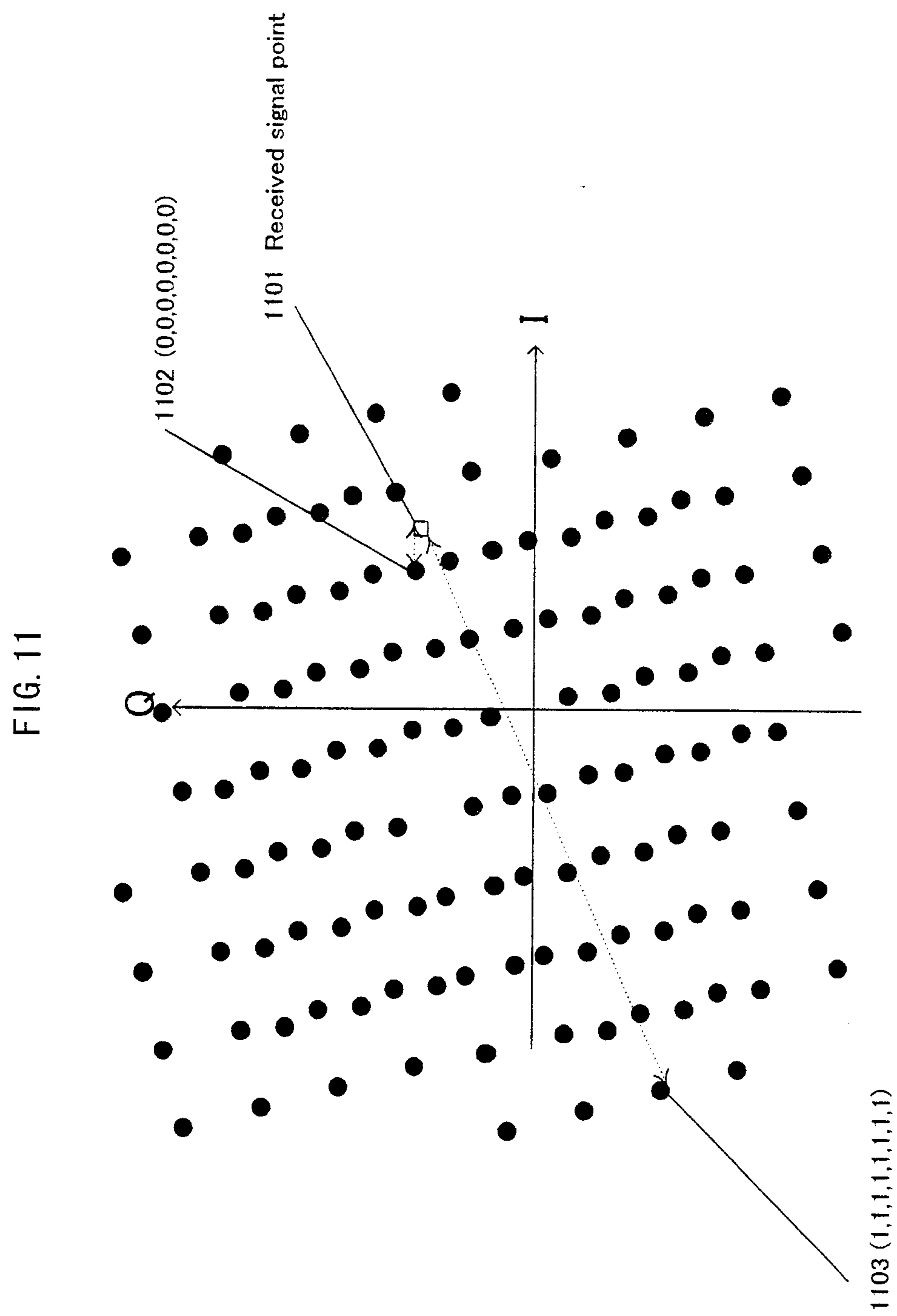


FIG. 10



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FIG. 12B

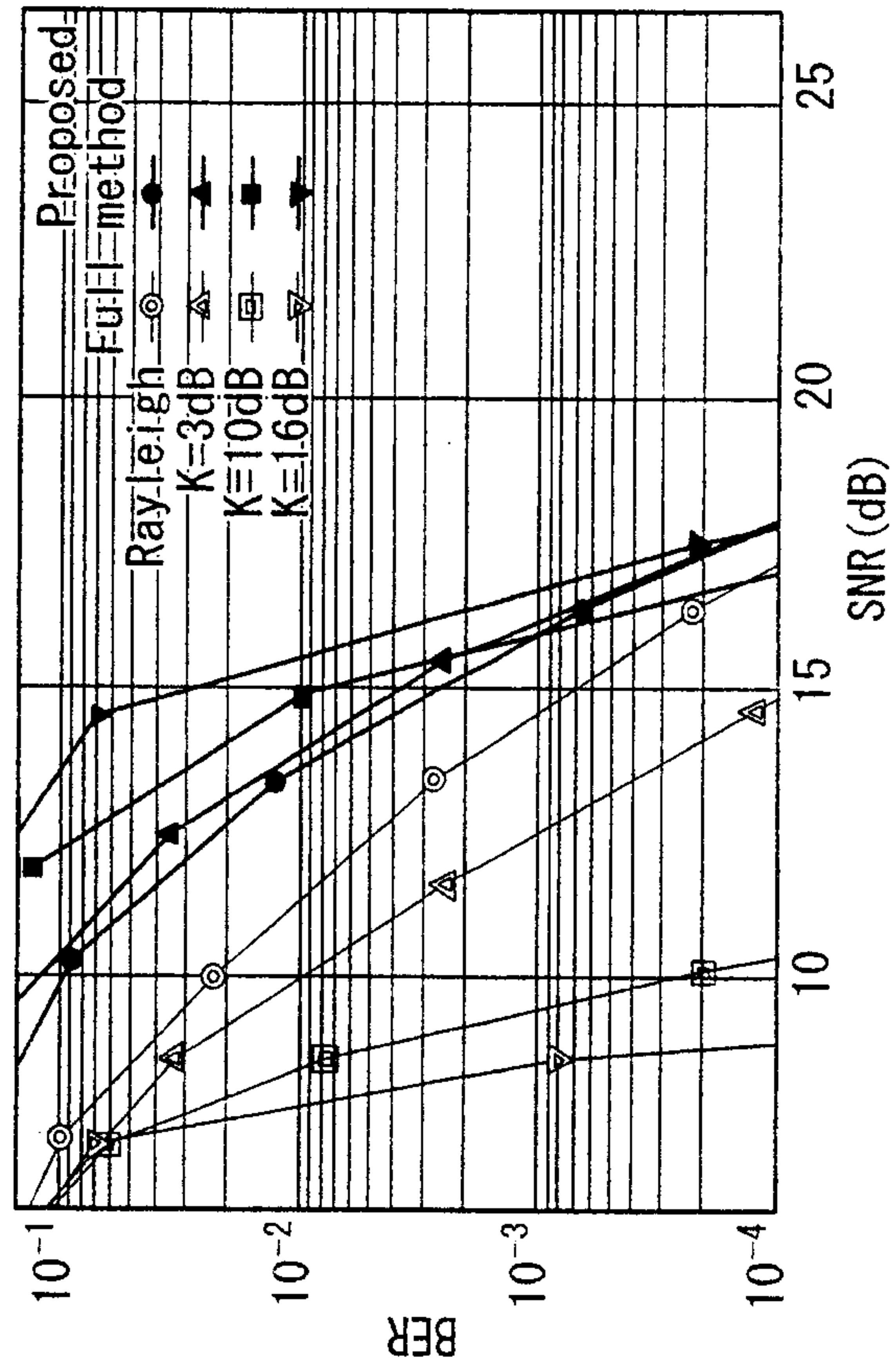
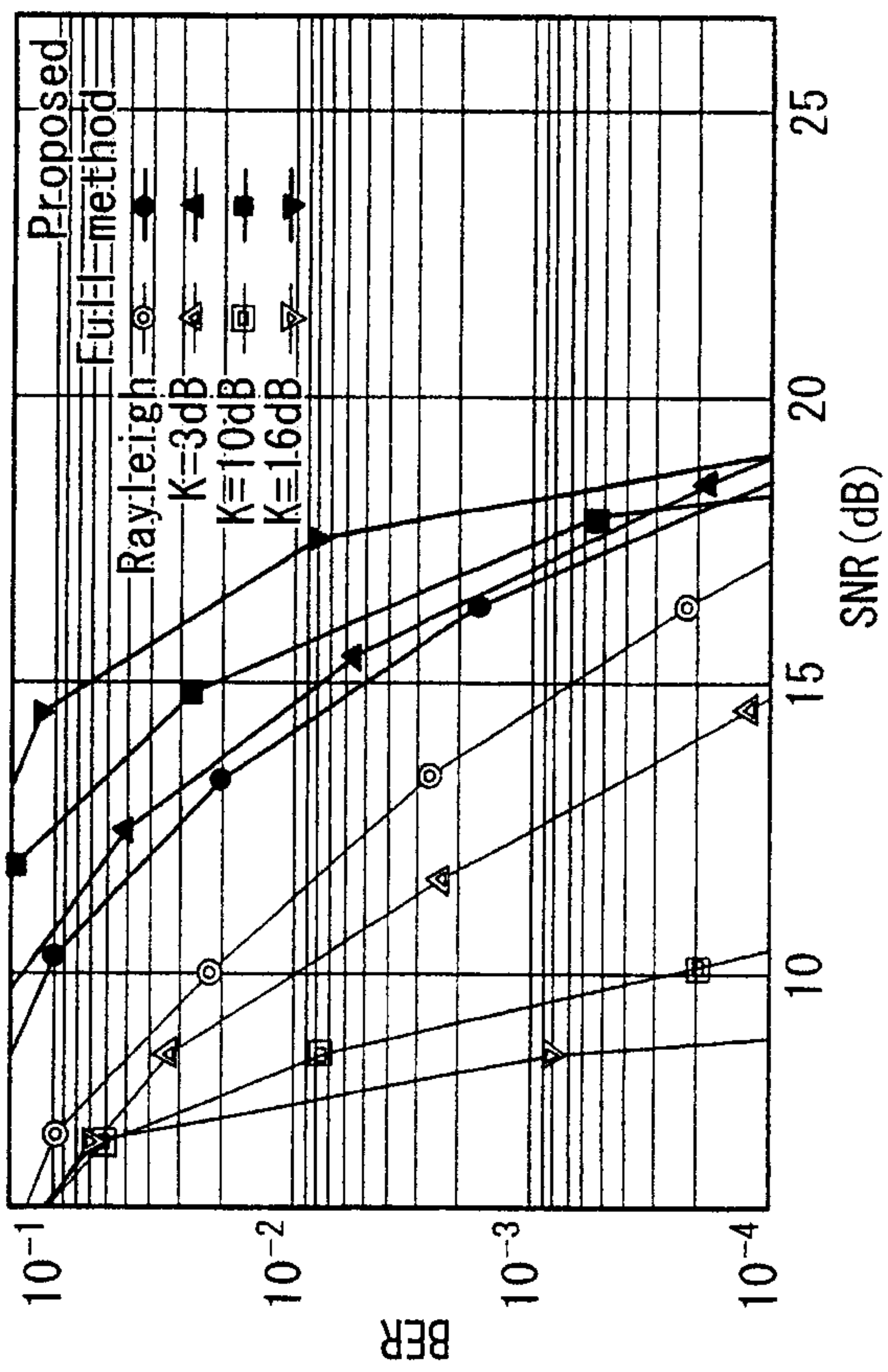


FIG. 12A



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FIG. 13

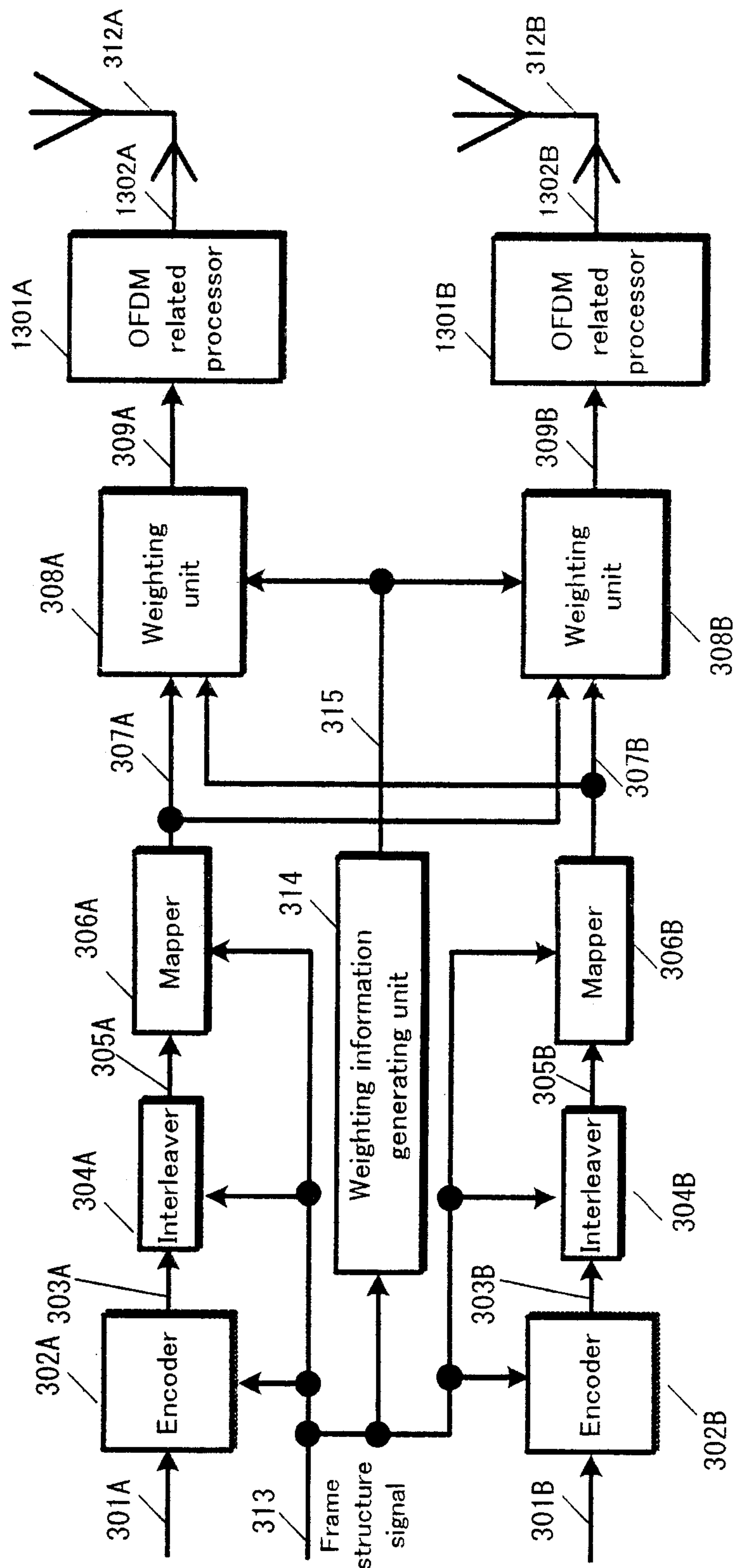


FIG. 14

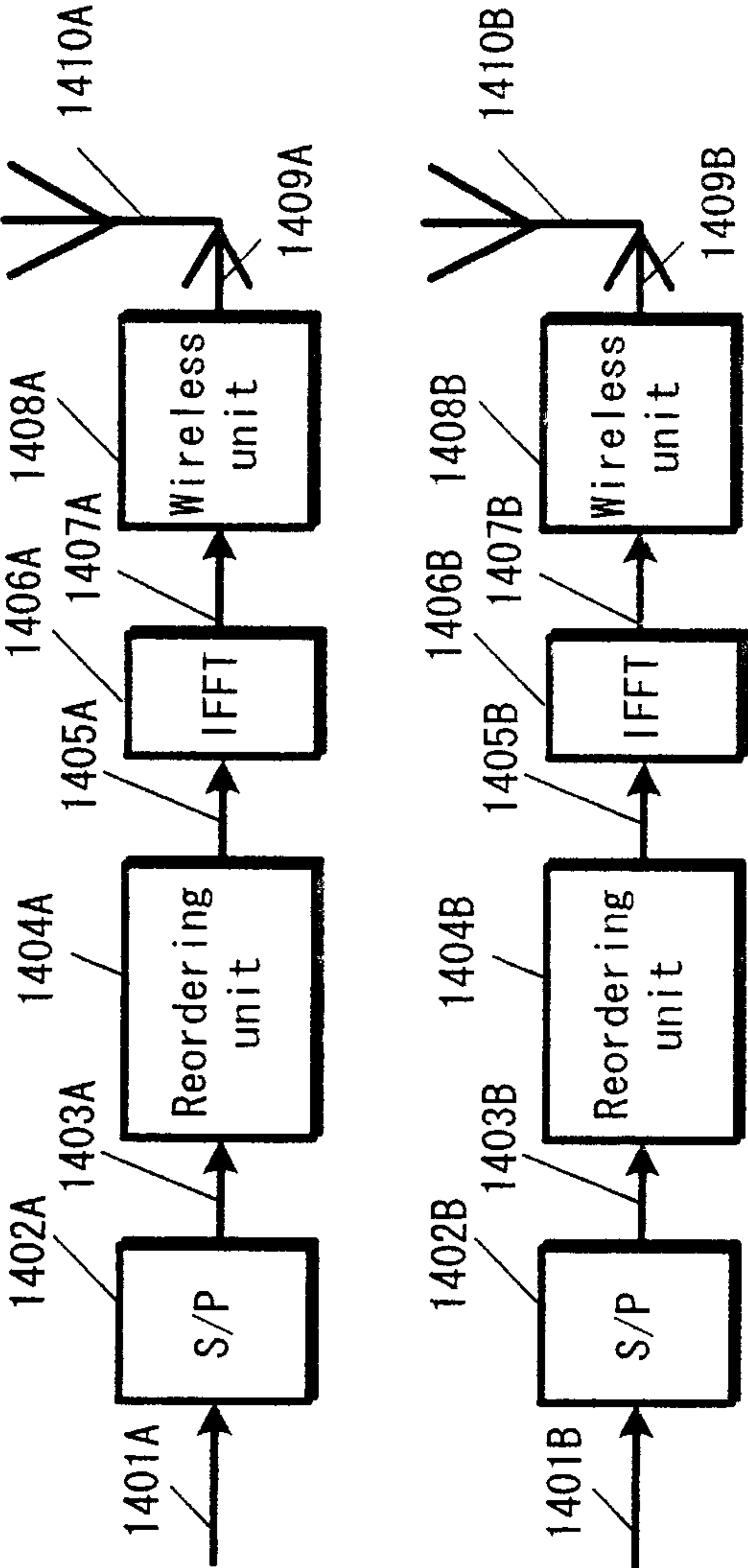


FIG. 15A

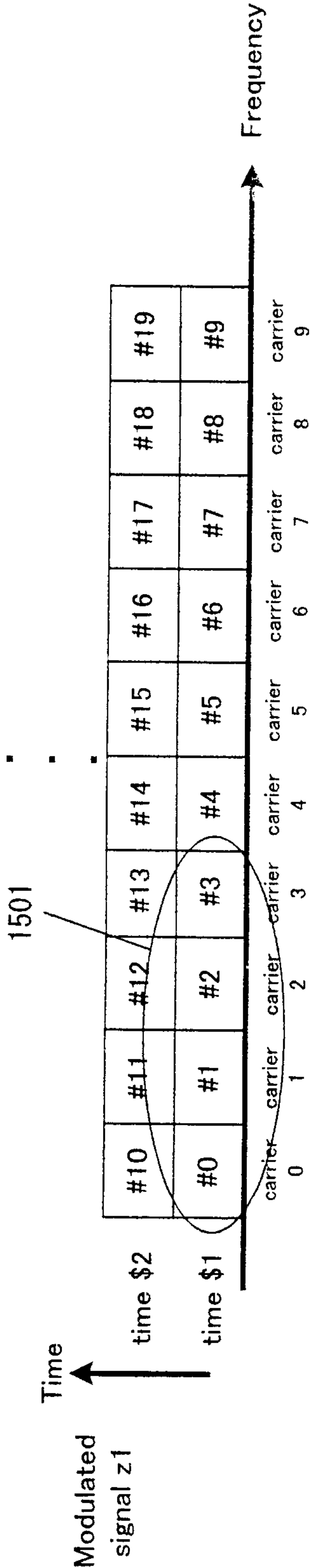


FIG. 15B

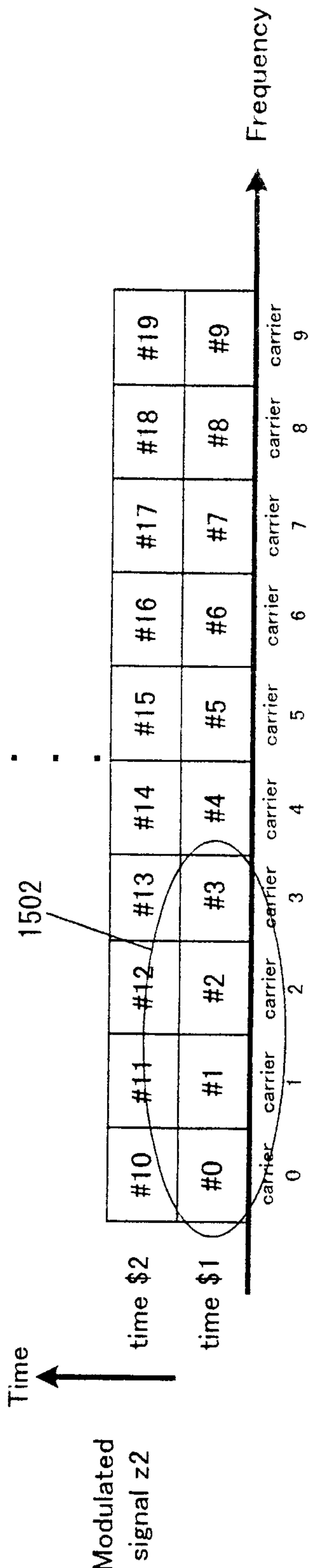


FIG. 16A

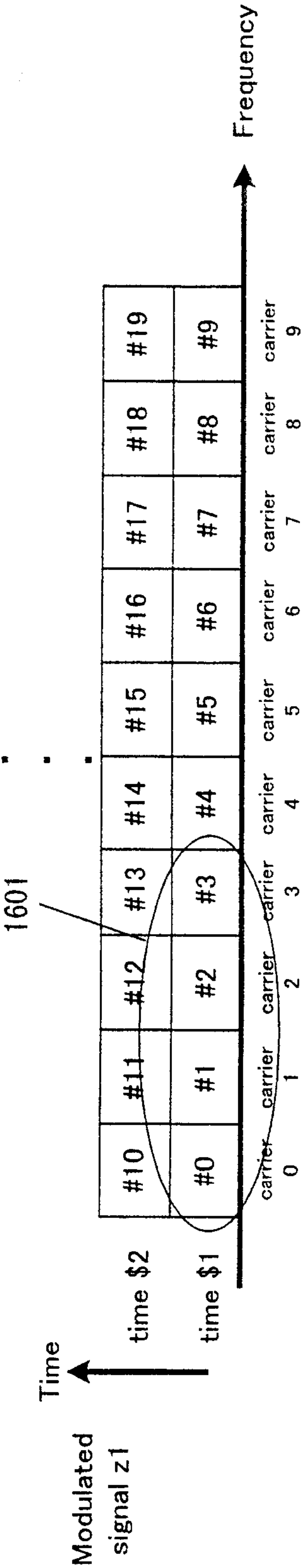


FIG. 16B

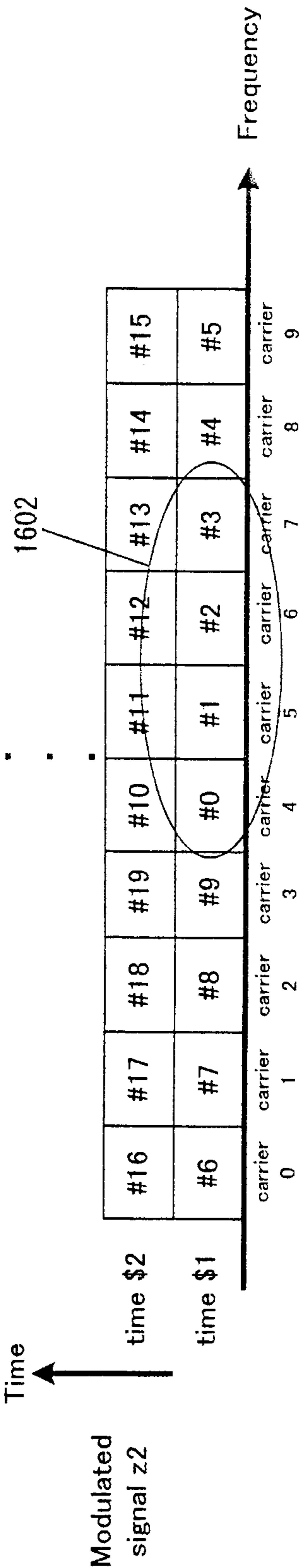


FIG. 17A

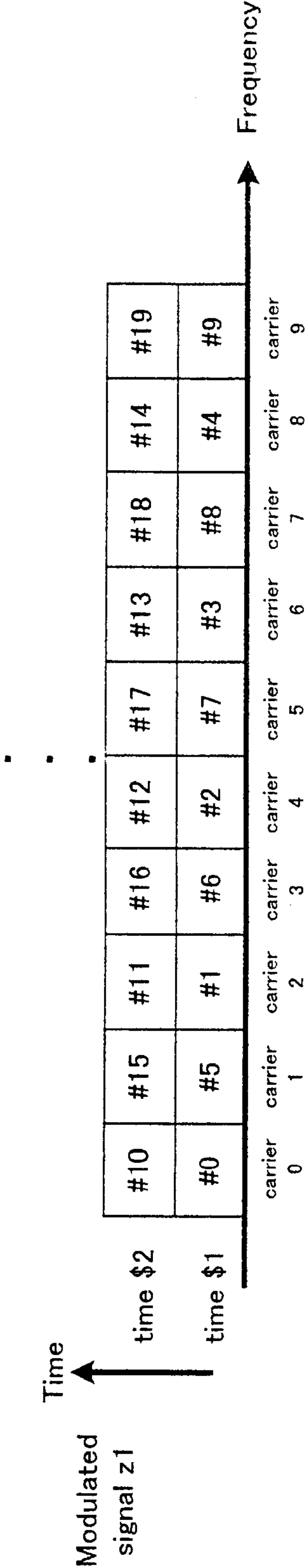


FIG. 17B

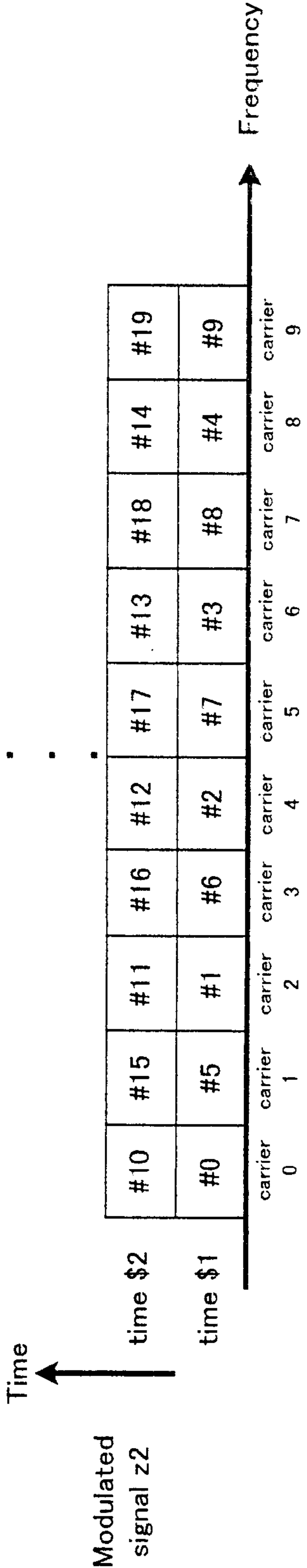


FIG. 18A

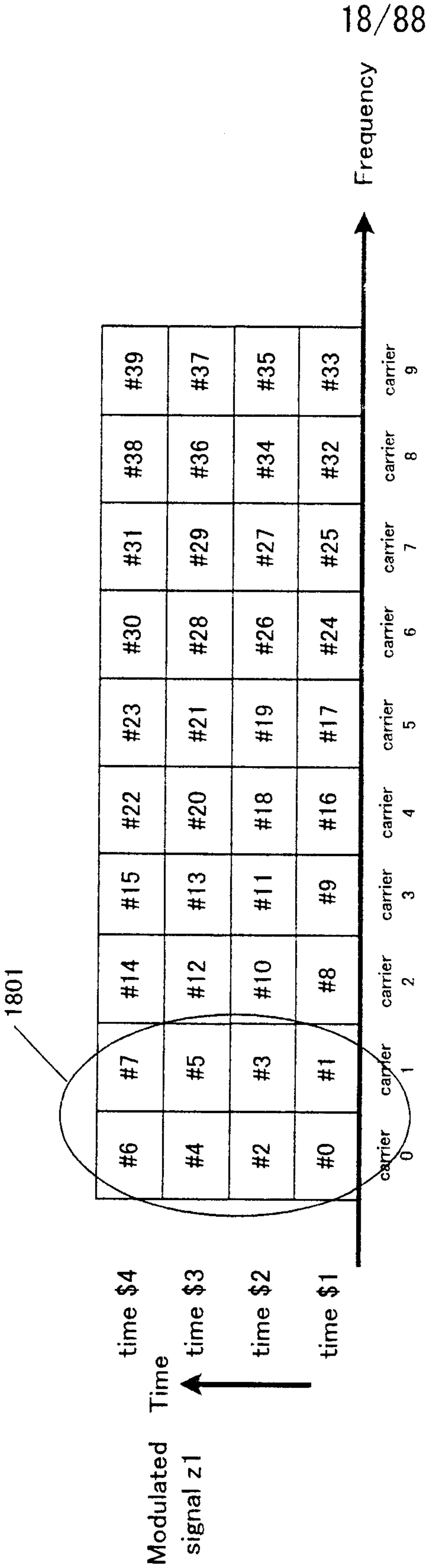


FIG. 18B

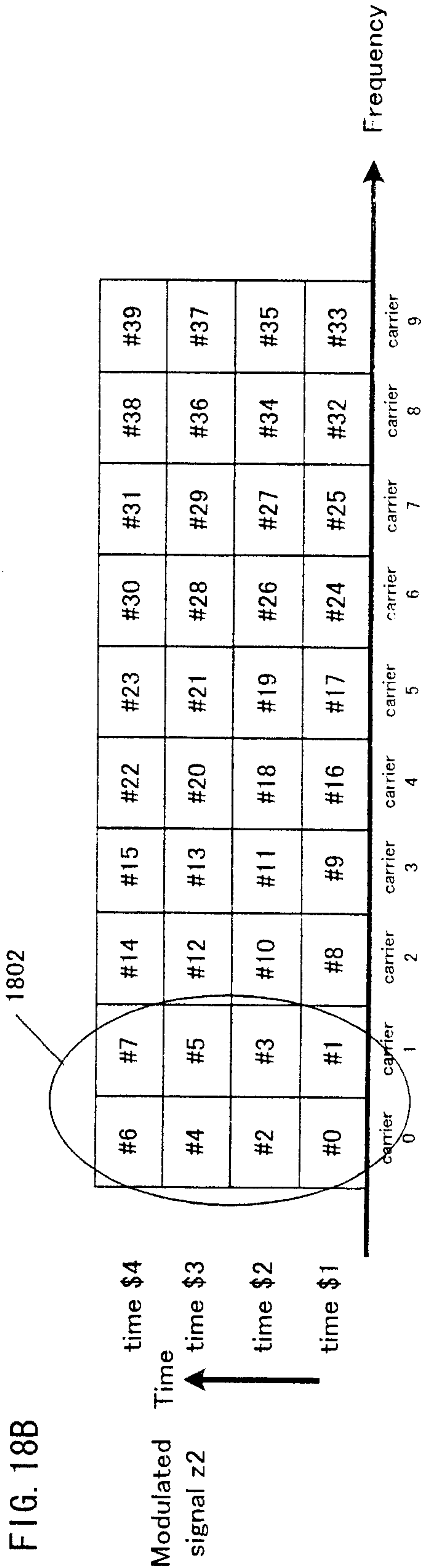


FIG. 19A

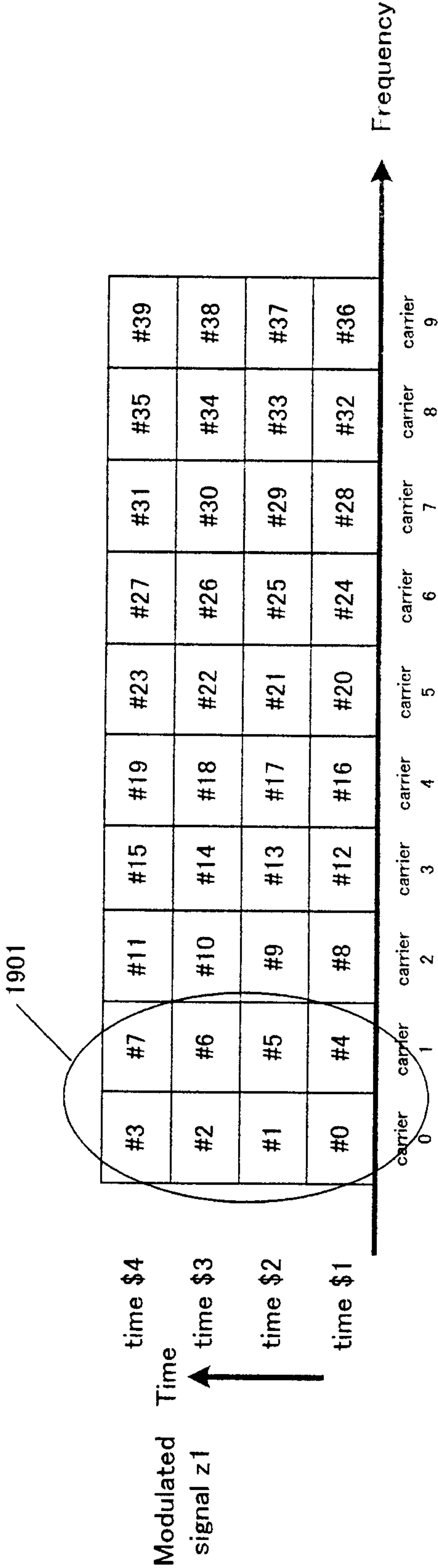
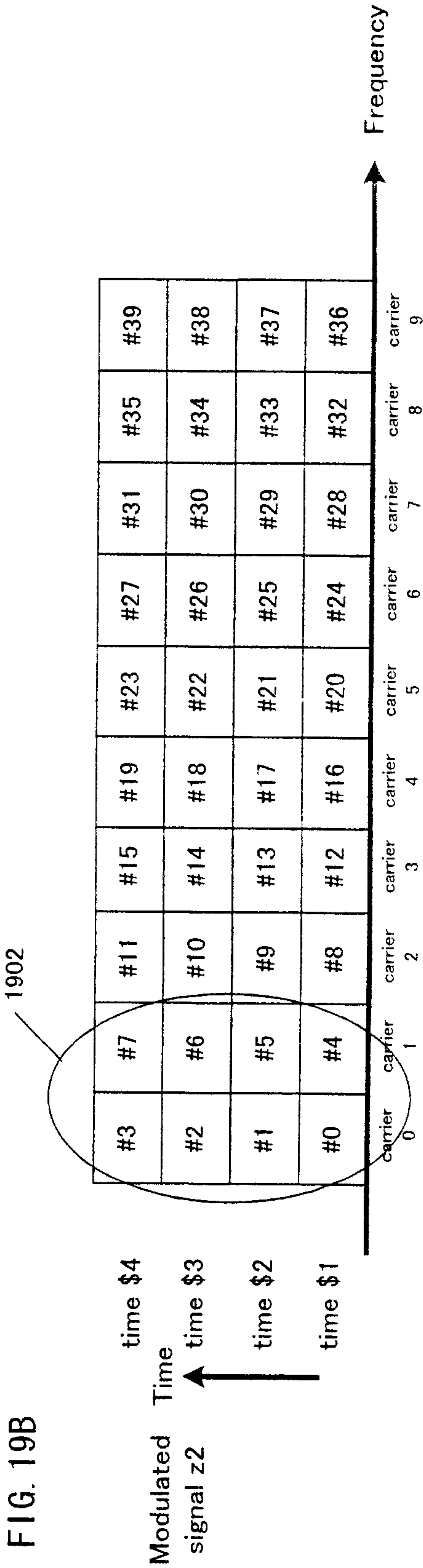


FIG. 19B



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FIG. 20

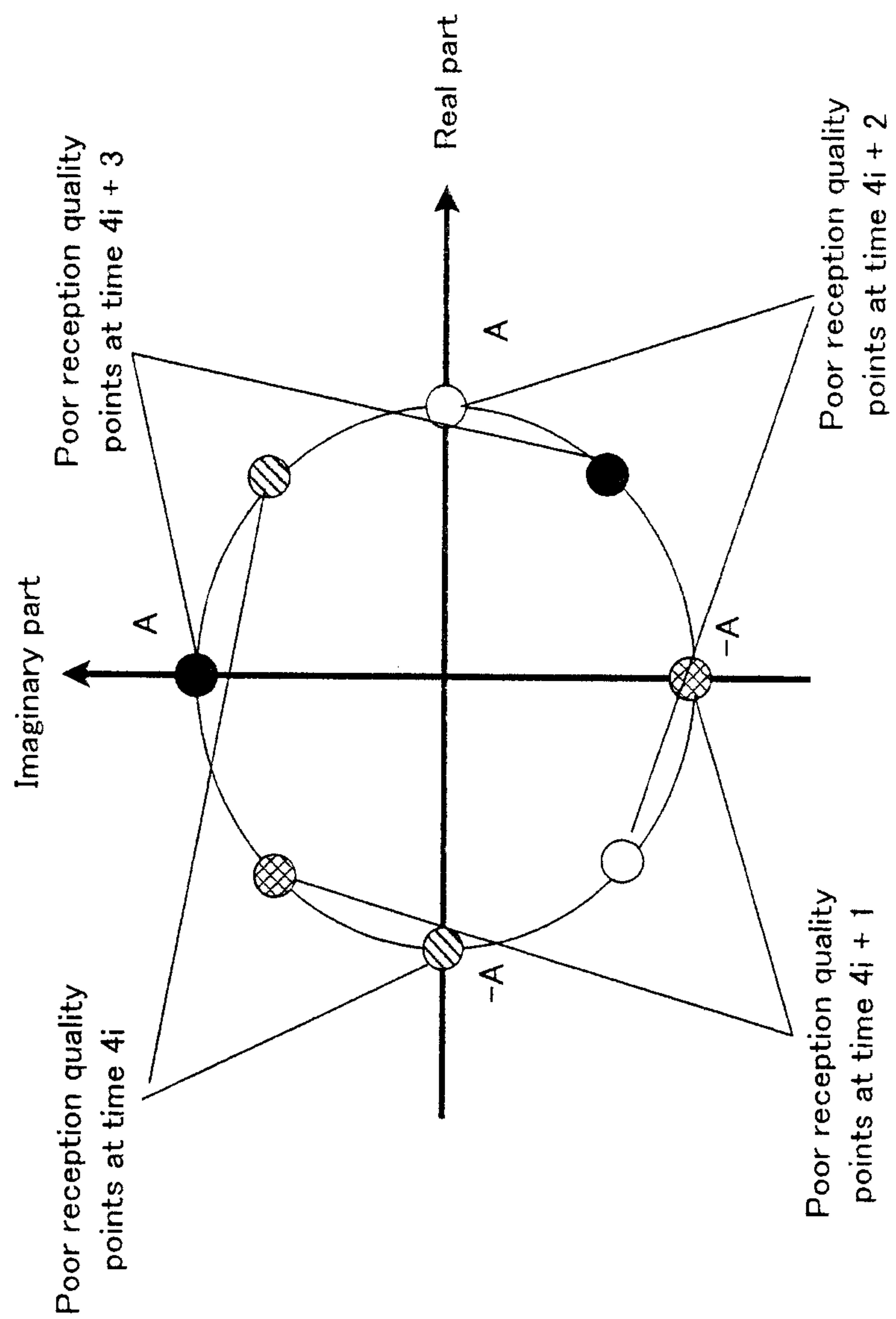


FIG. 21

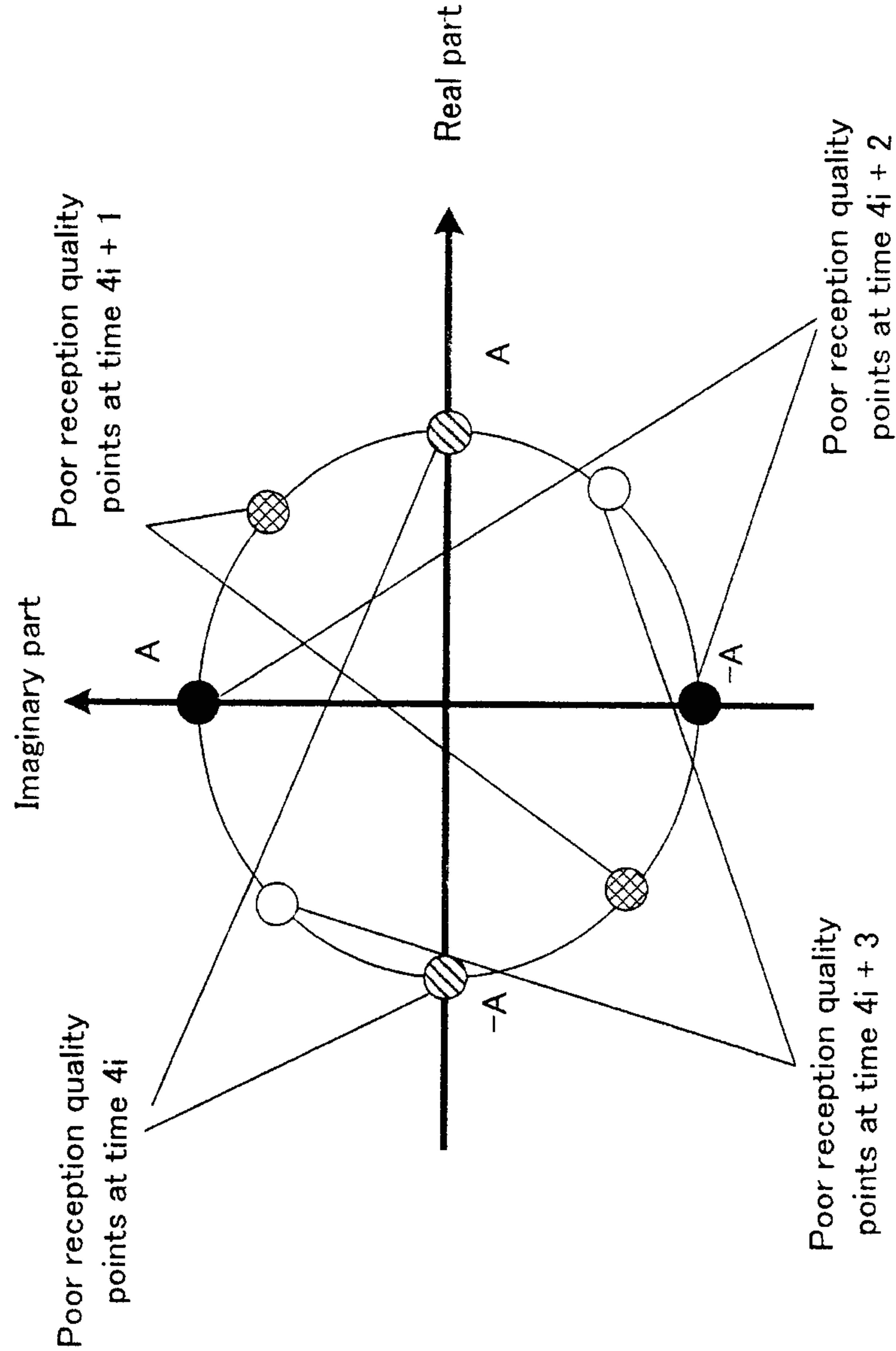


FIG. 22

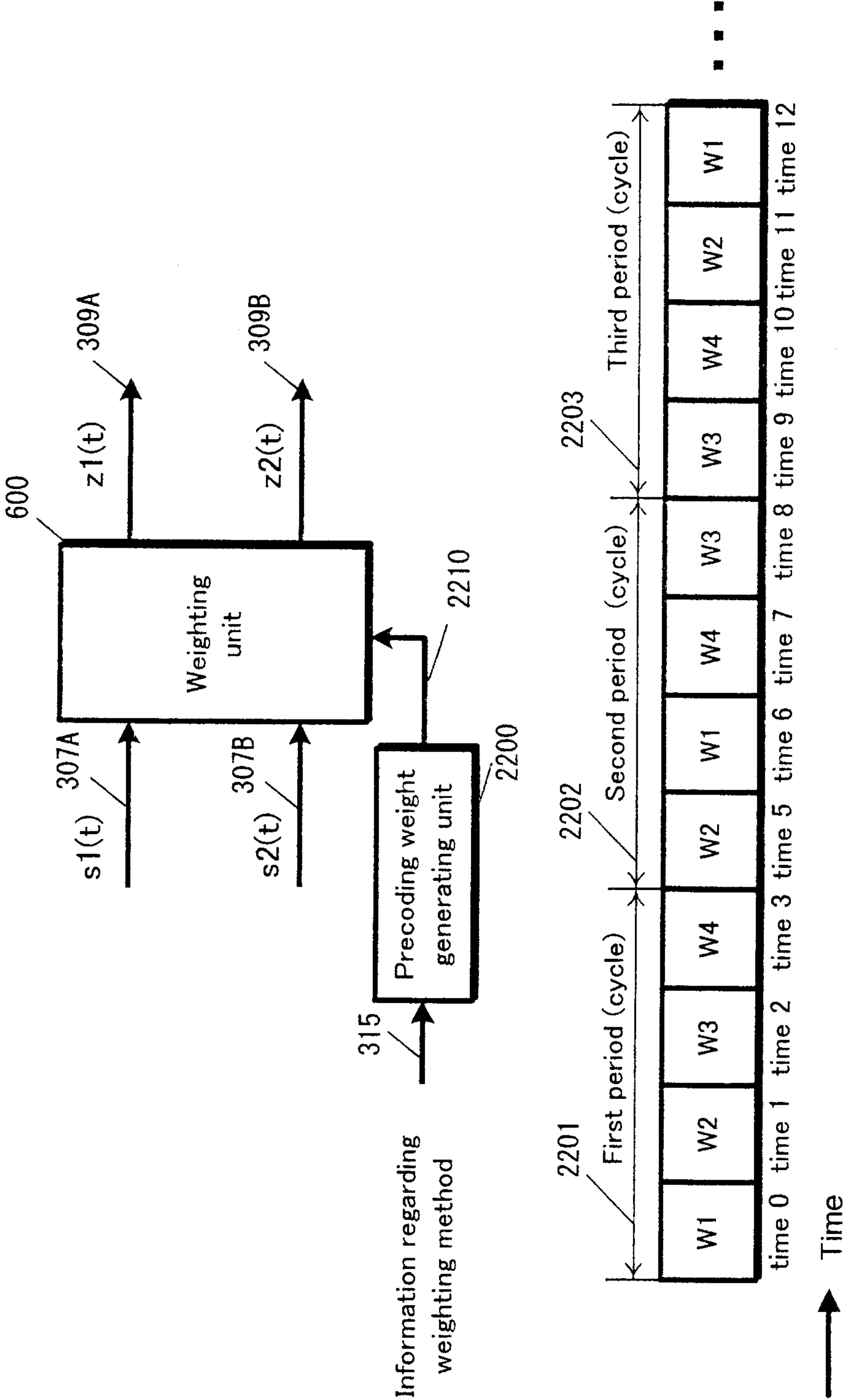
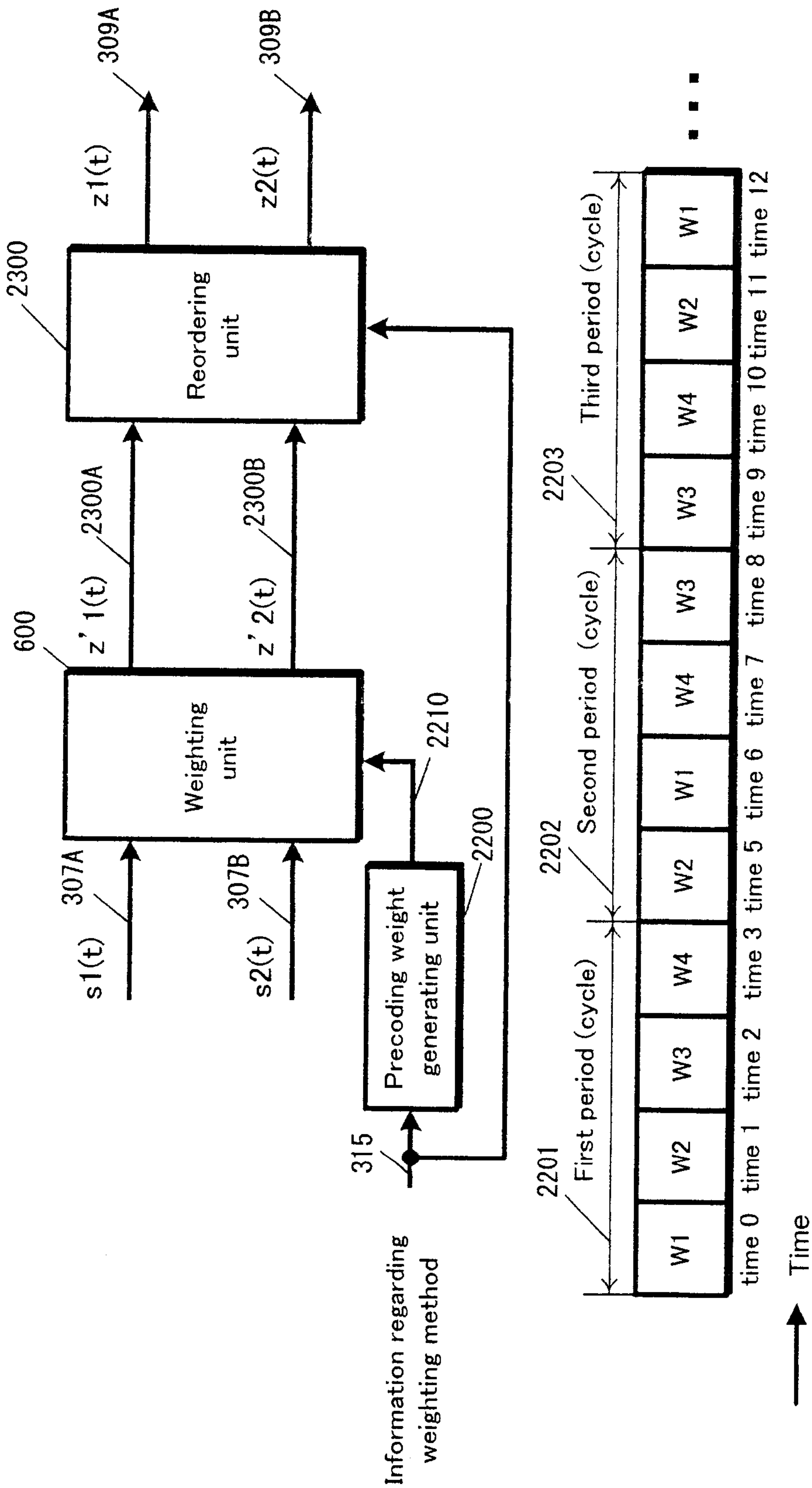


FIG. 23



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FIG. 24B

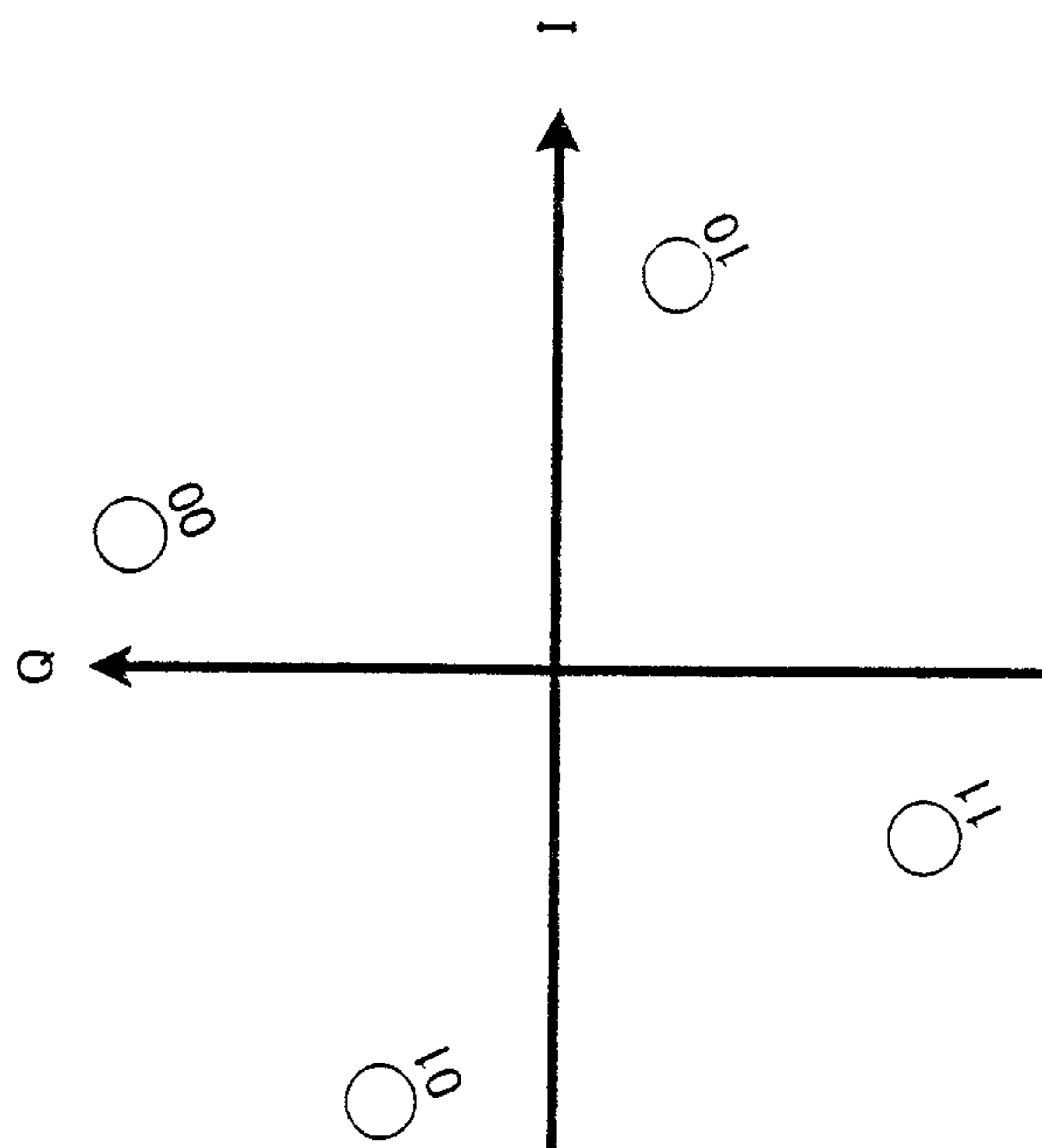
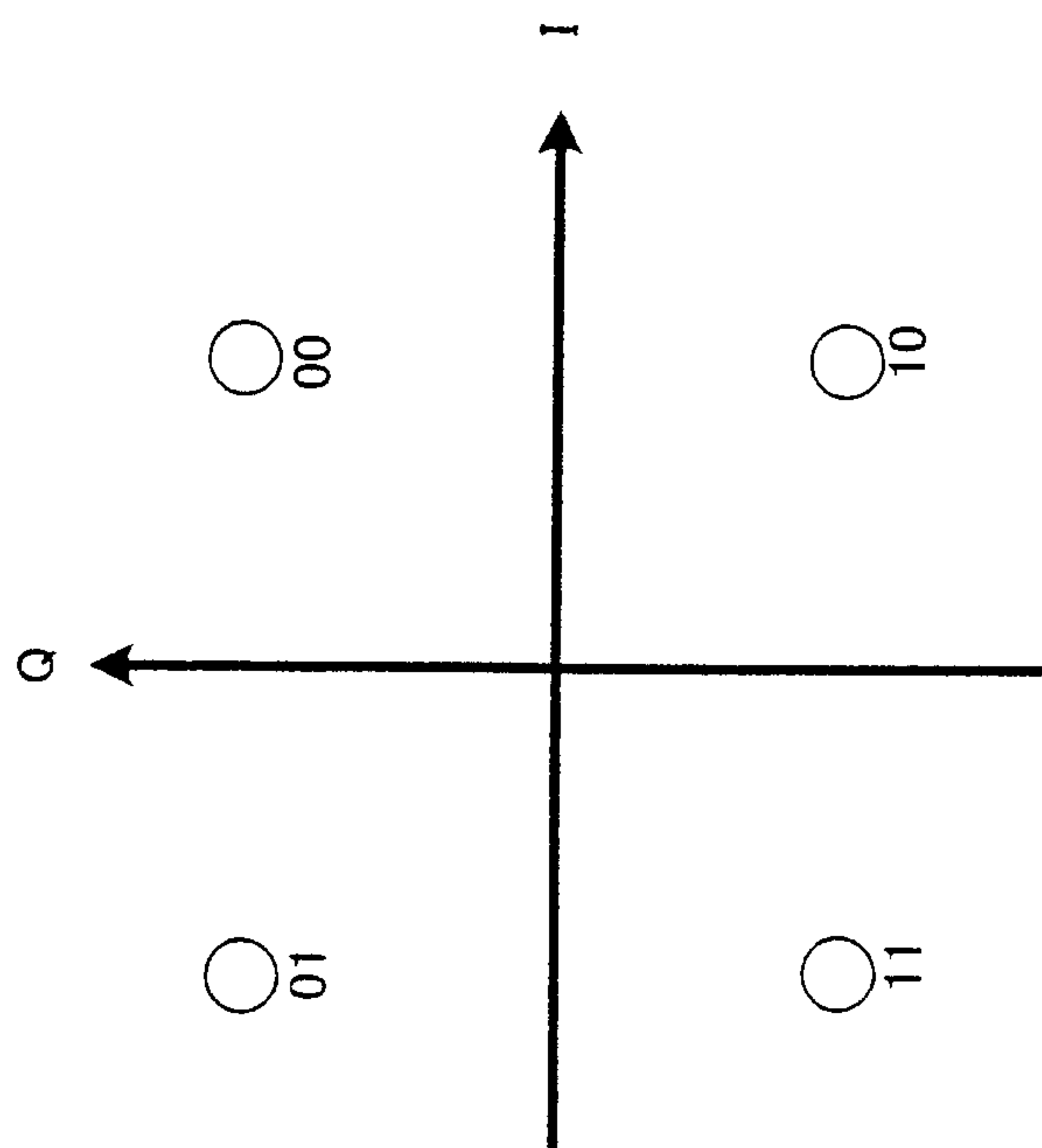


FIG. 24A



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FIG. 25B

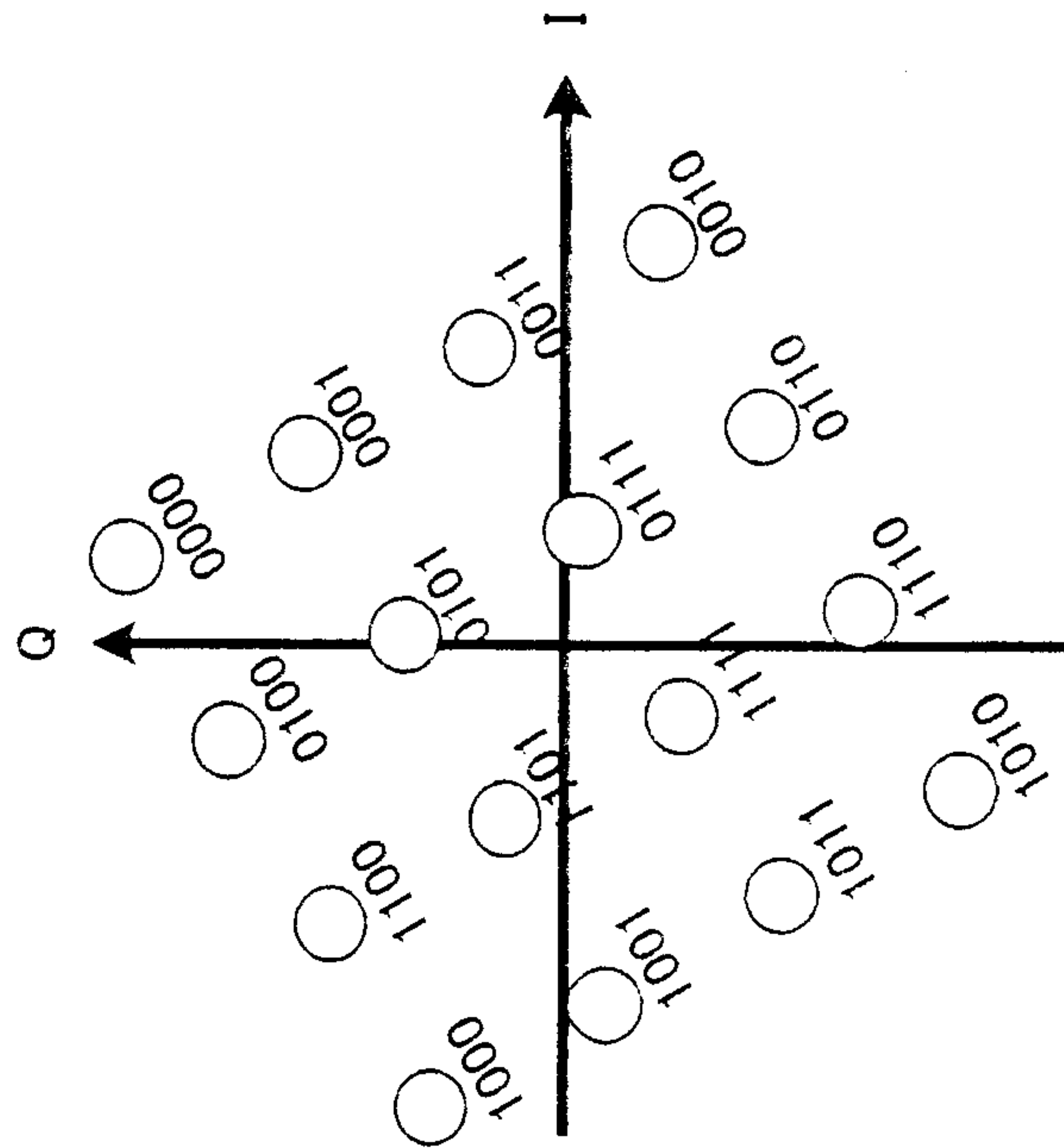
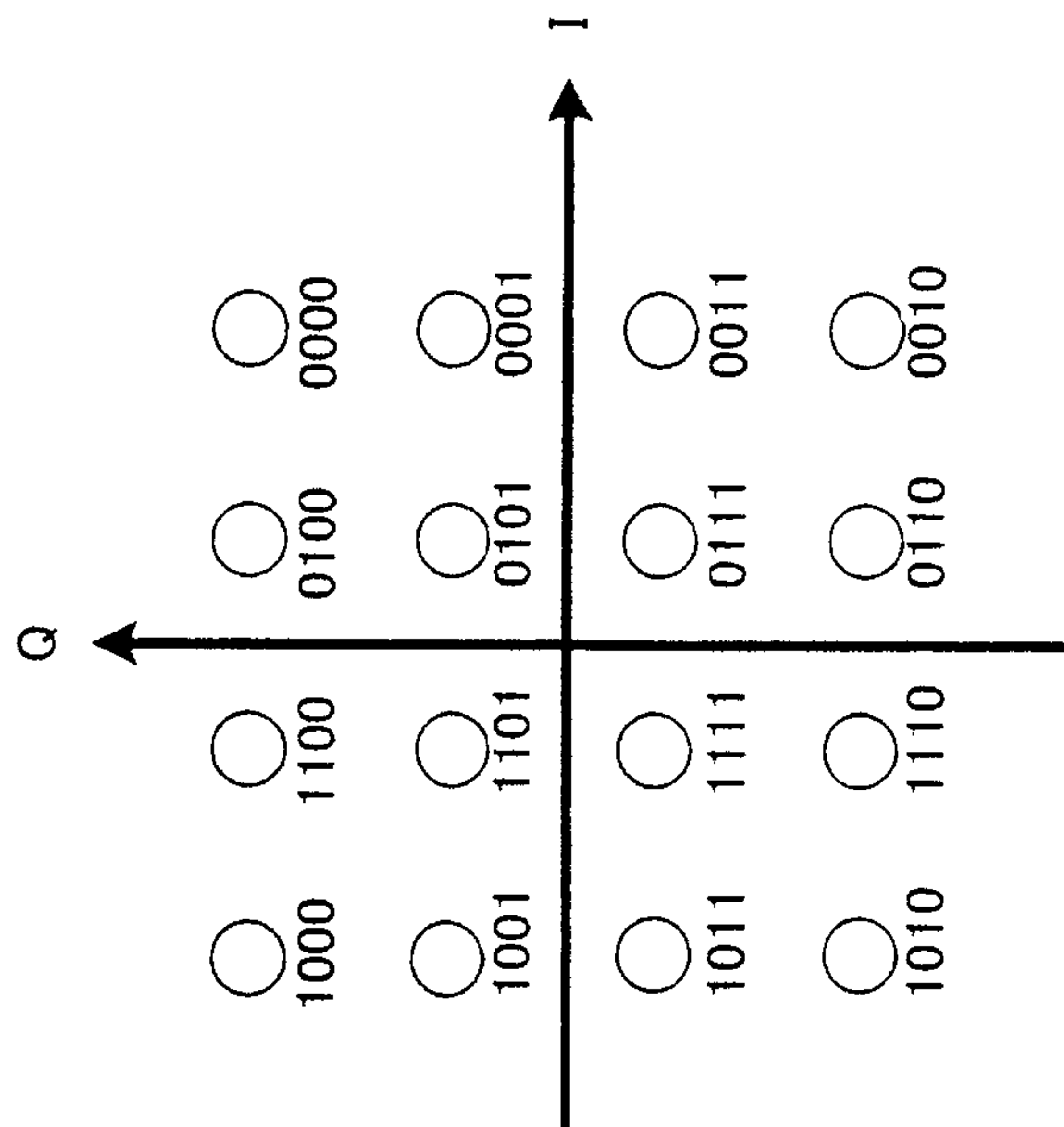


FIG. 25A



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FIG. 26

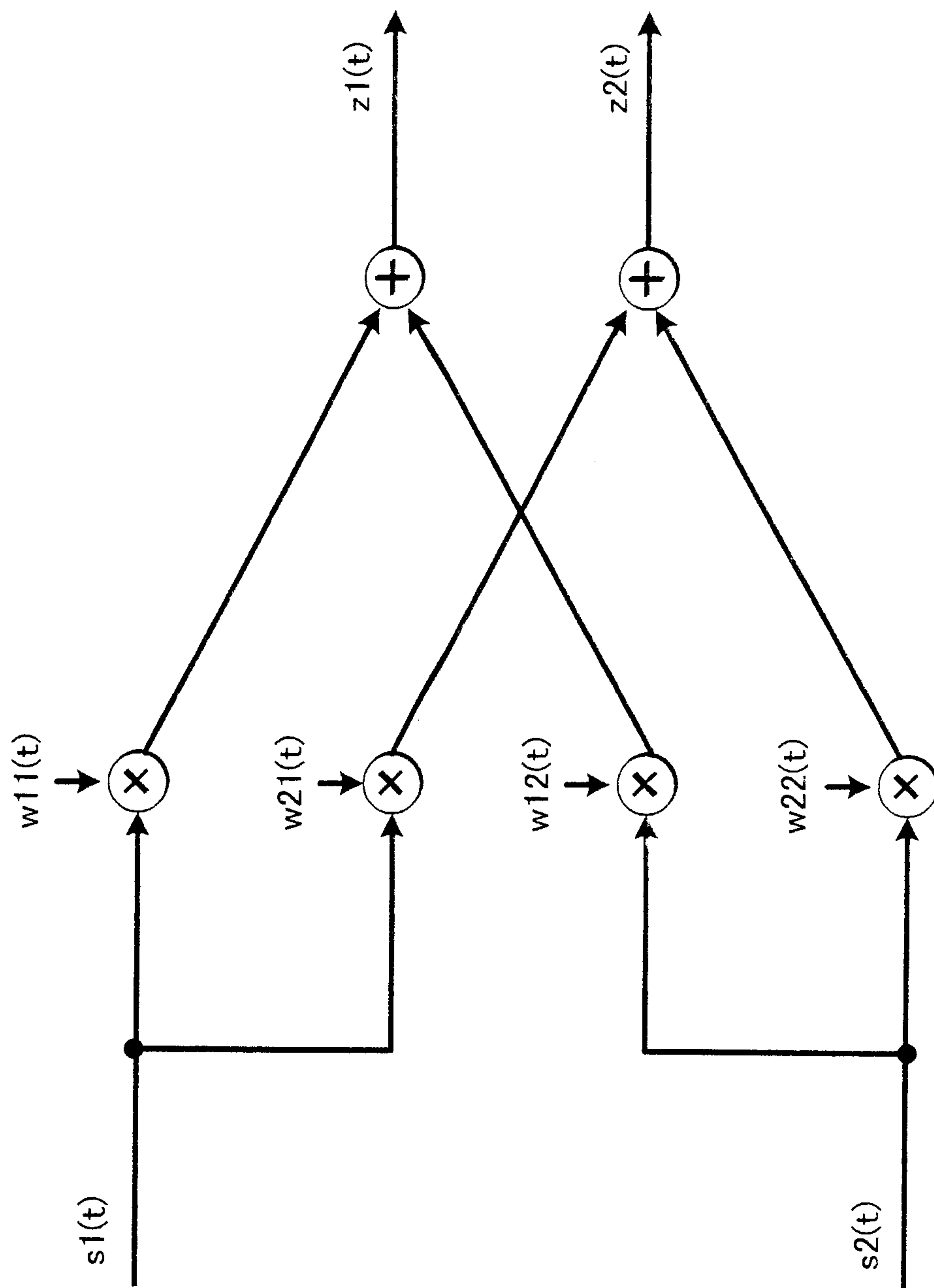
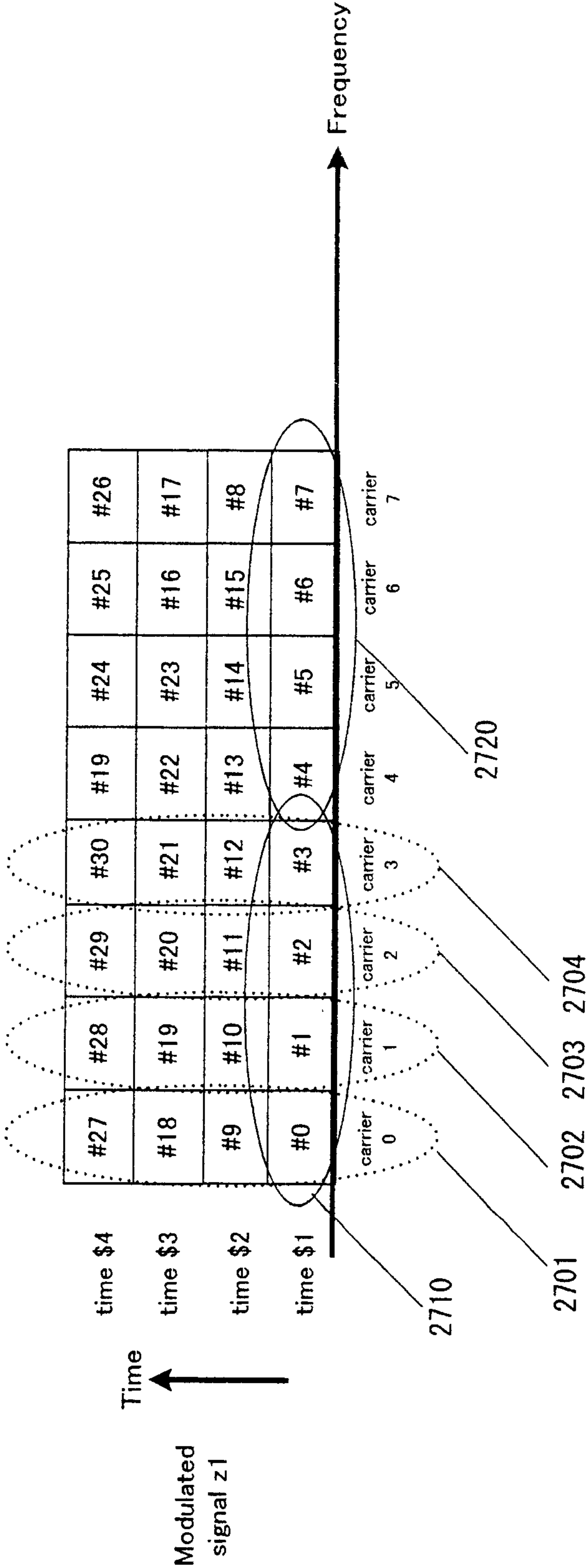
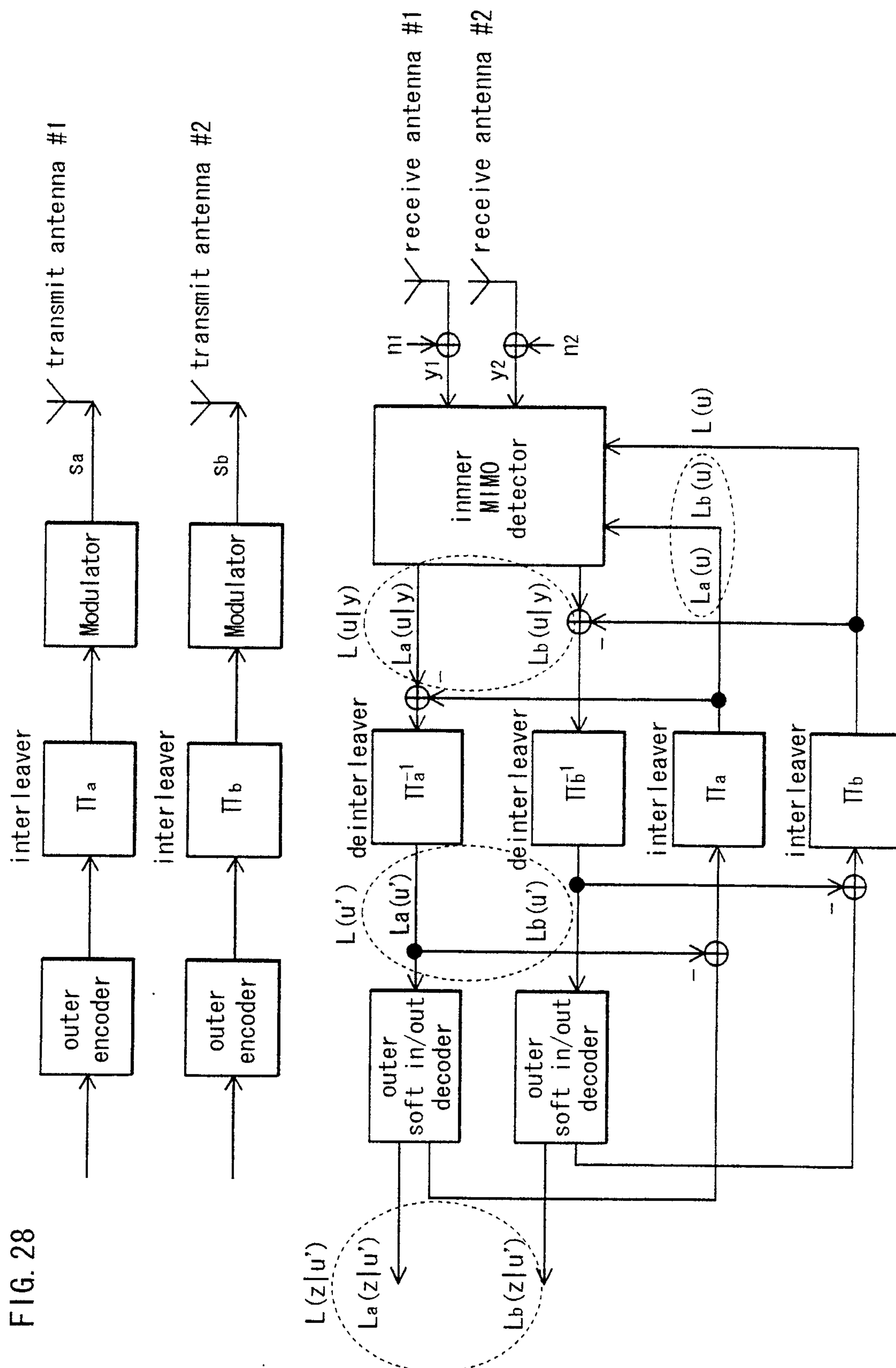


FIG. 27



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FIG. 29B

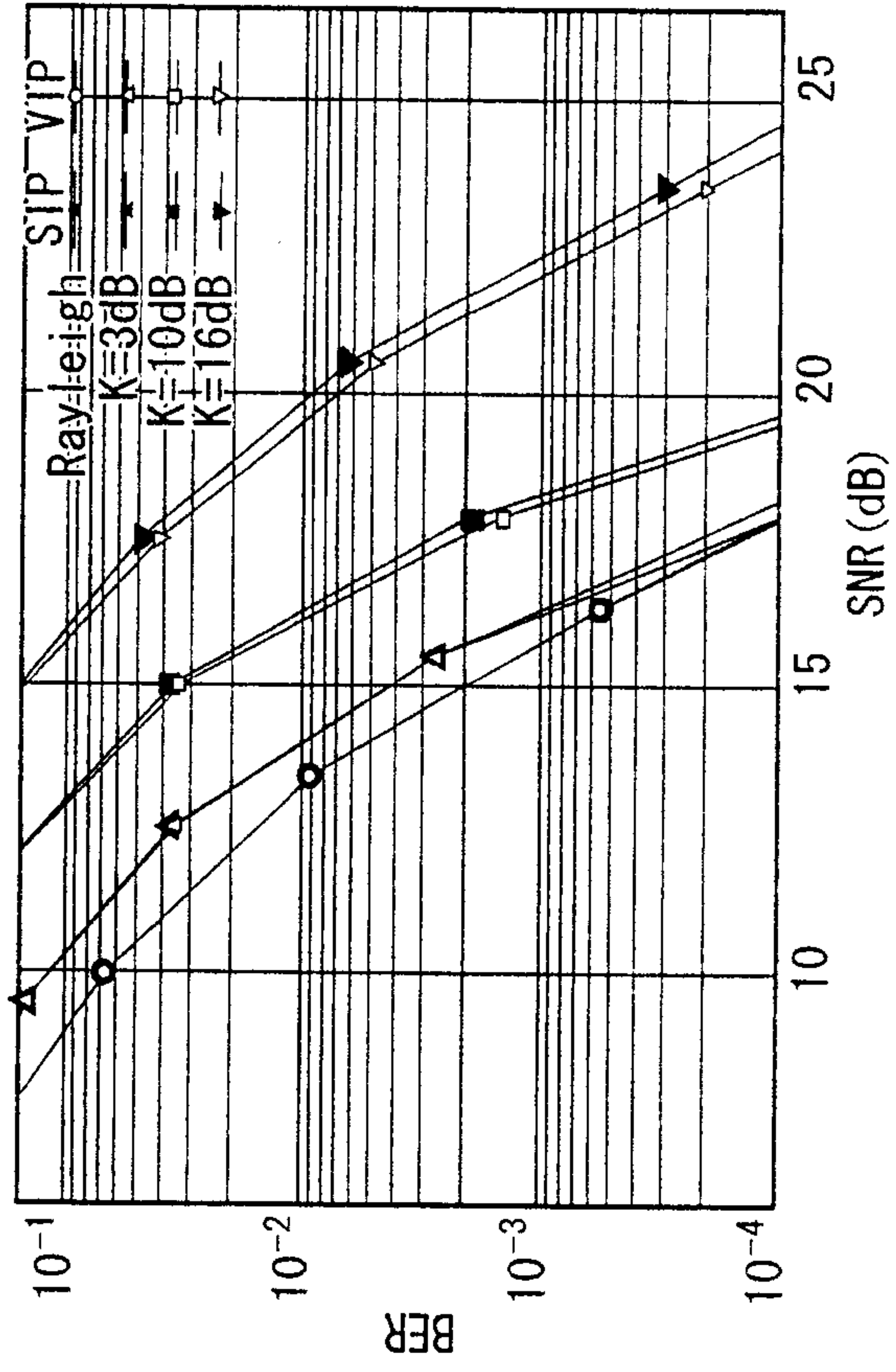
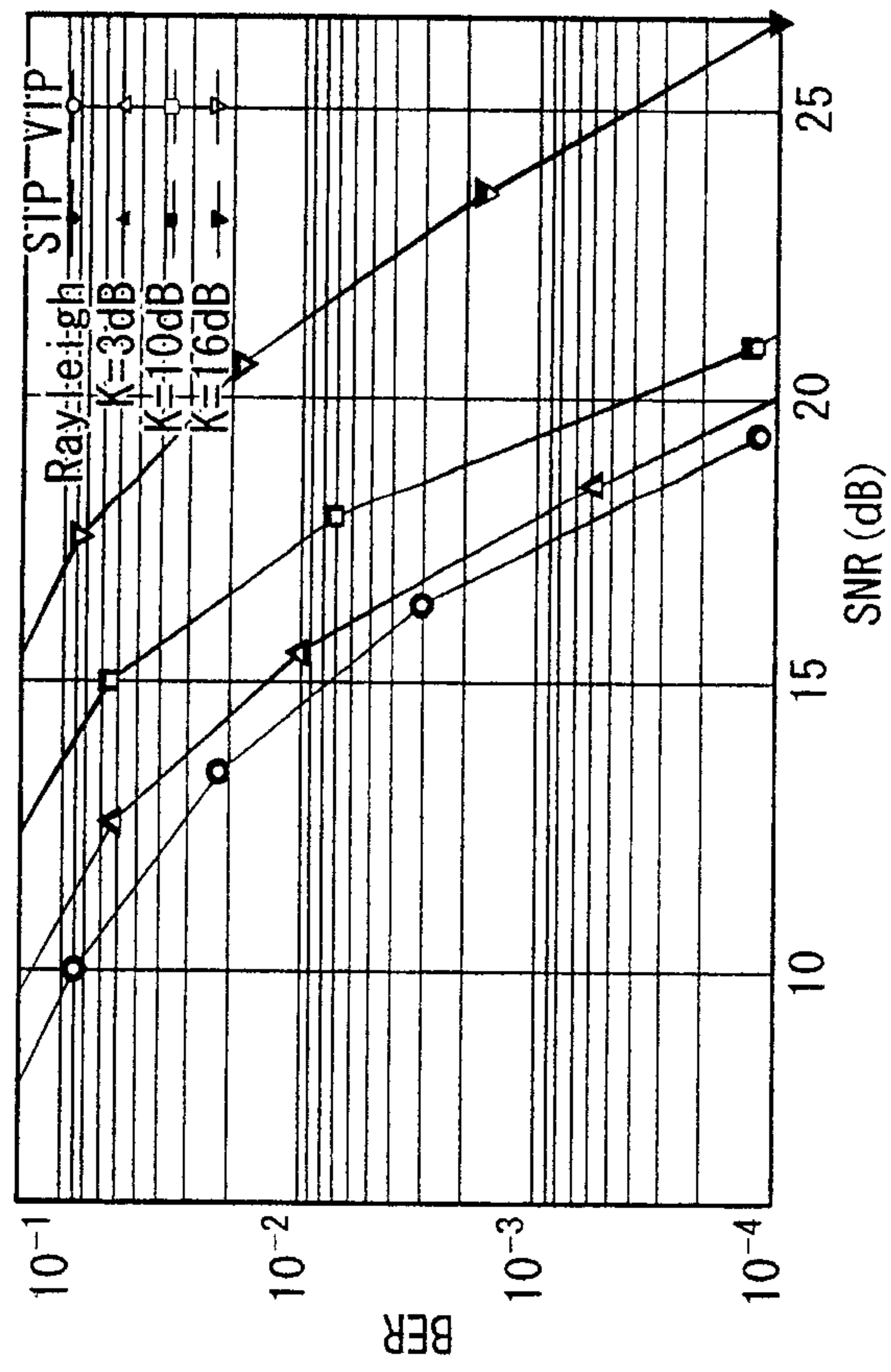


FIG. 29A



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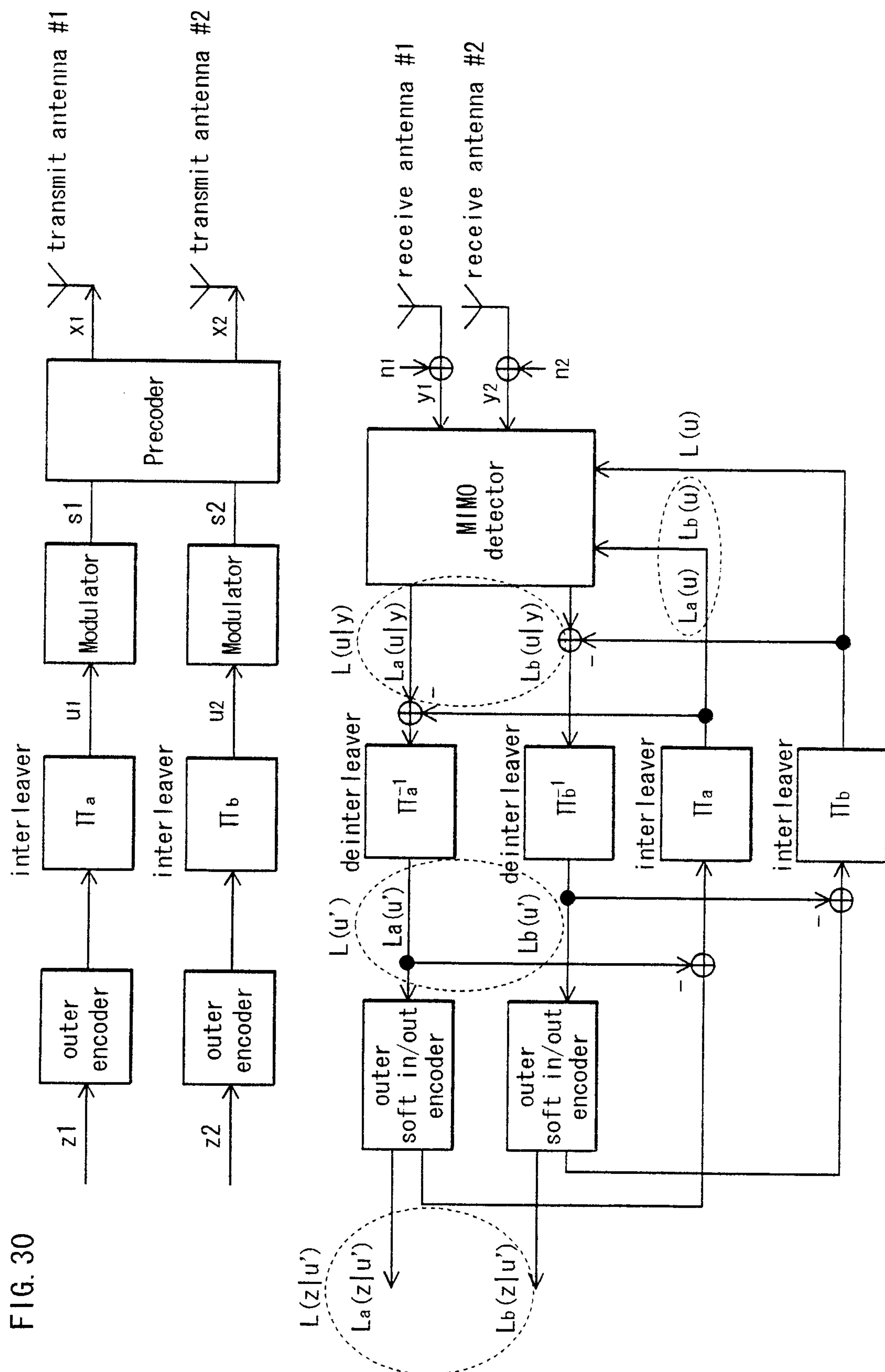


FIG. 31A

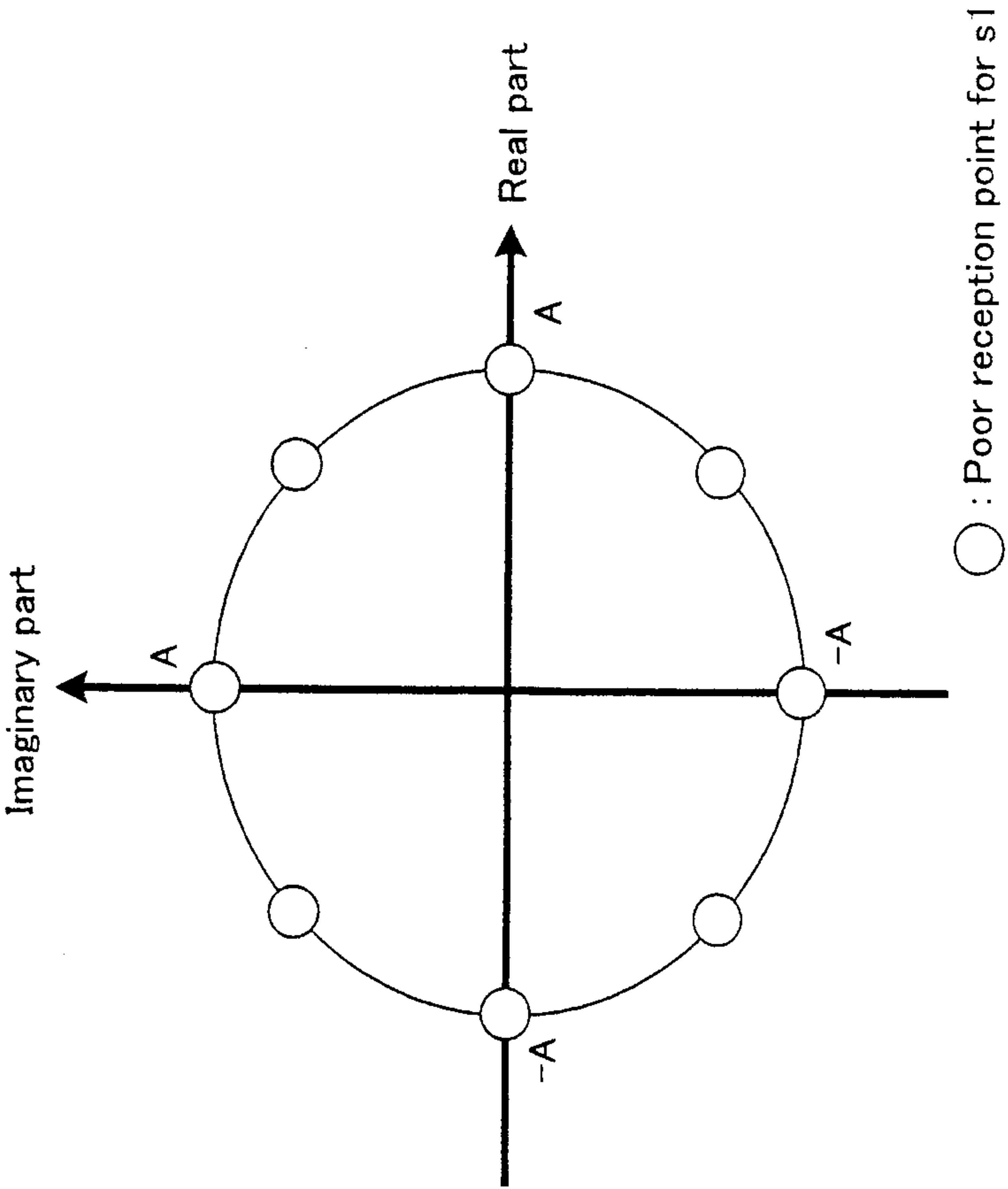


FIG. 31B

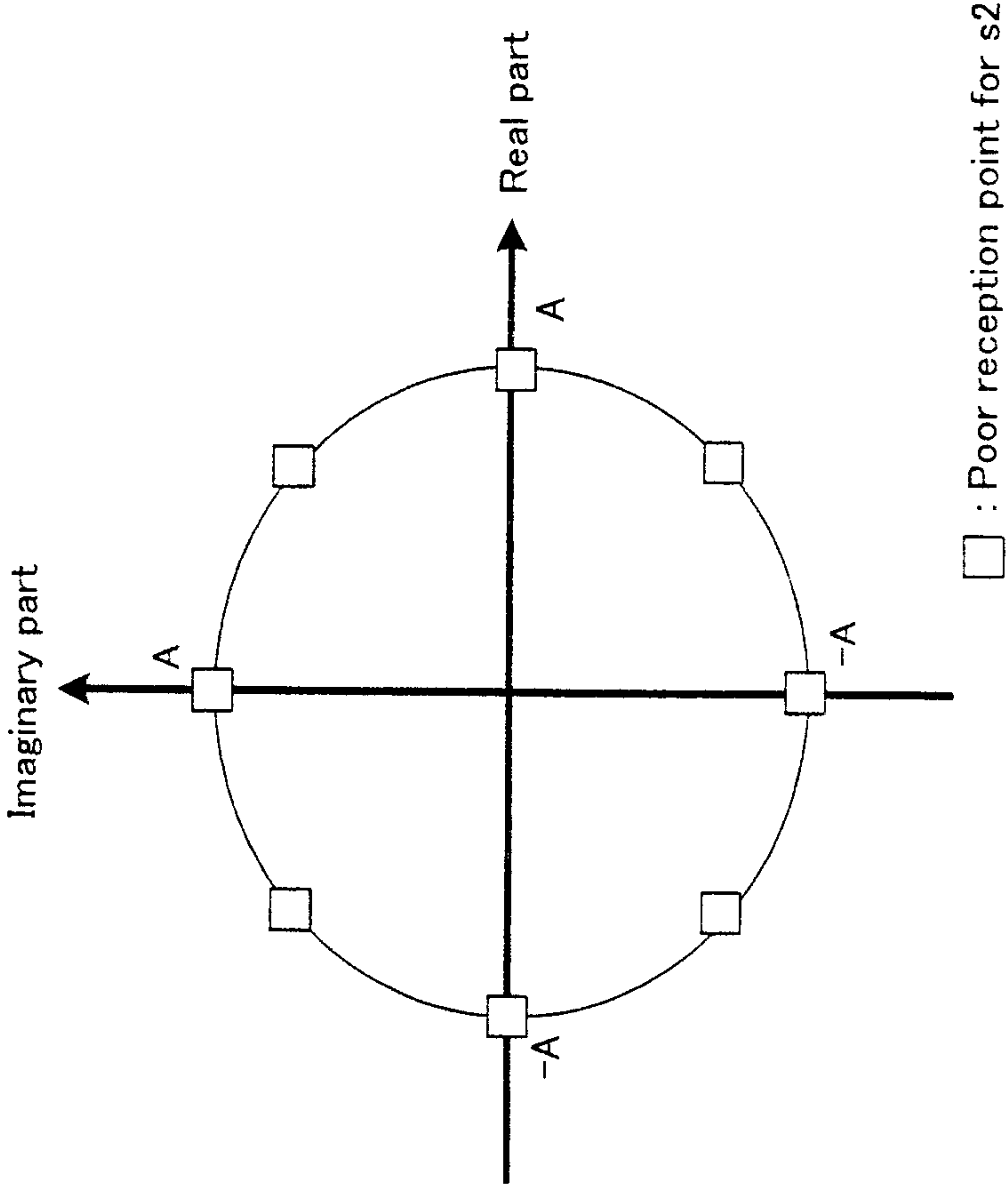


FIG. 32

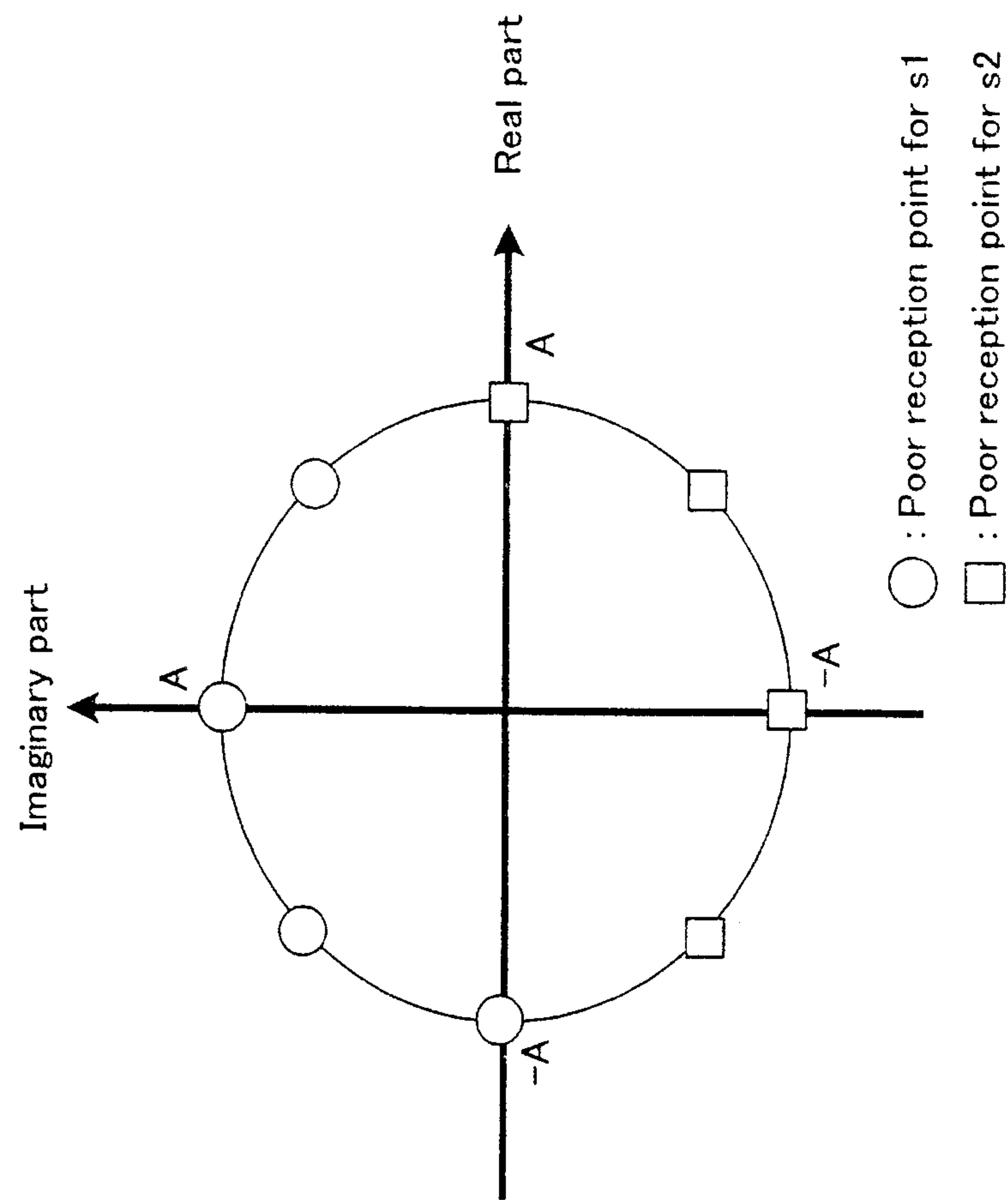
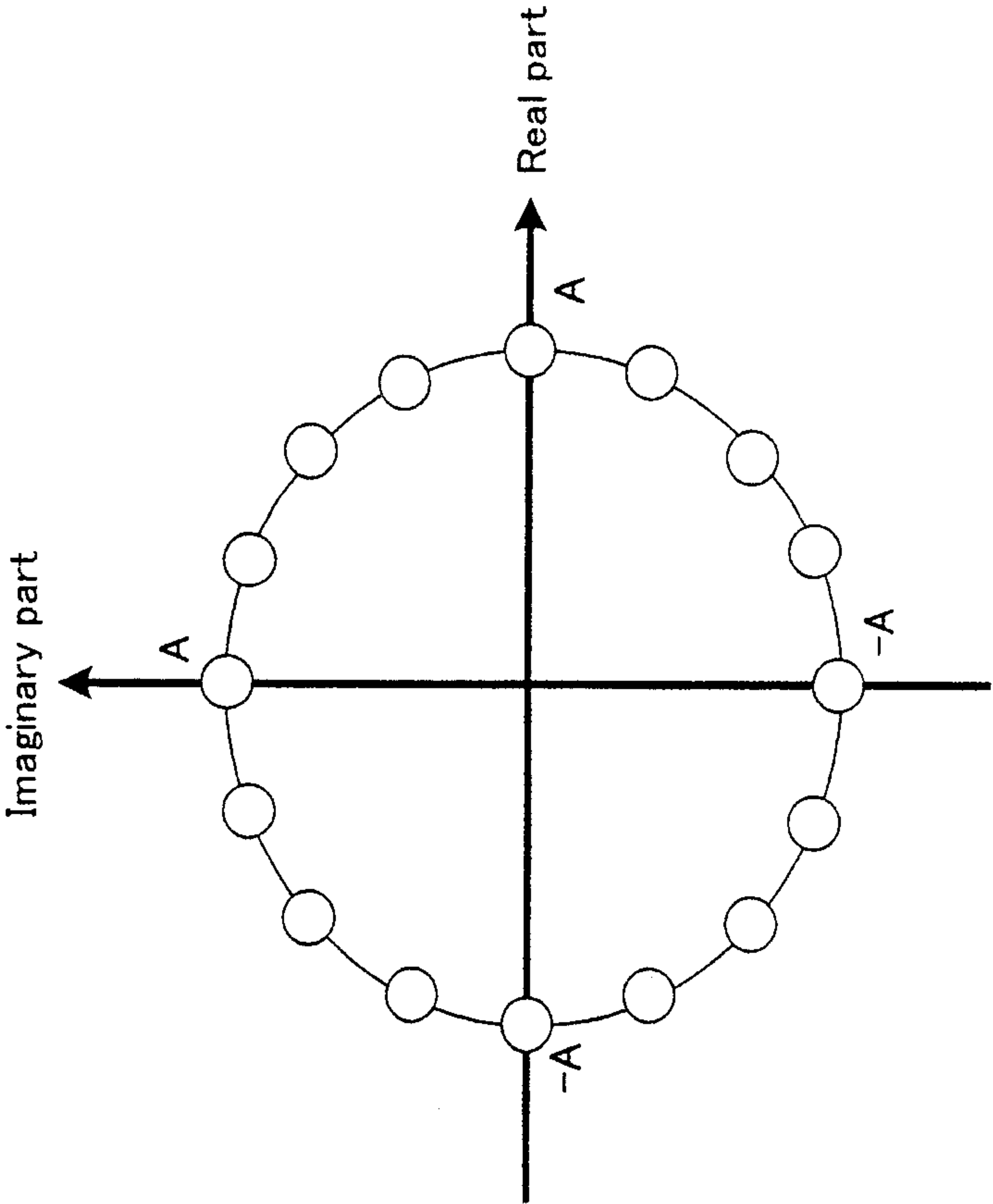
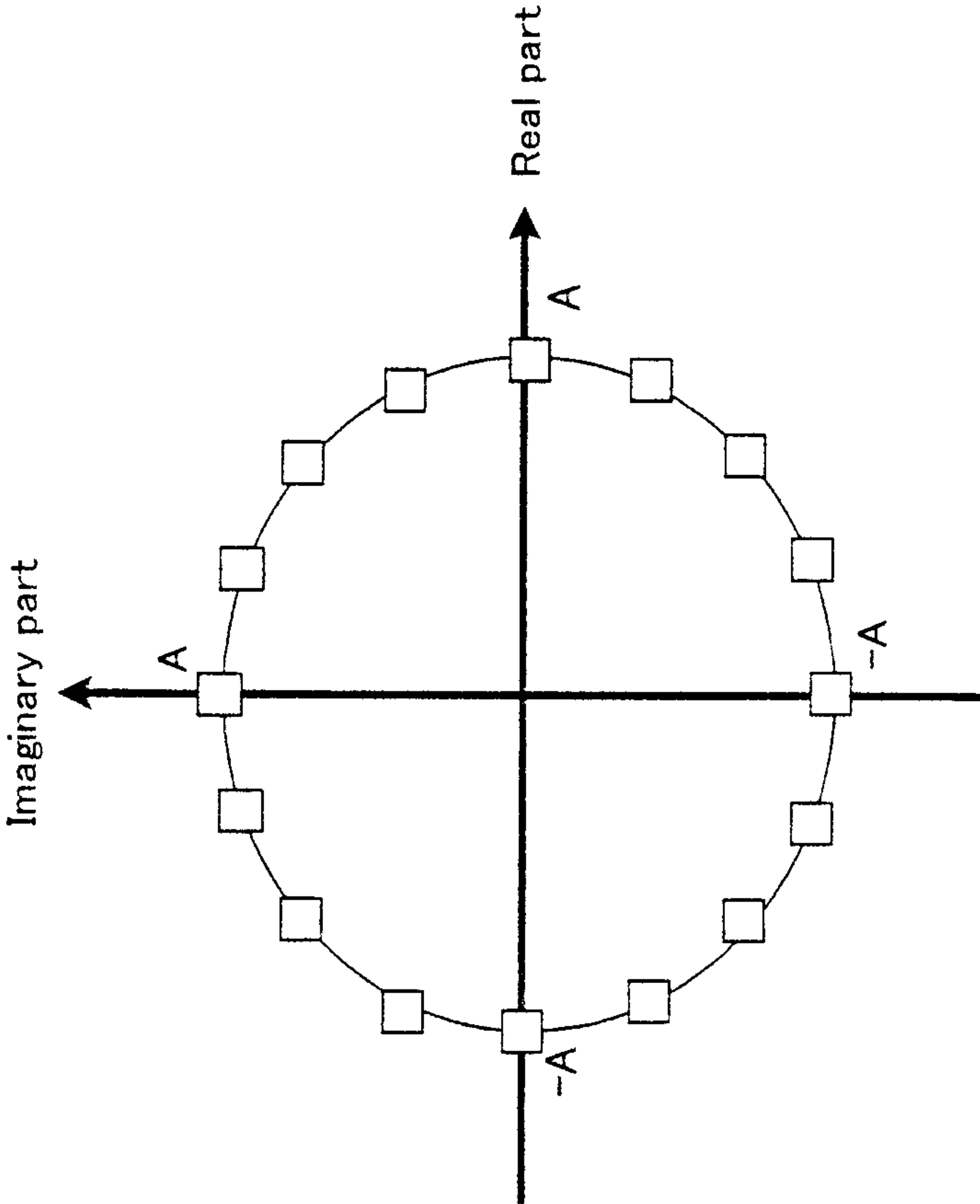


FIG. 33A



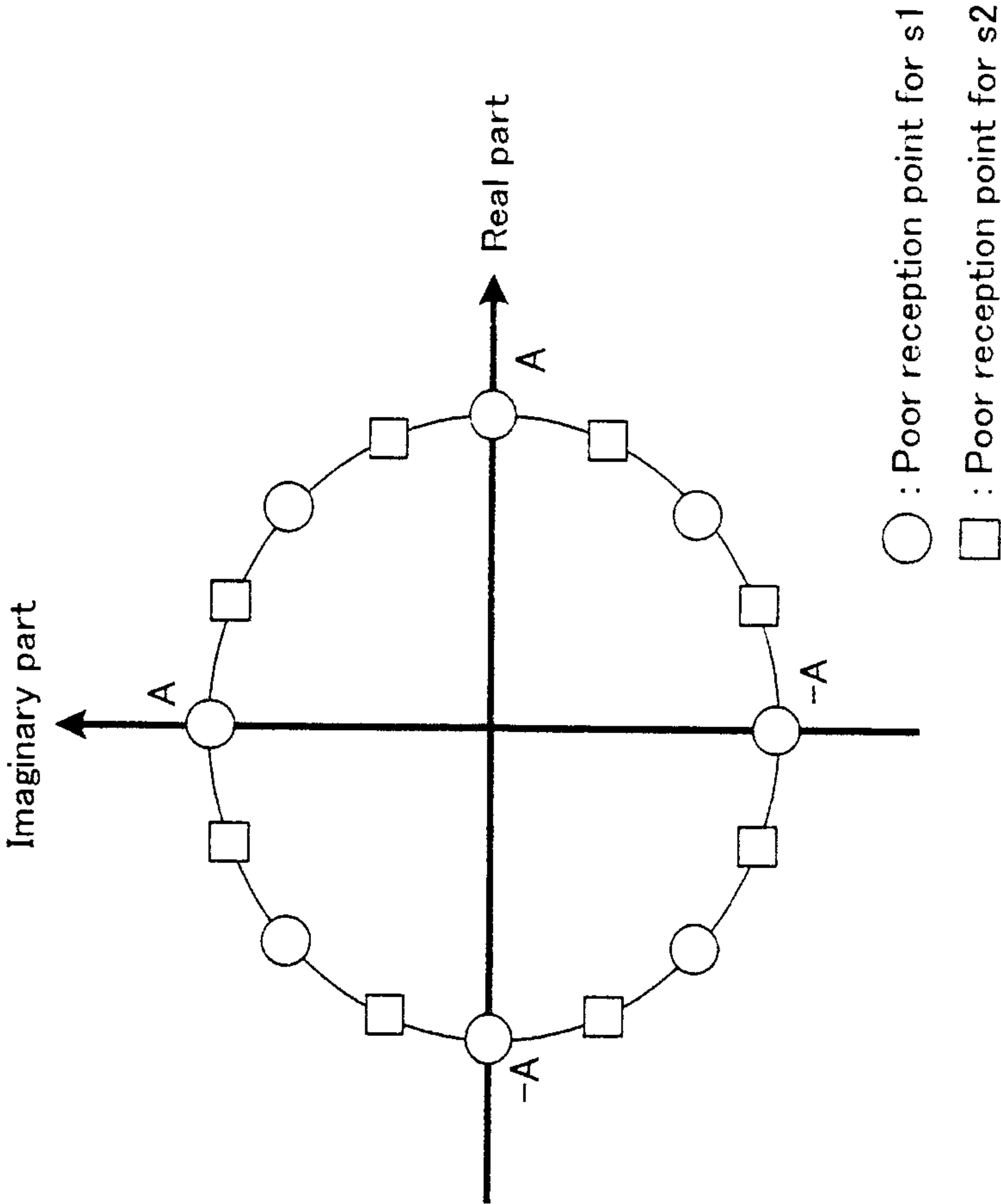
○ : Poor reception point for s1

FIG. 33B



□ : Poor reception point for s2

FIG. 34



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FIG. 35A

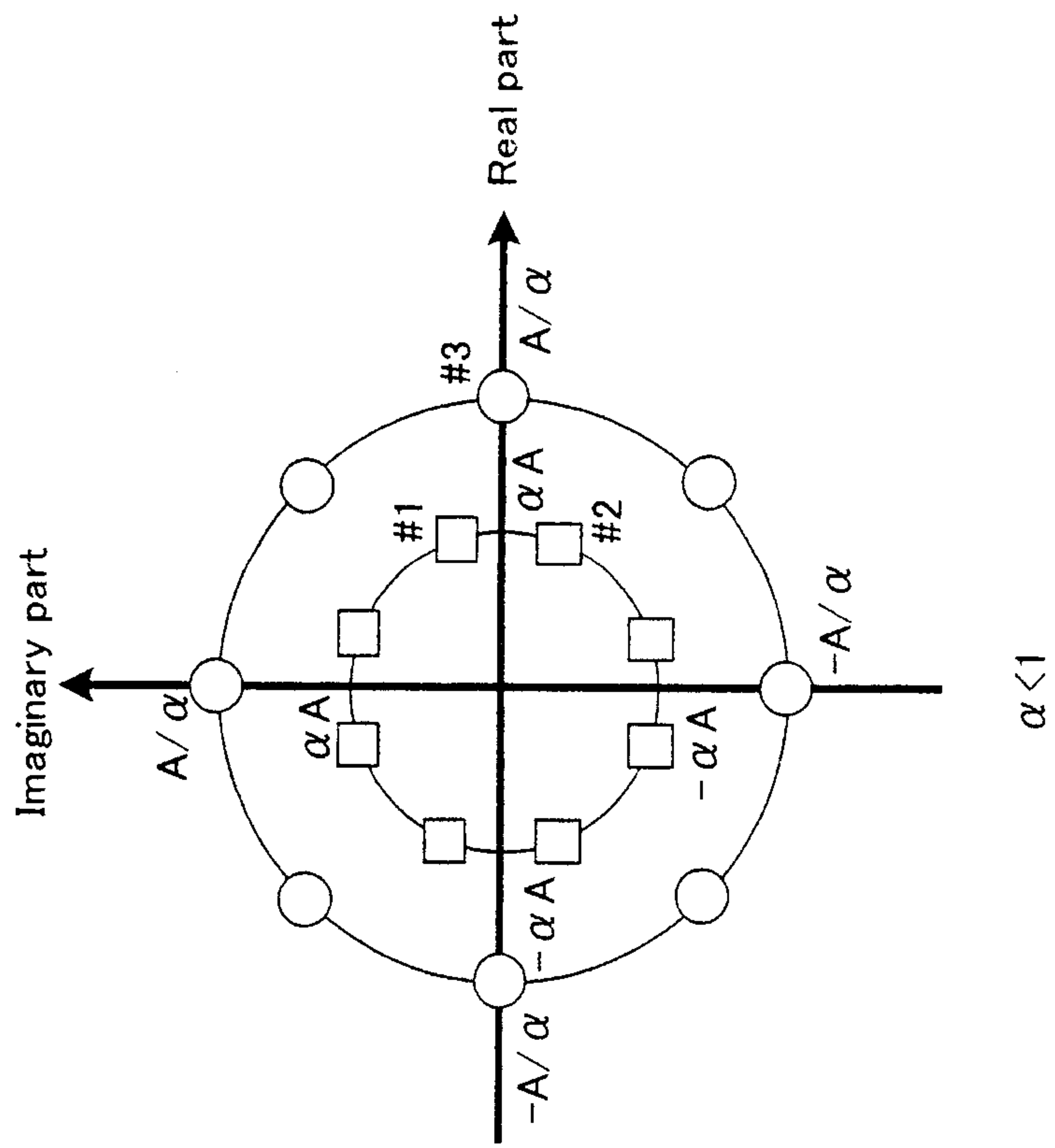
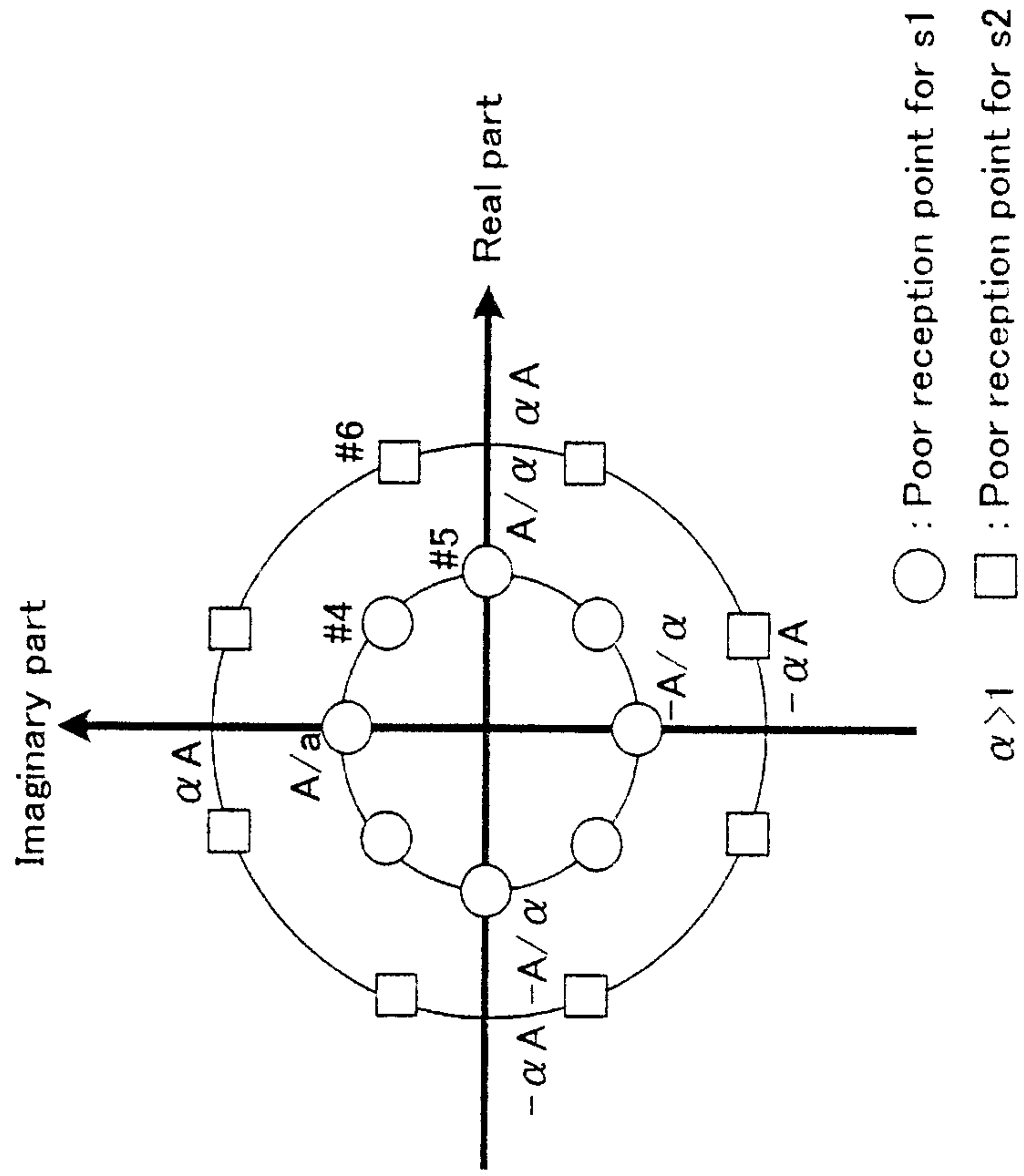
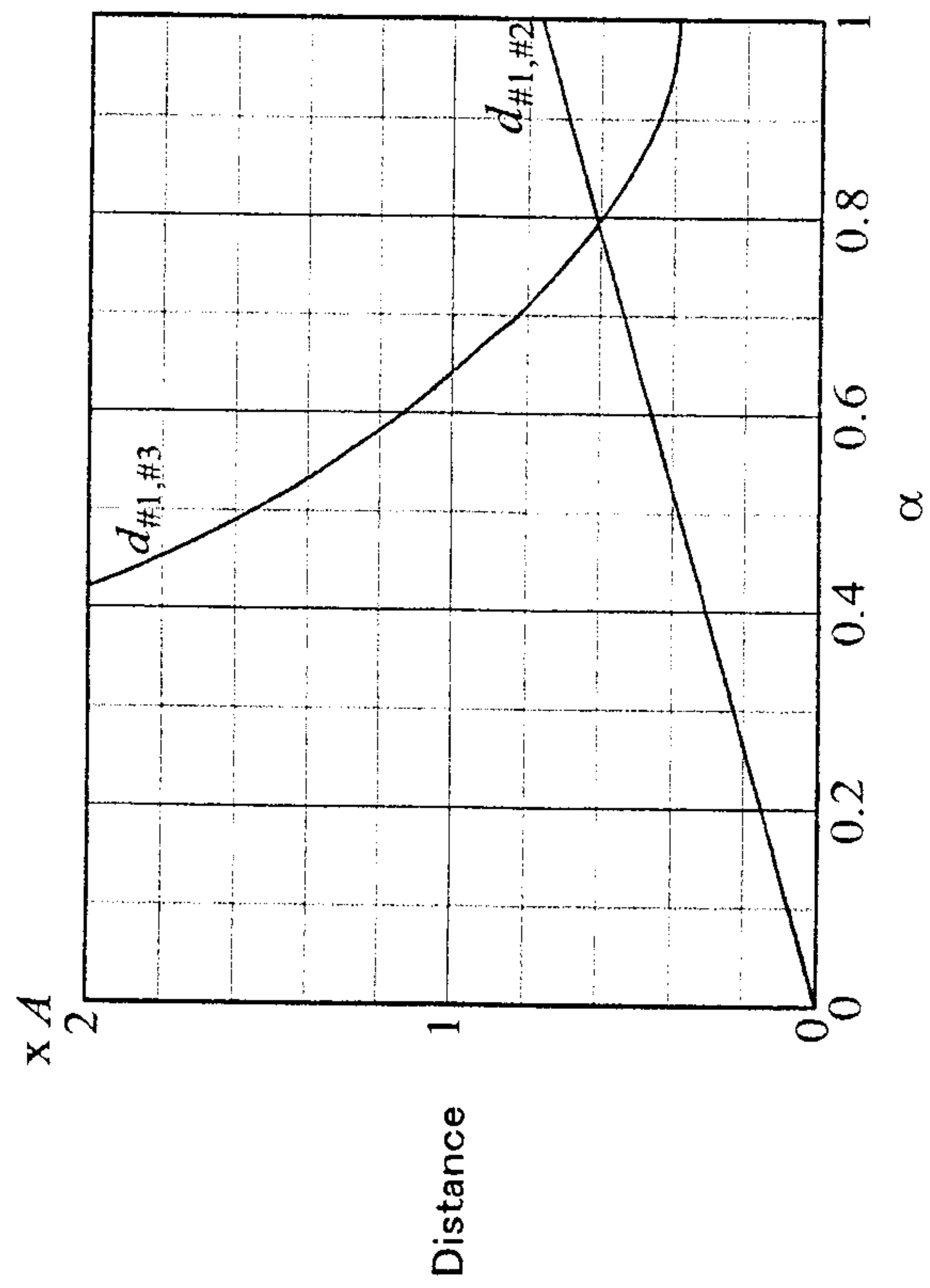


FIG. 35B



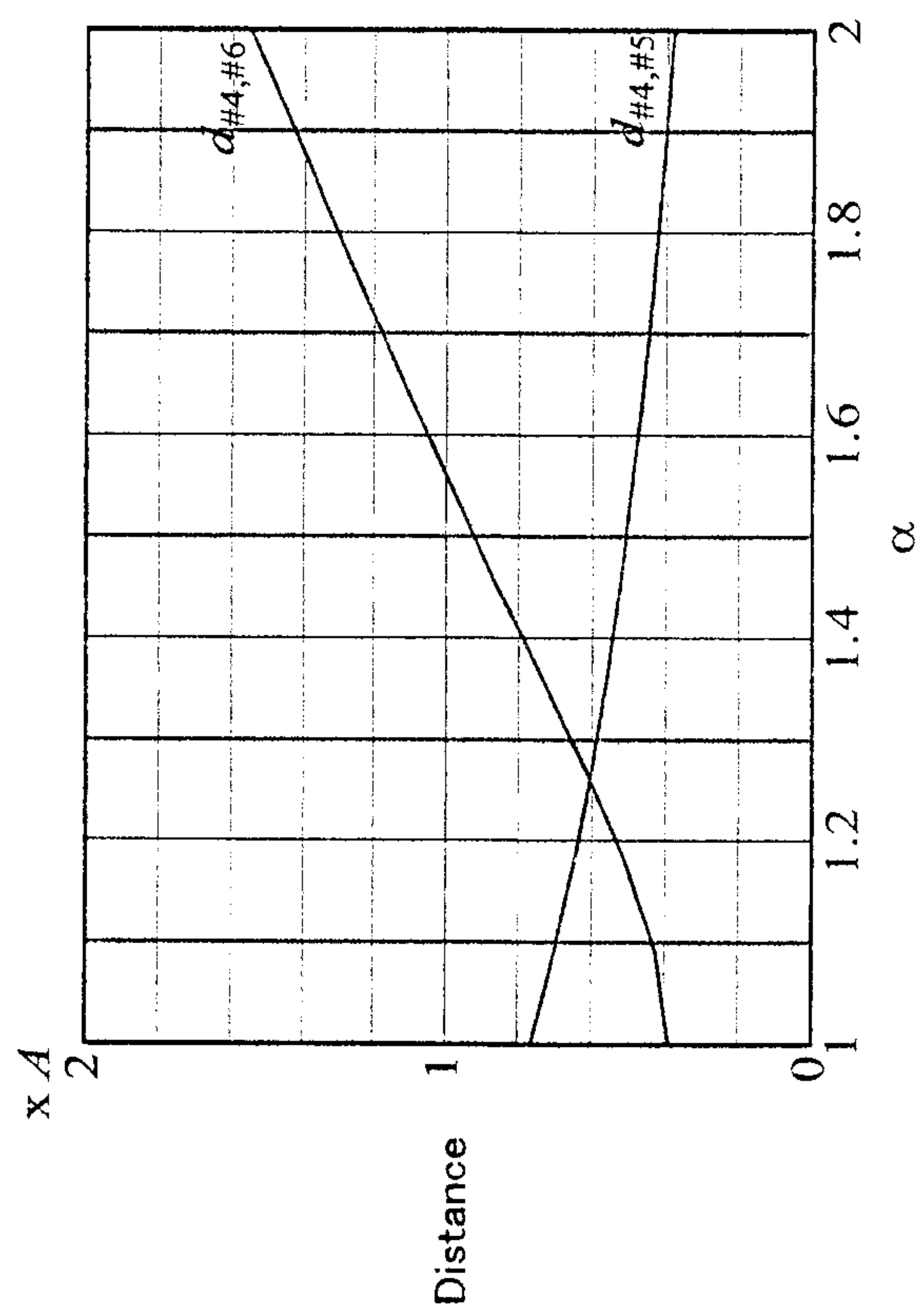
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FIG. 36



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FIG. 37



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FIG. 38A

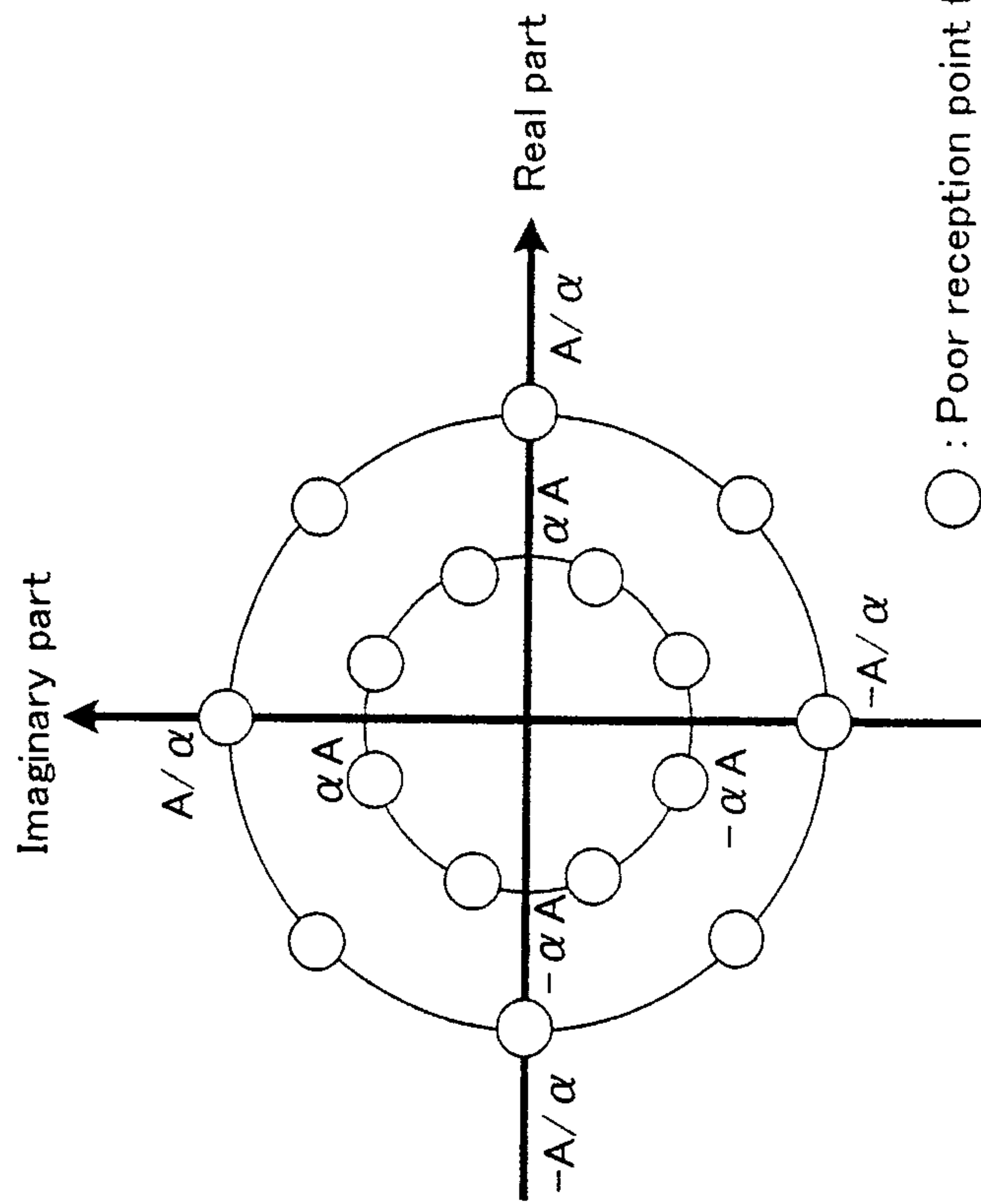
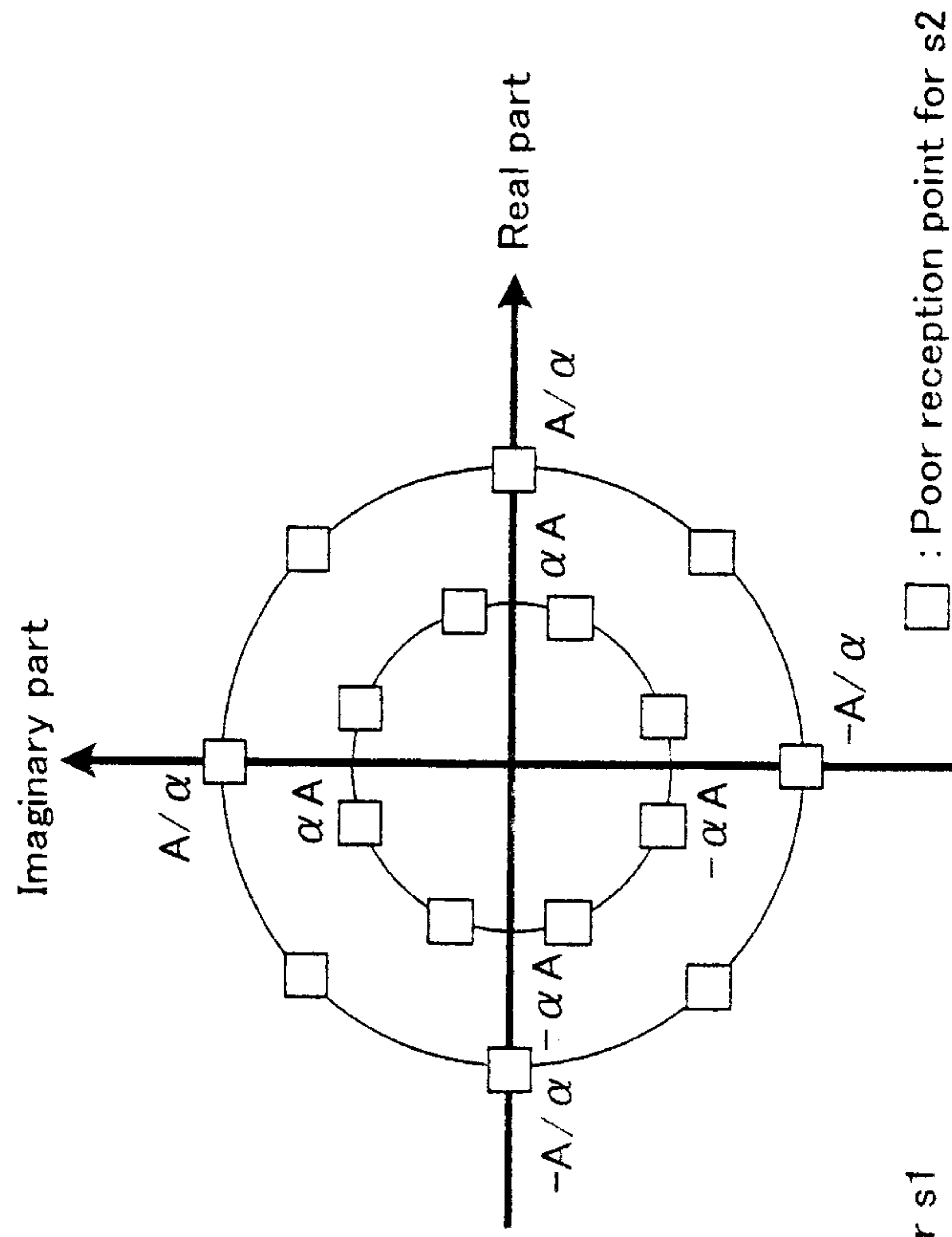


FIG. 38B



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FIG. 39A

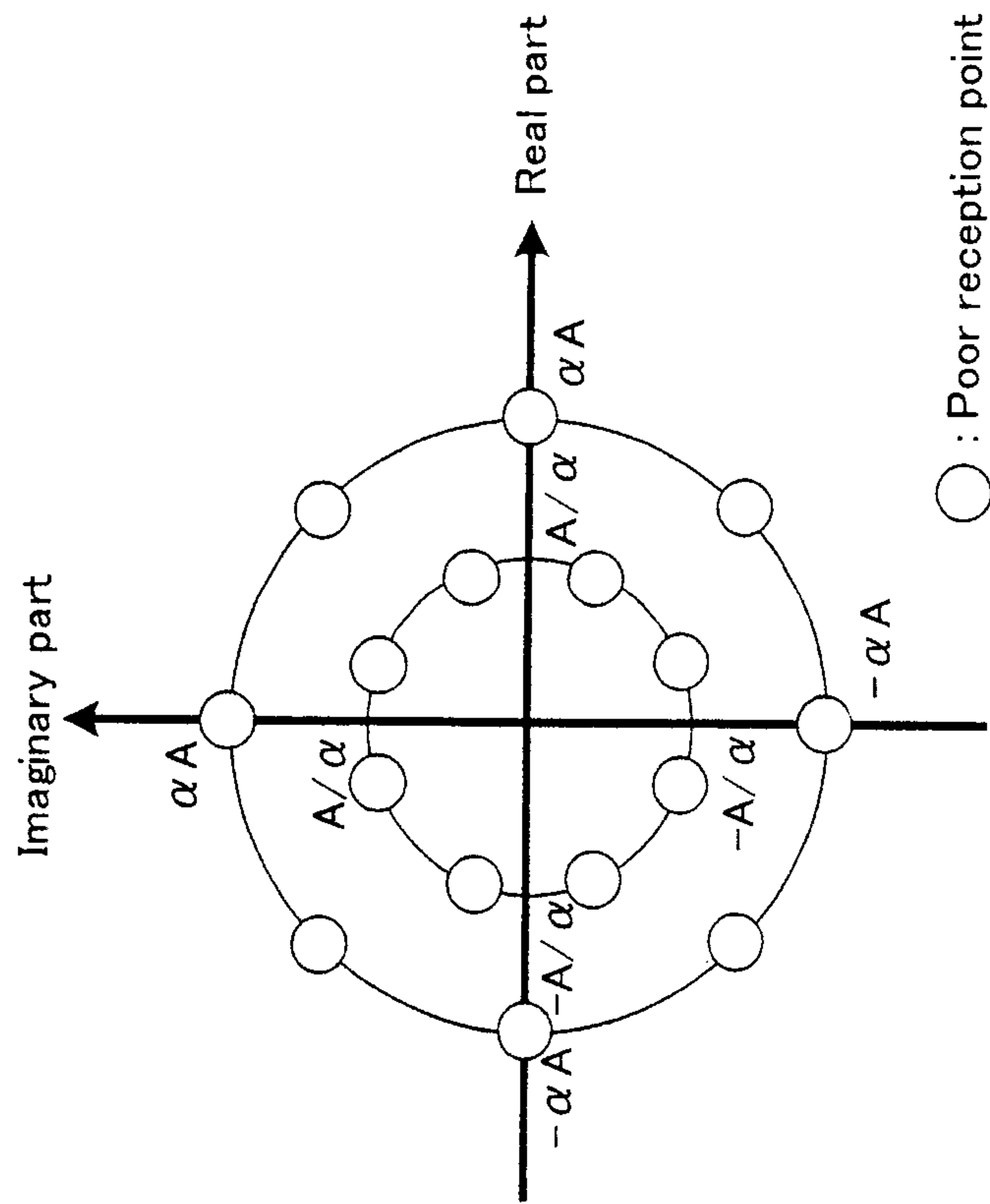
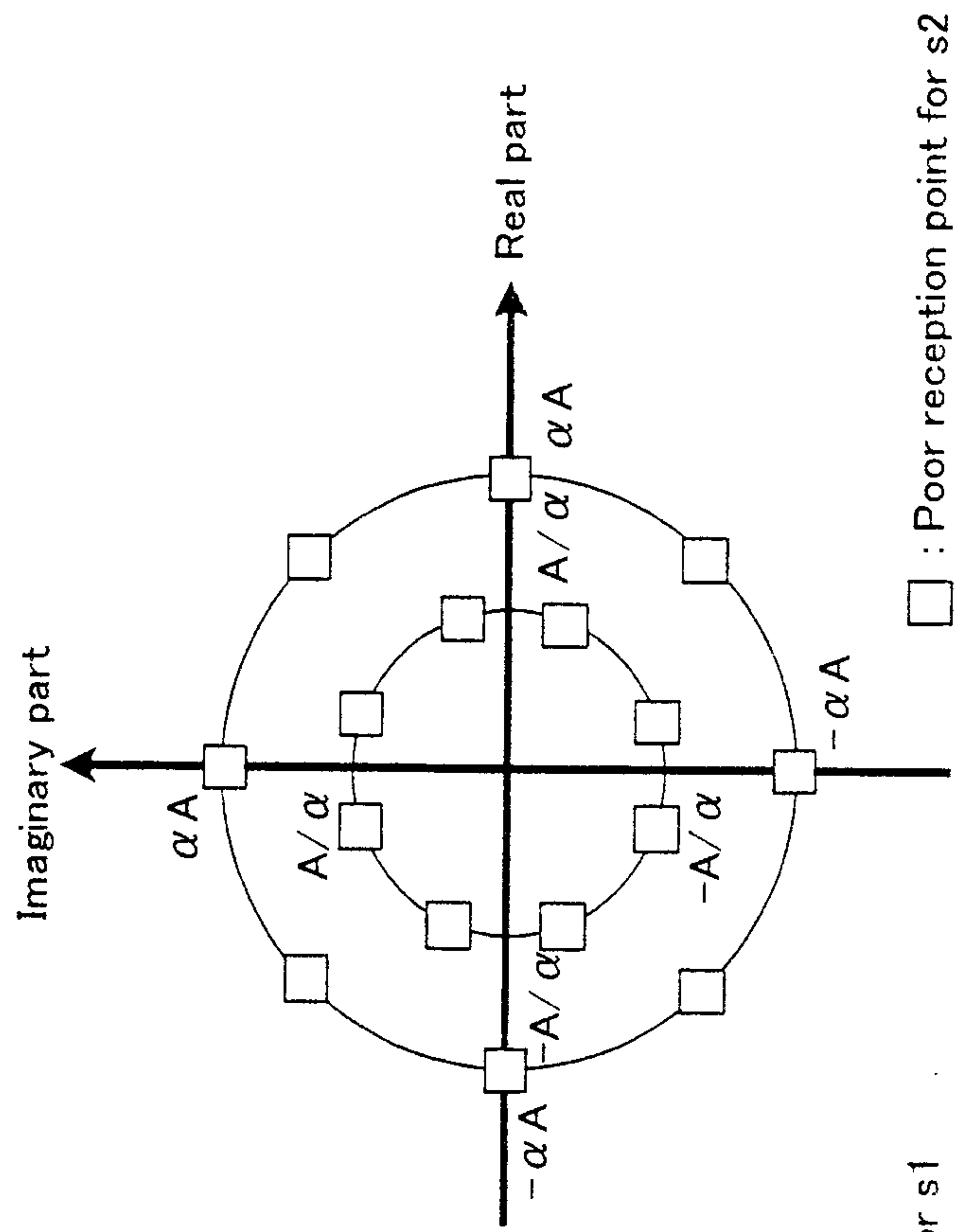


FIG. 39B



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FIG. 40

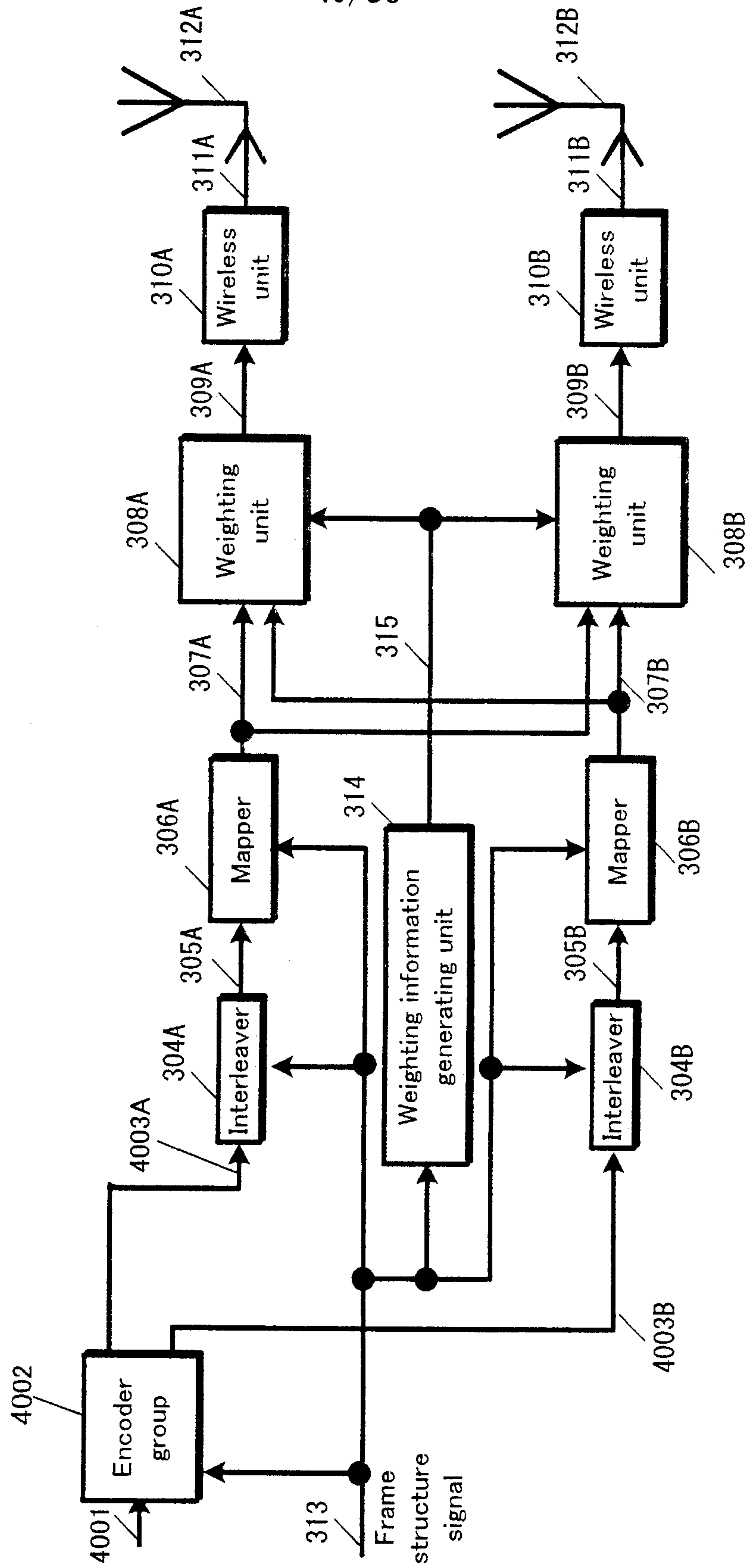
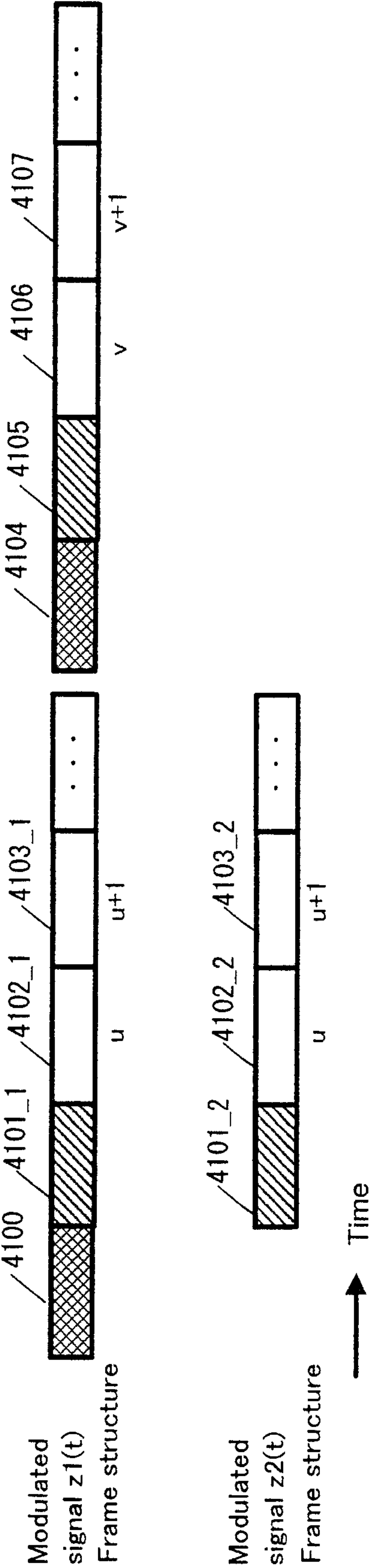


FIG. 41



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FIG. 42A

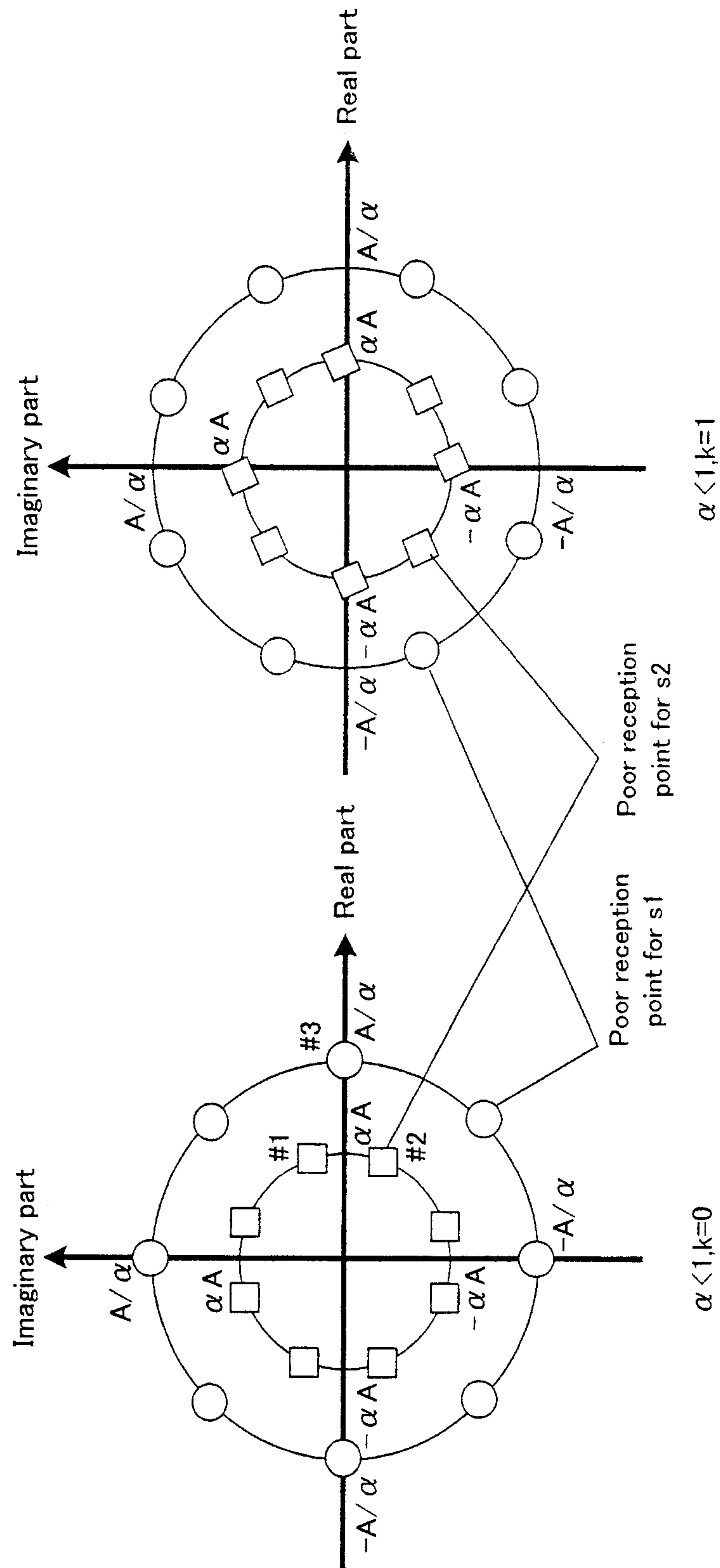


FIG. 42B

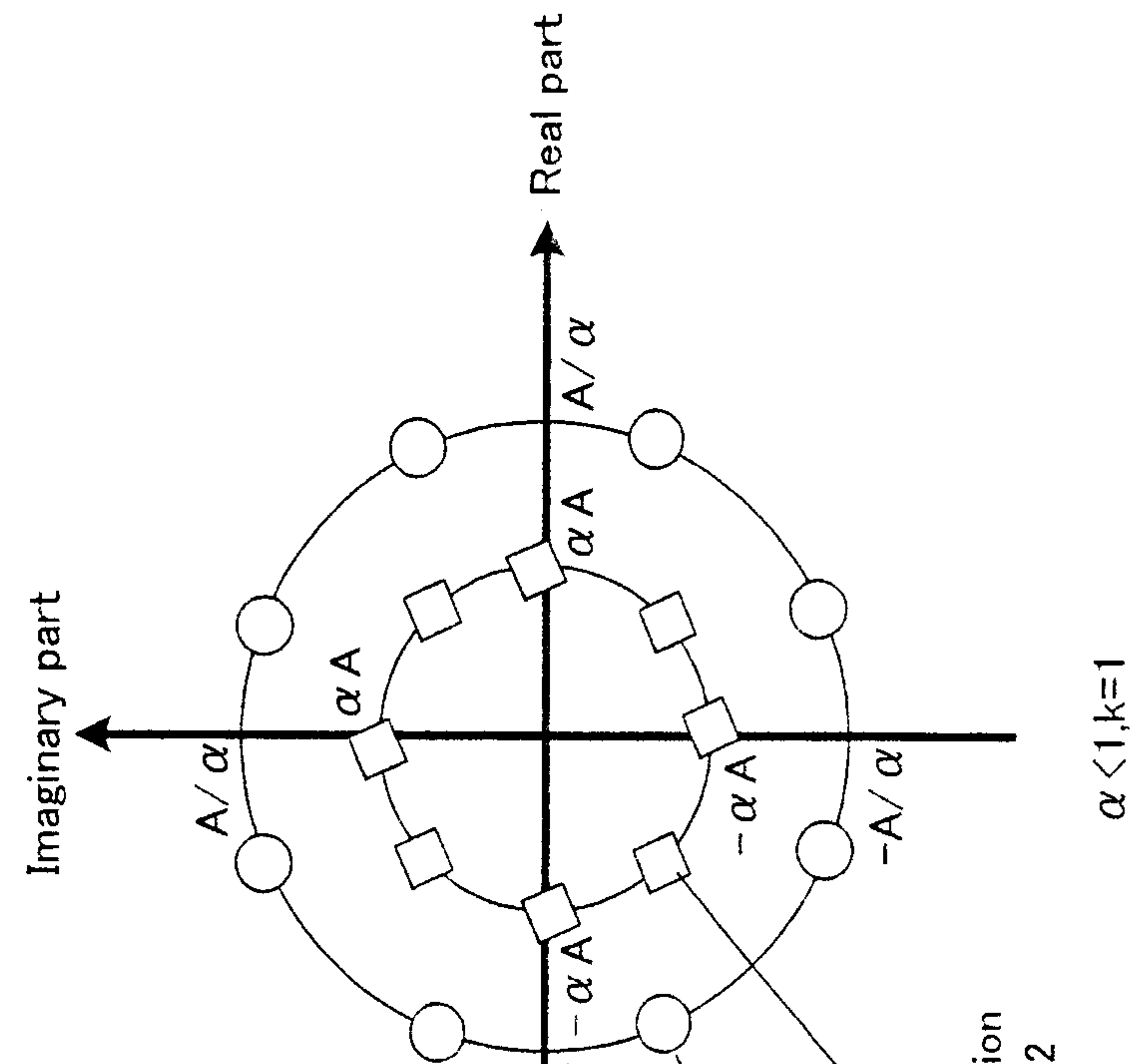


FIG. 43A

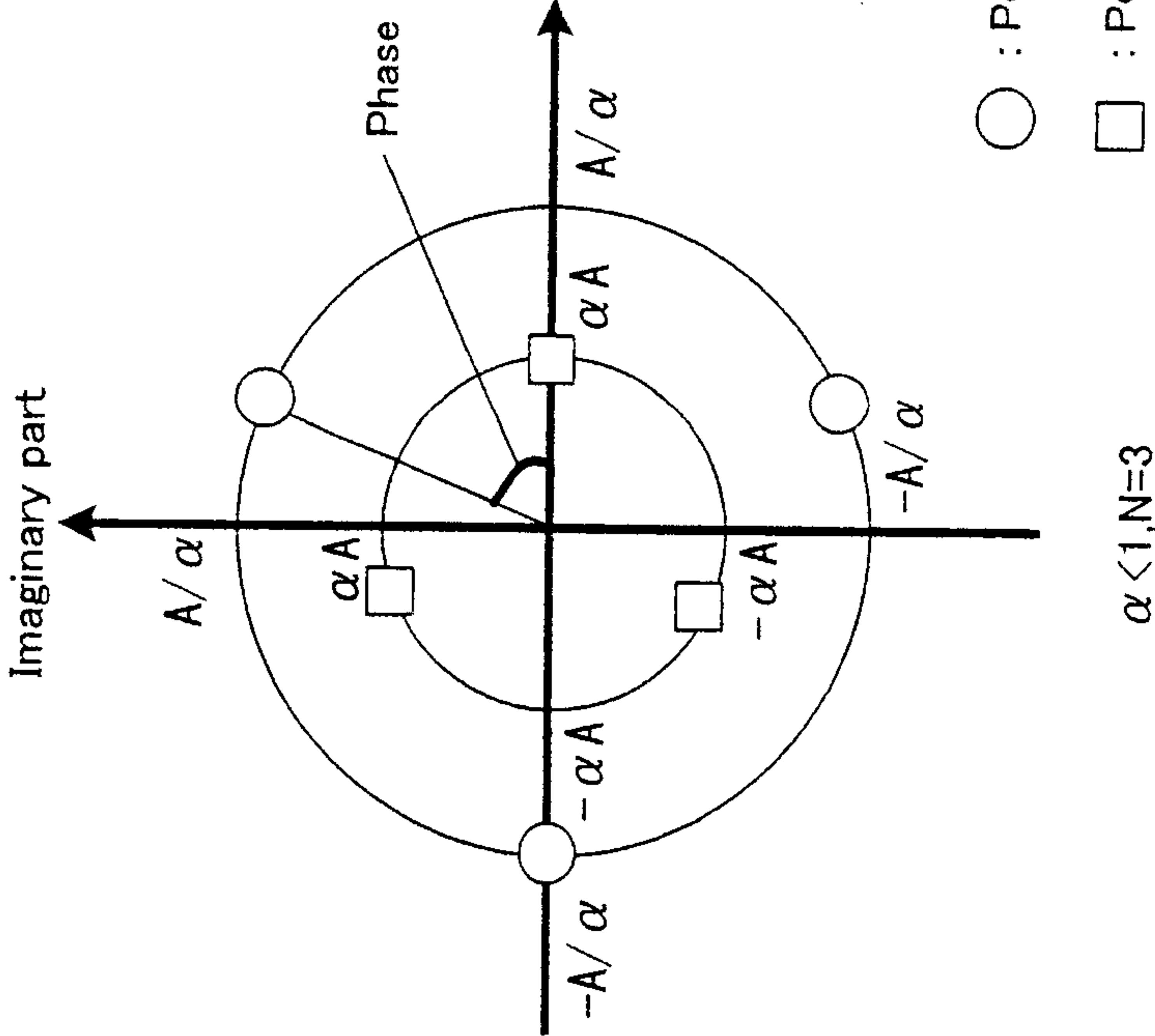


FIG. 43B

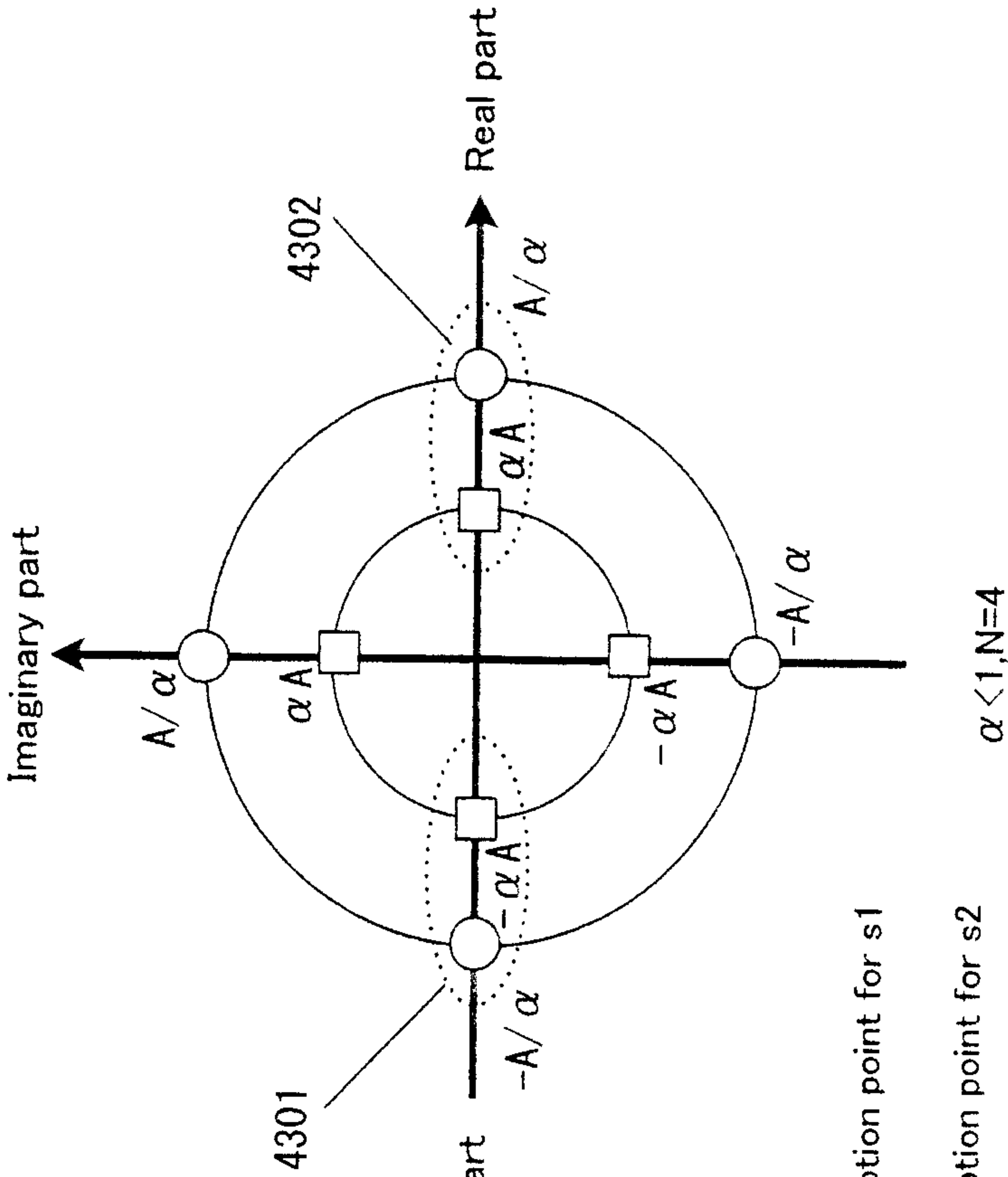


FIG. 44A

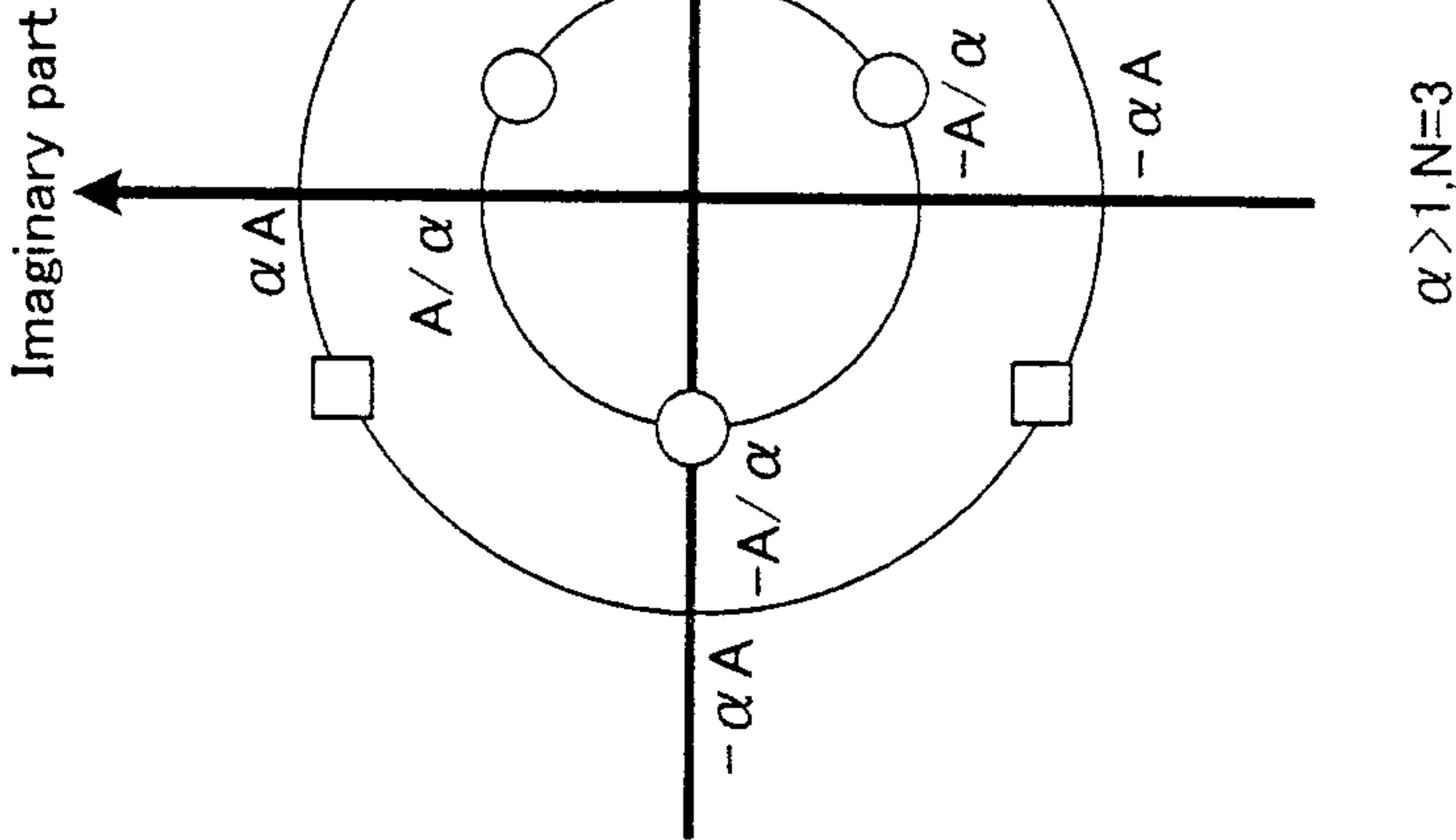


FIG. 44B

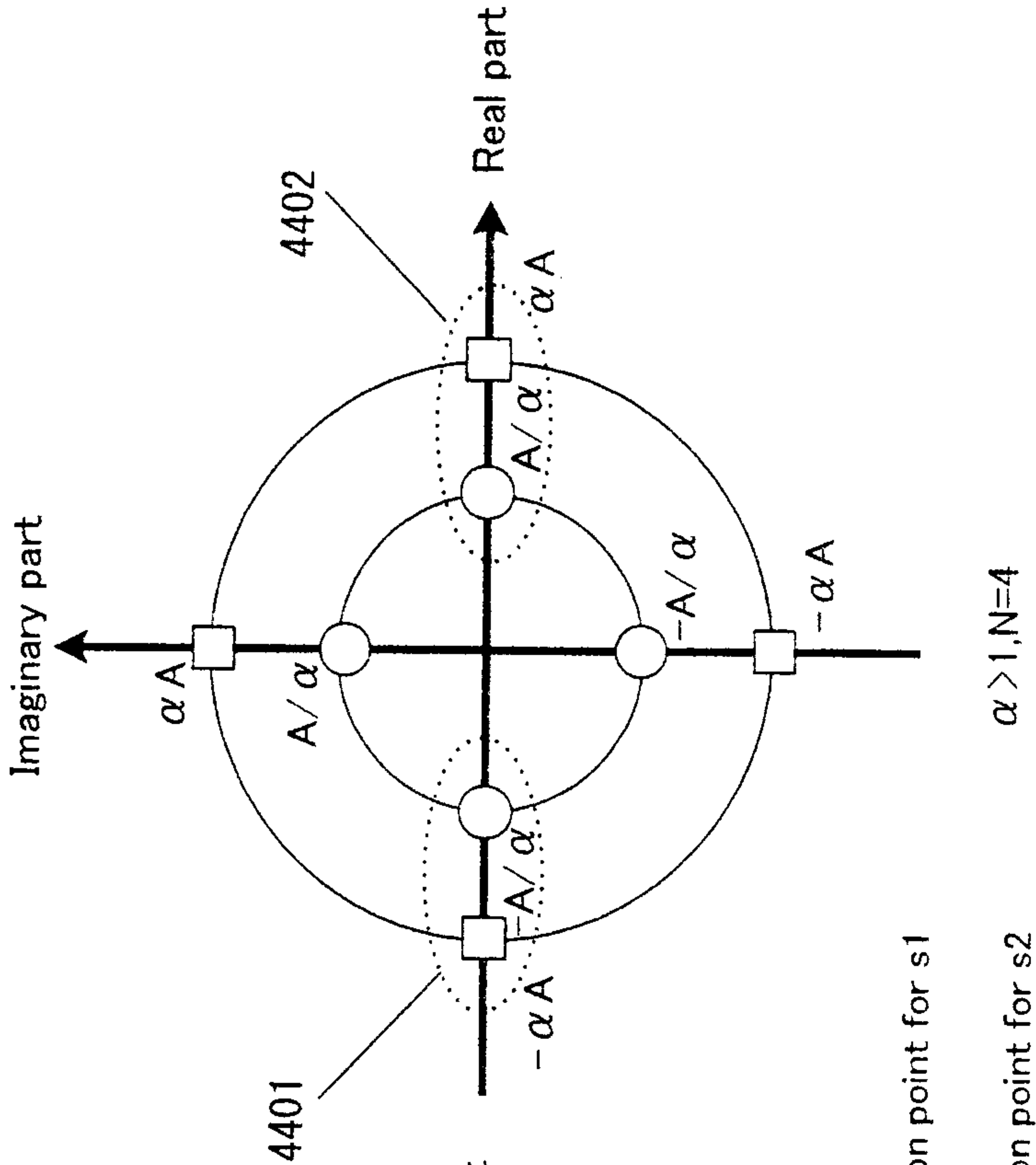


FIG. 45A

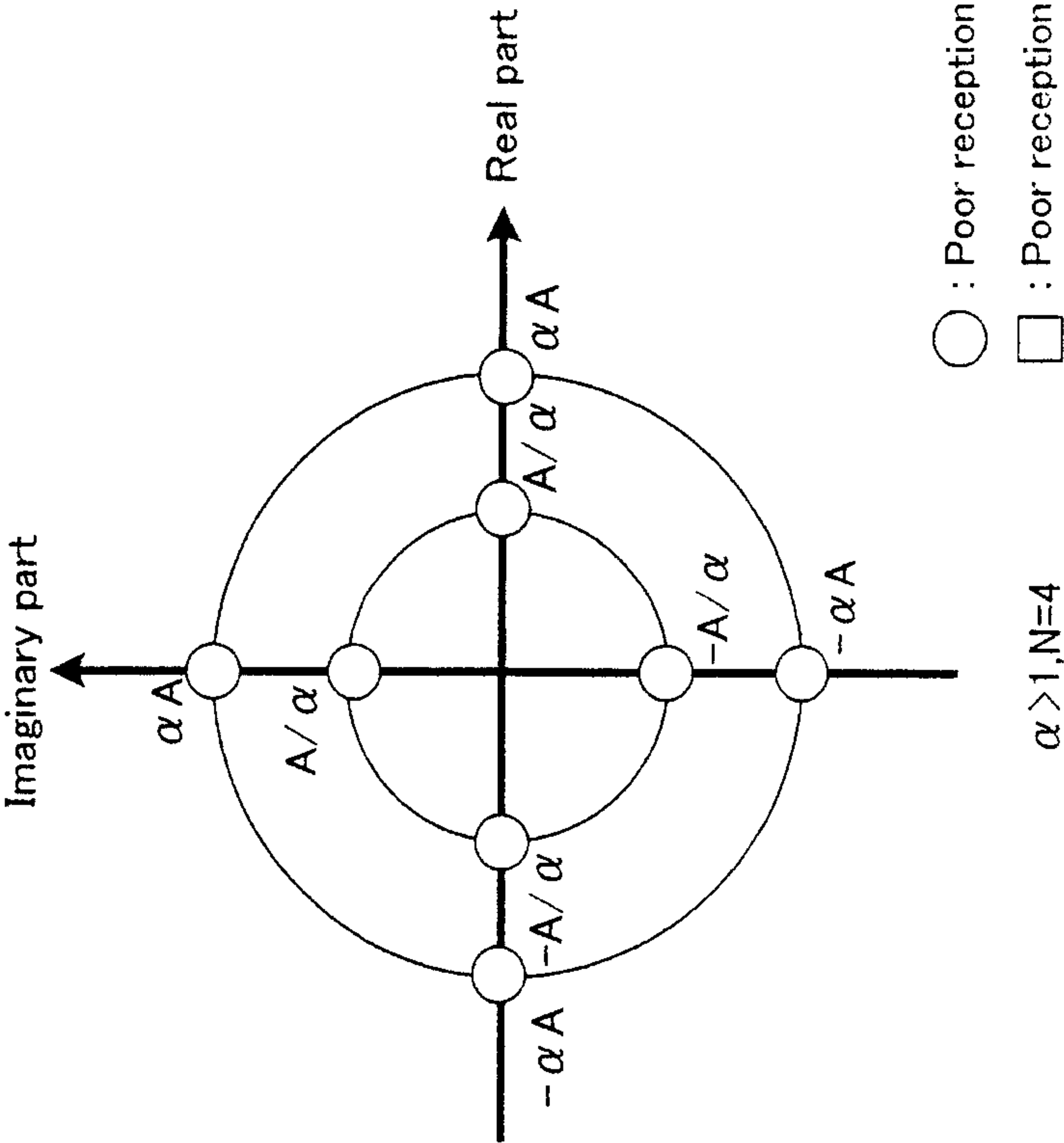


FIG. 45B

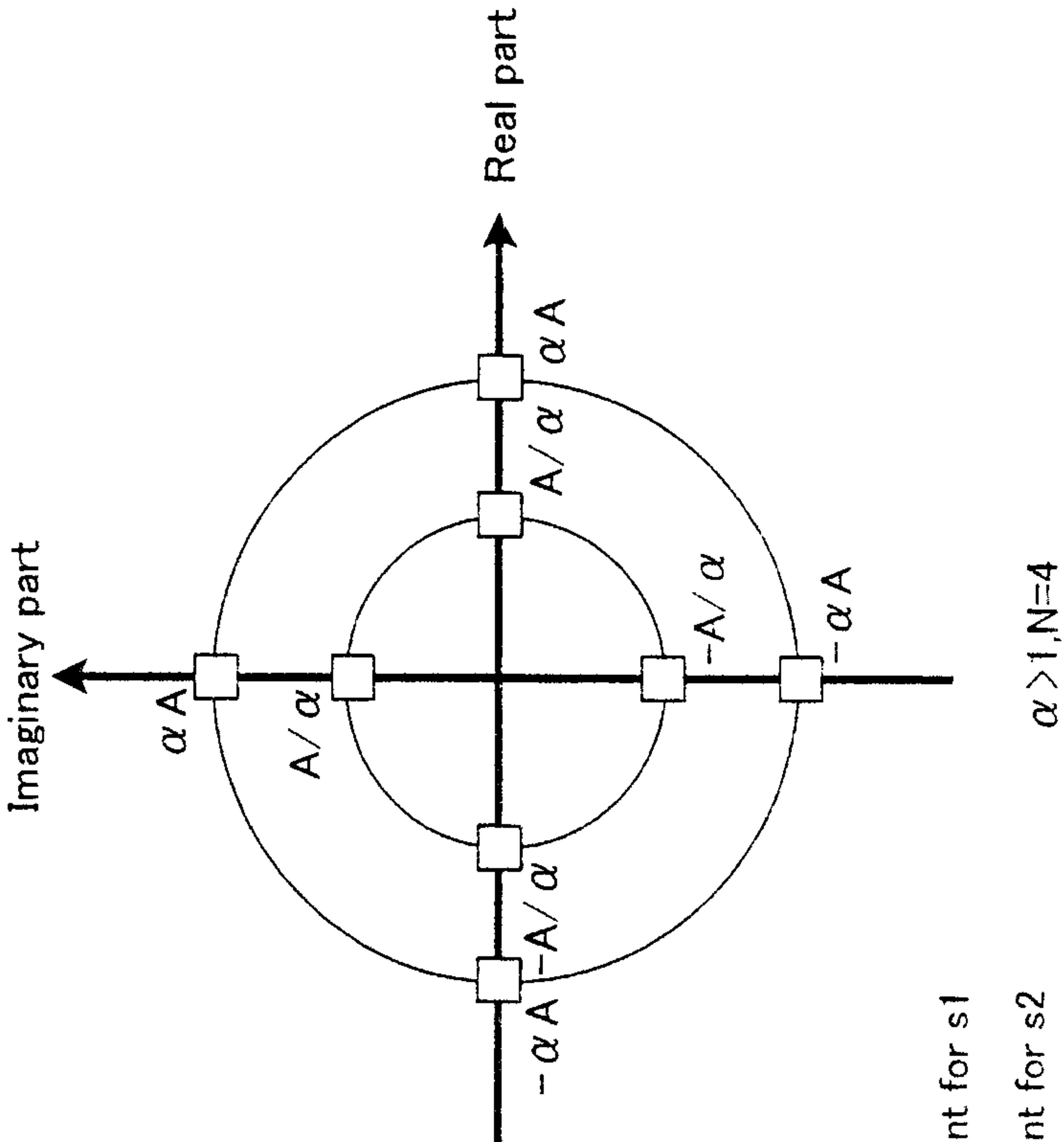


FIG. 46A

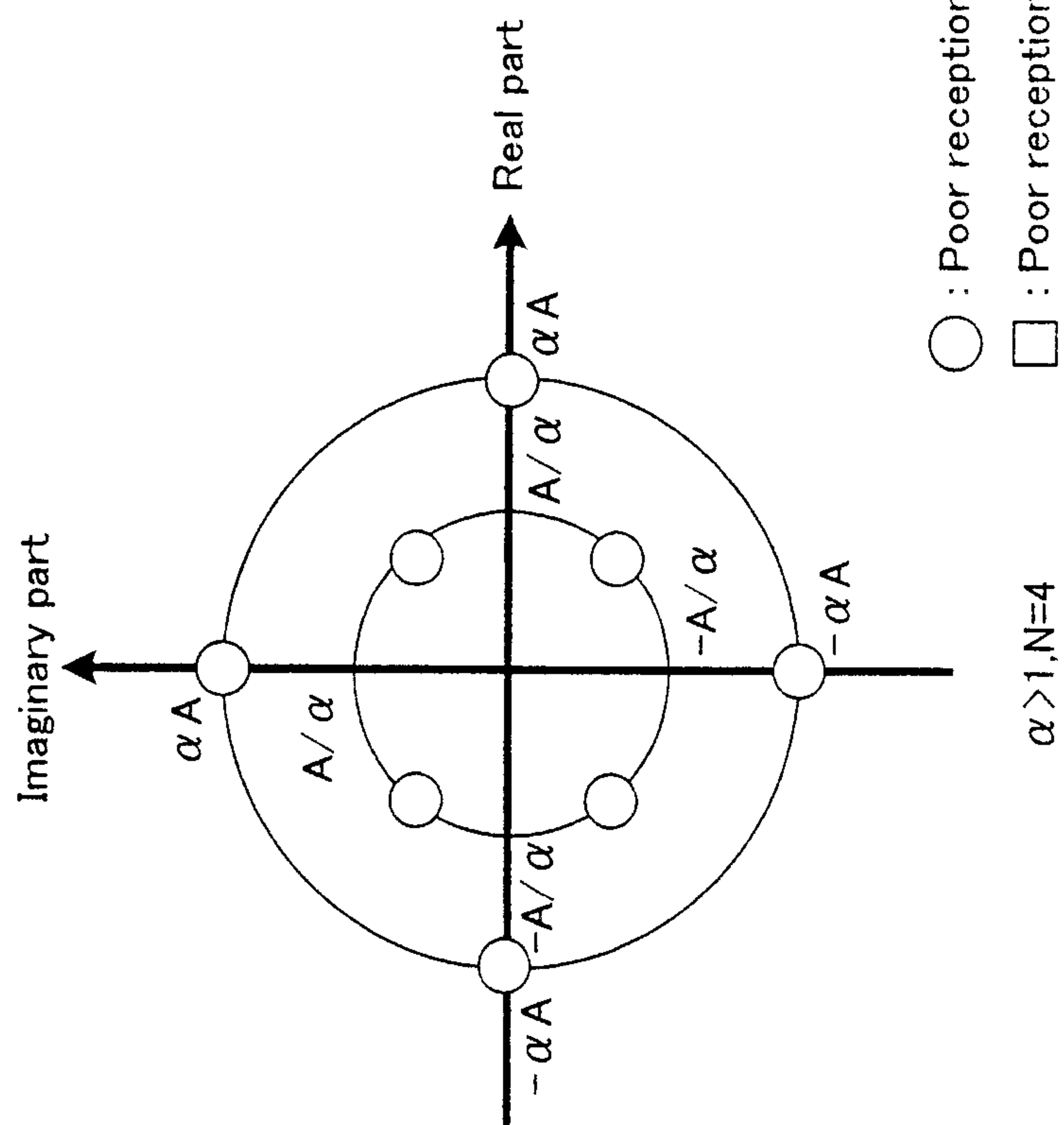


FIG. 46B

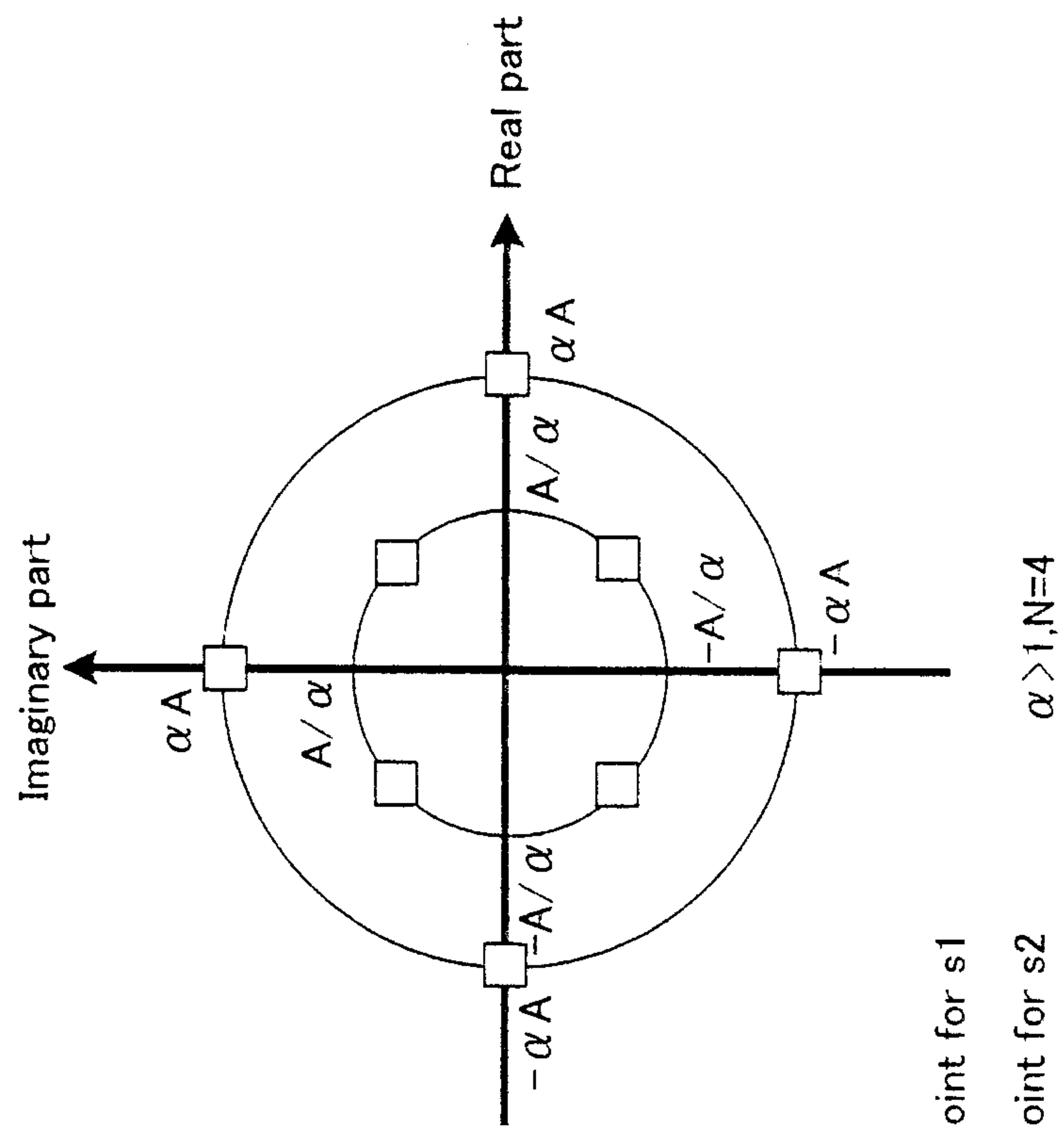


FIG. 47A

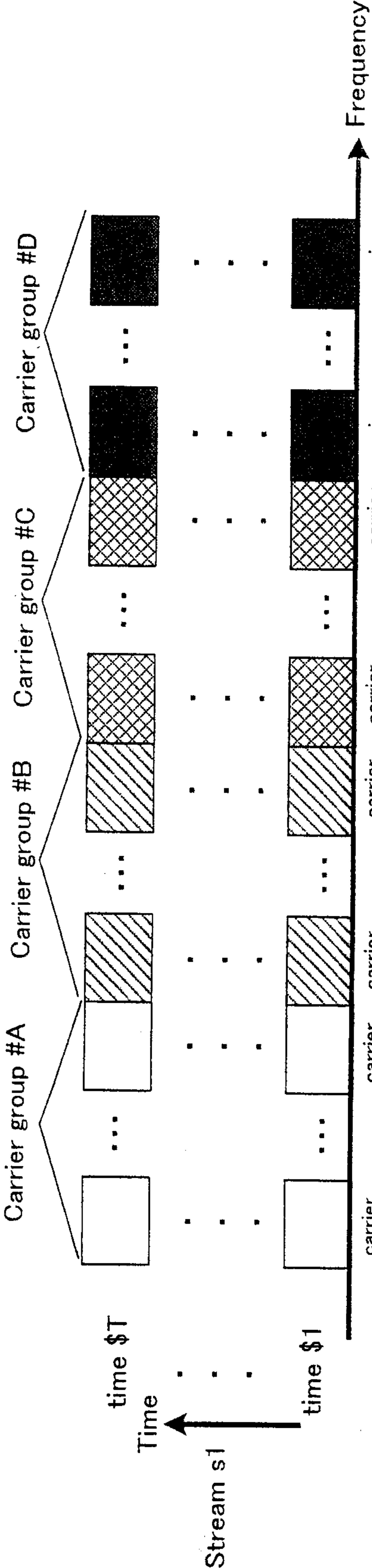
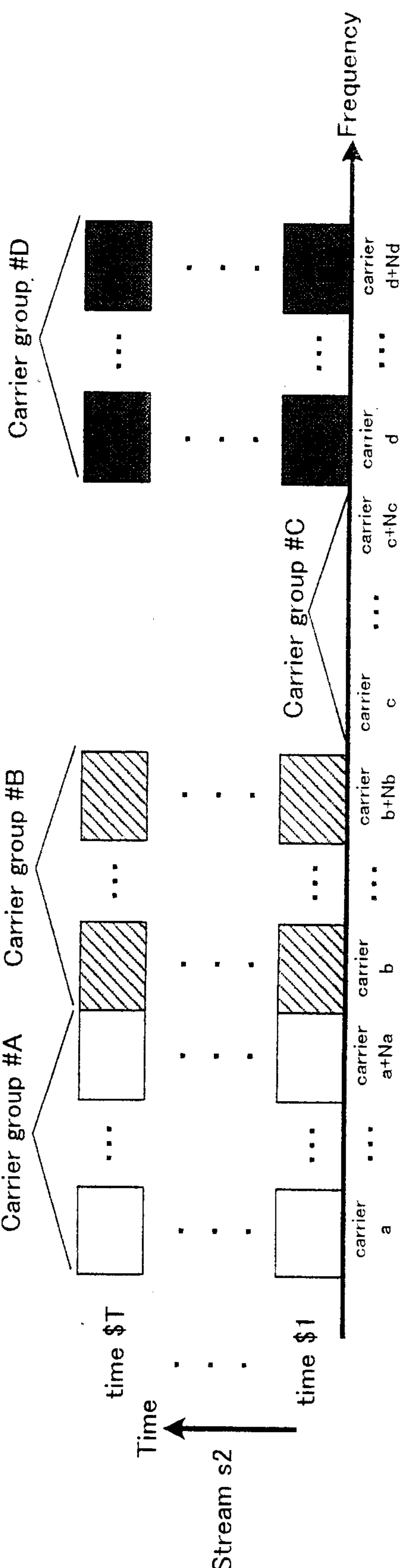


FIG. 47B



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FIG. 48A

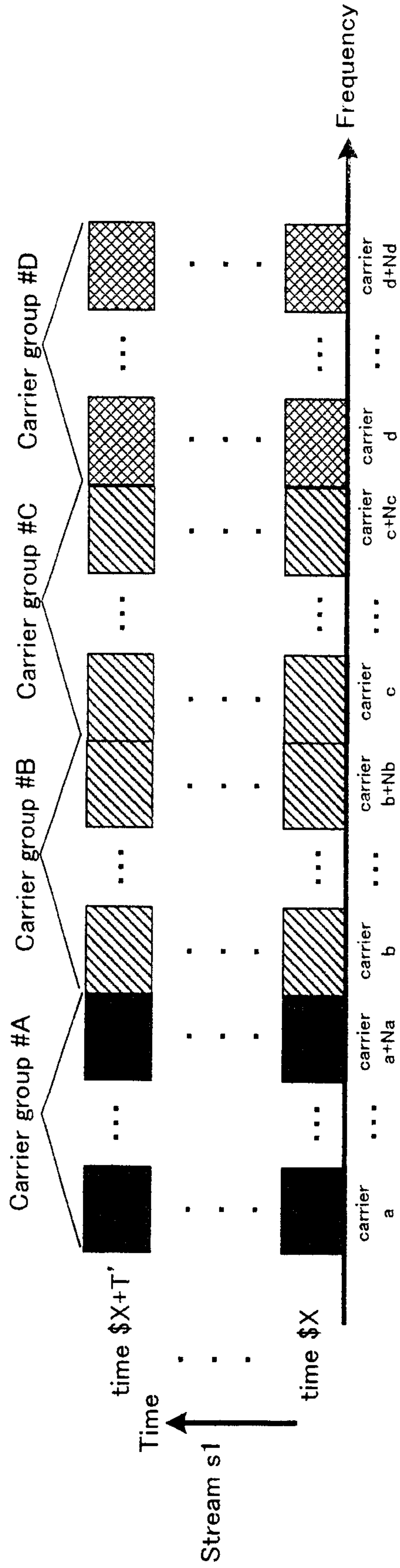


FIG. 48B

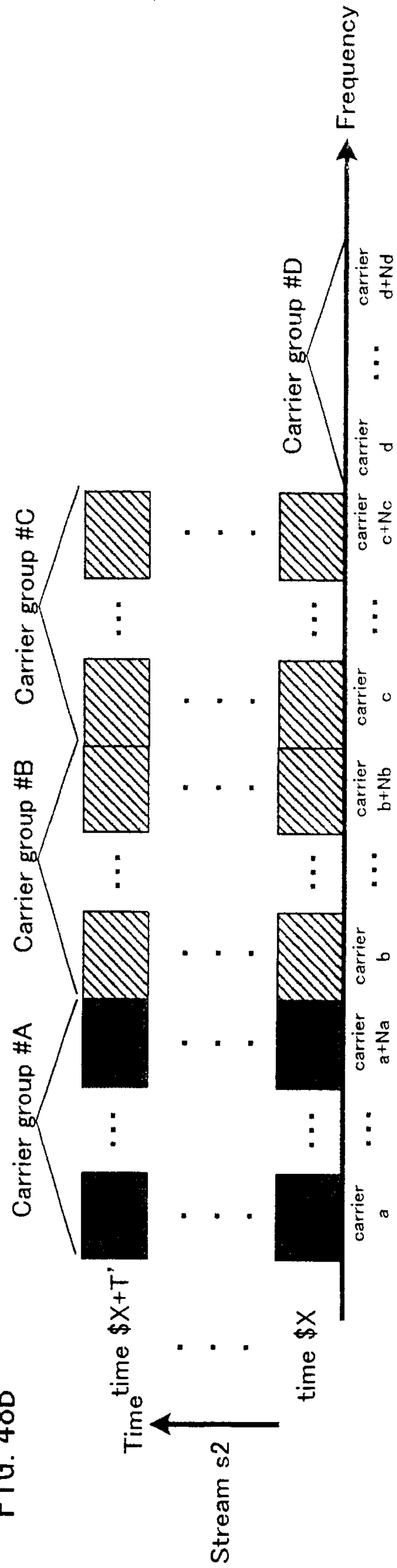
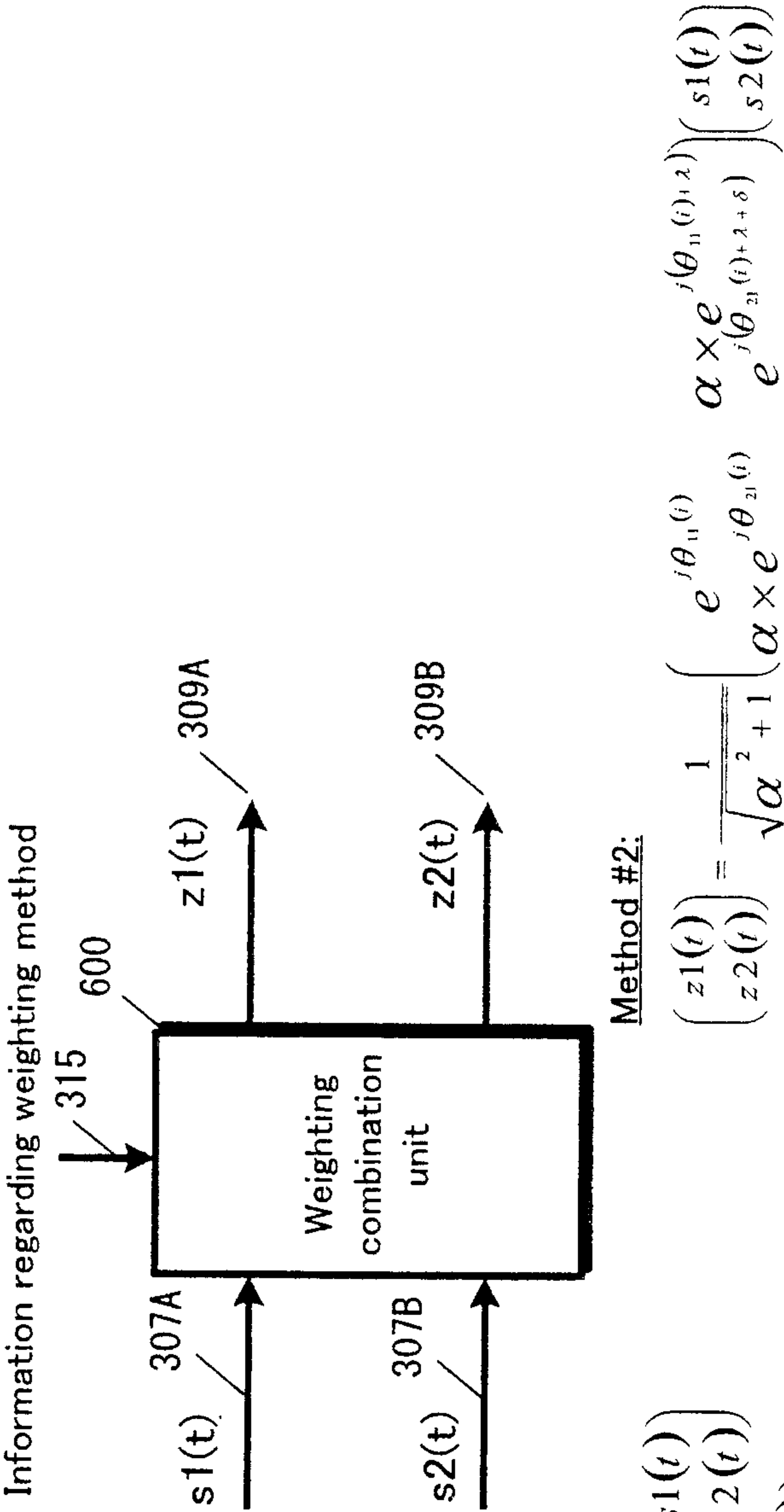


FIG. 49



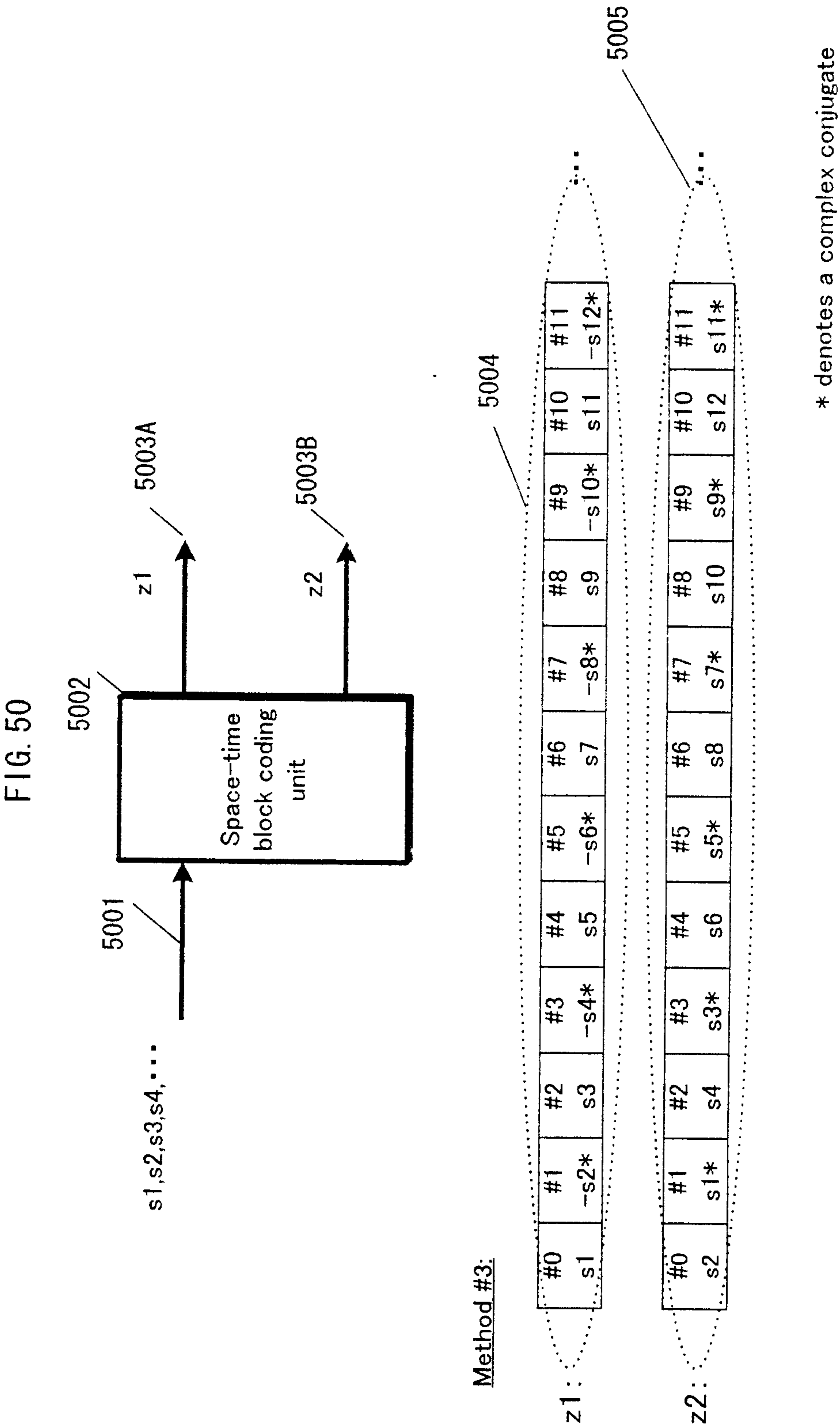
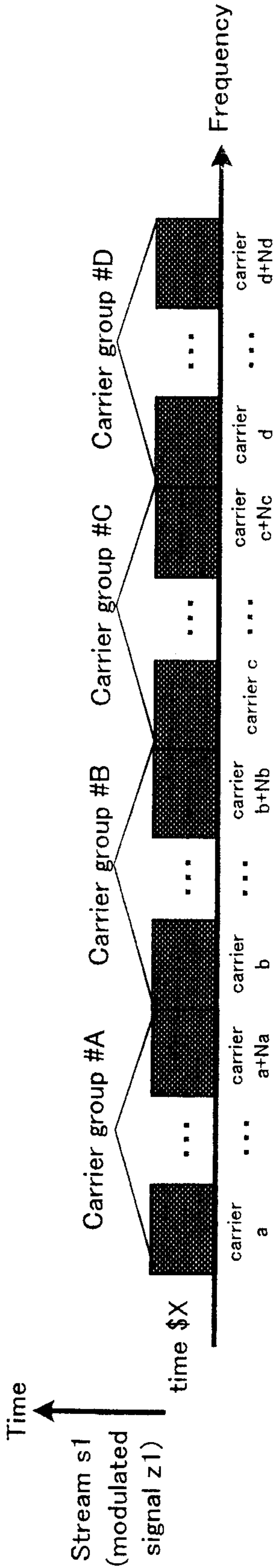
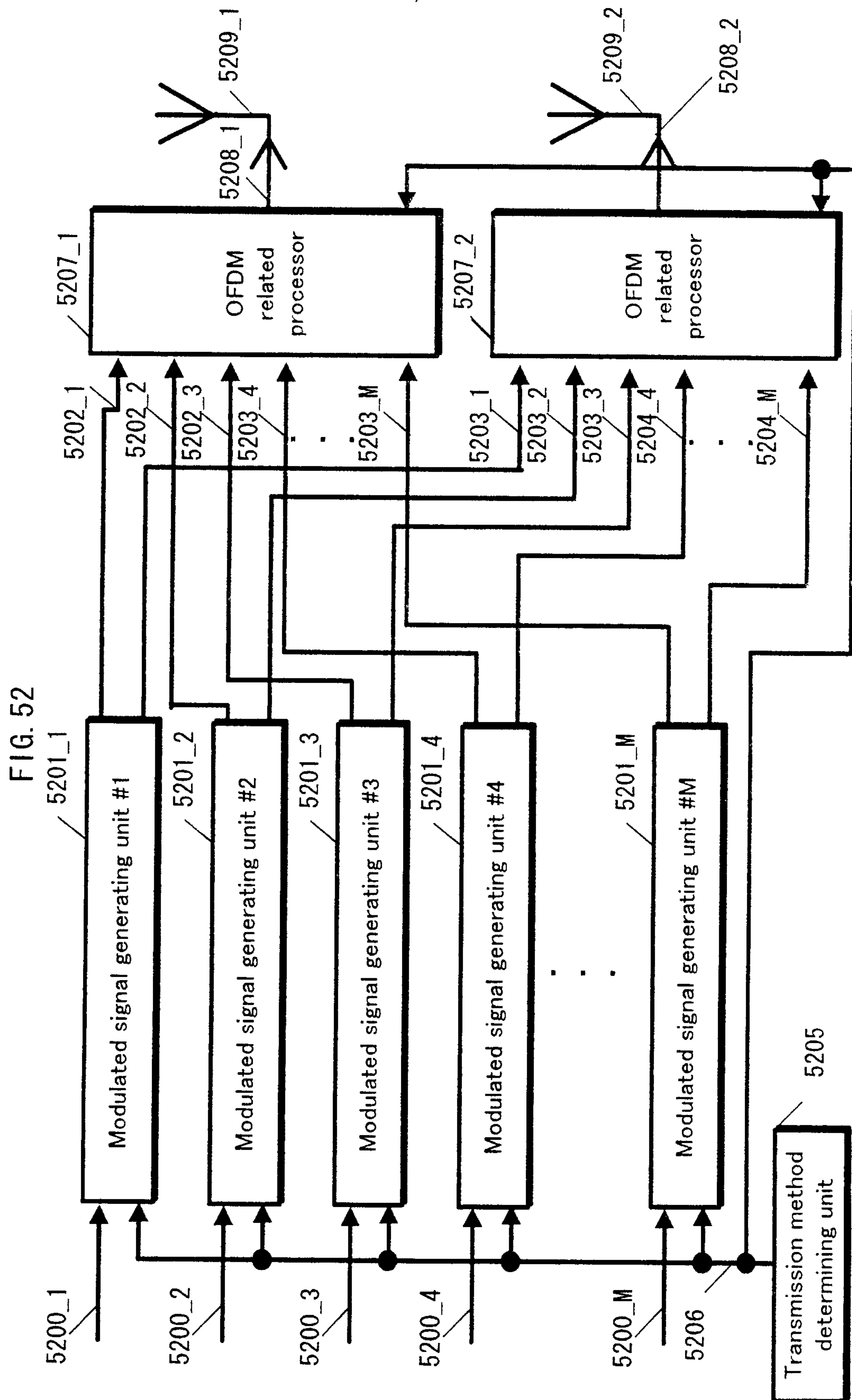


FIG. 51

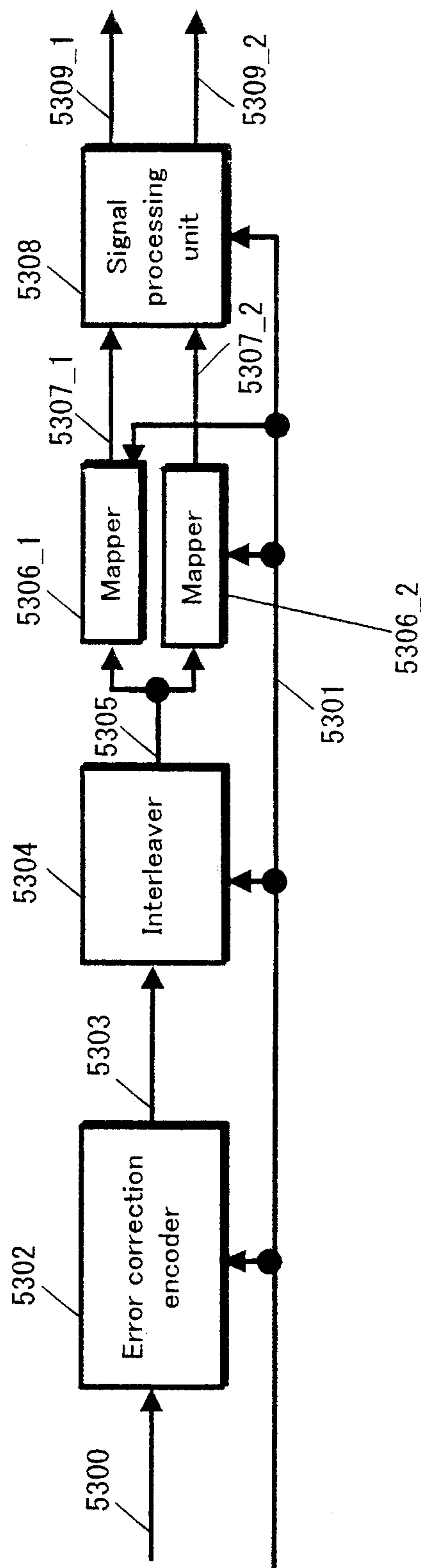


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FIG. 53



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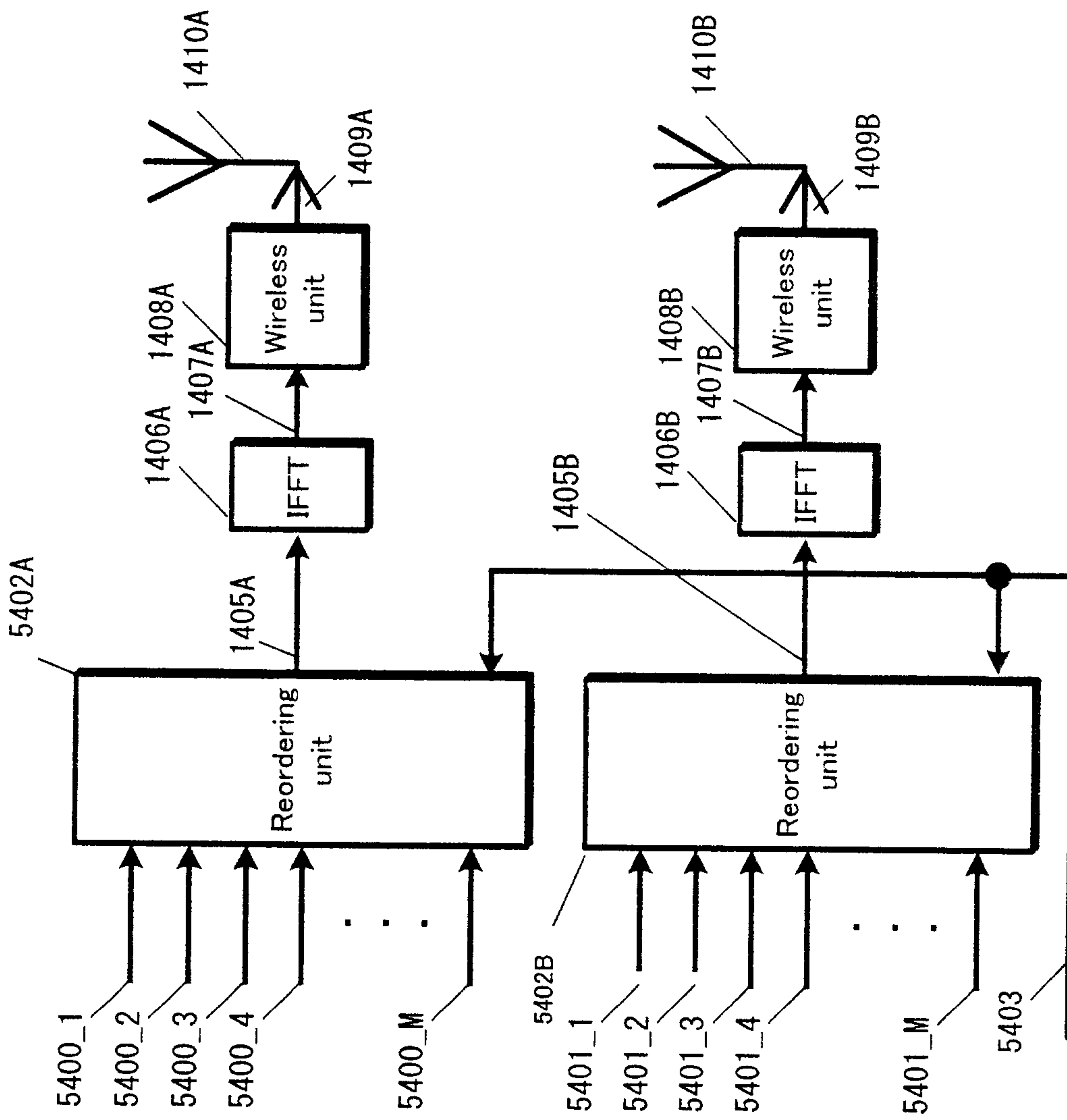


FIG. 55A

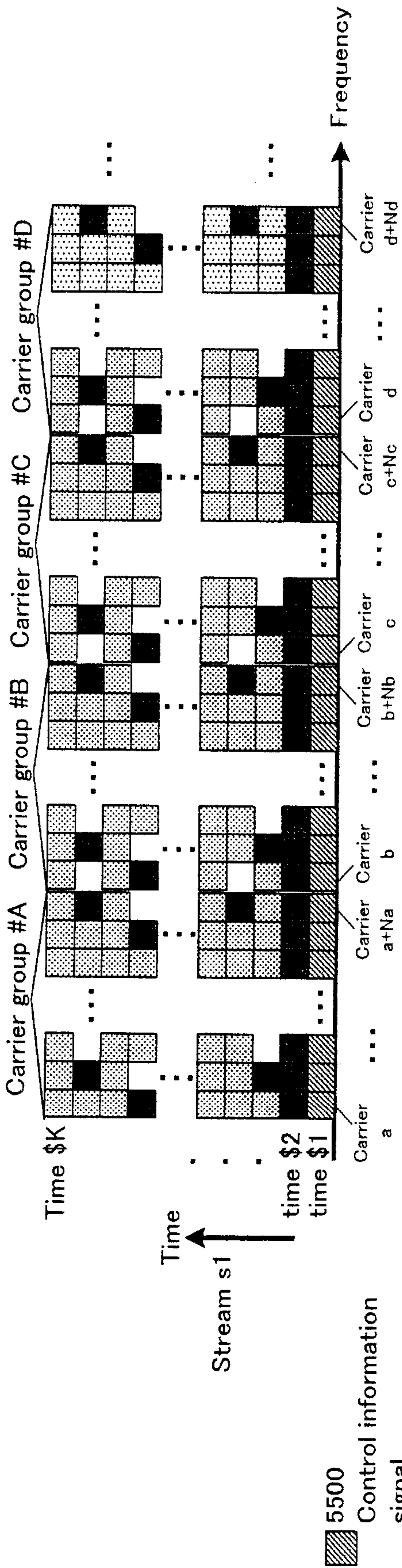
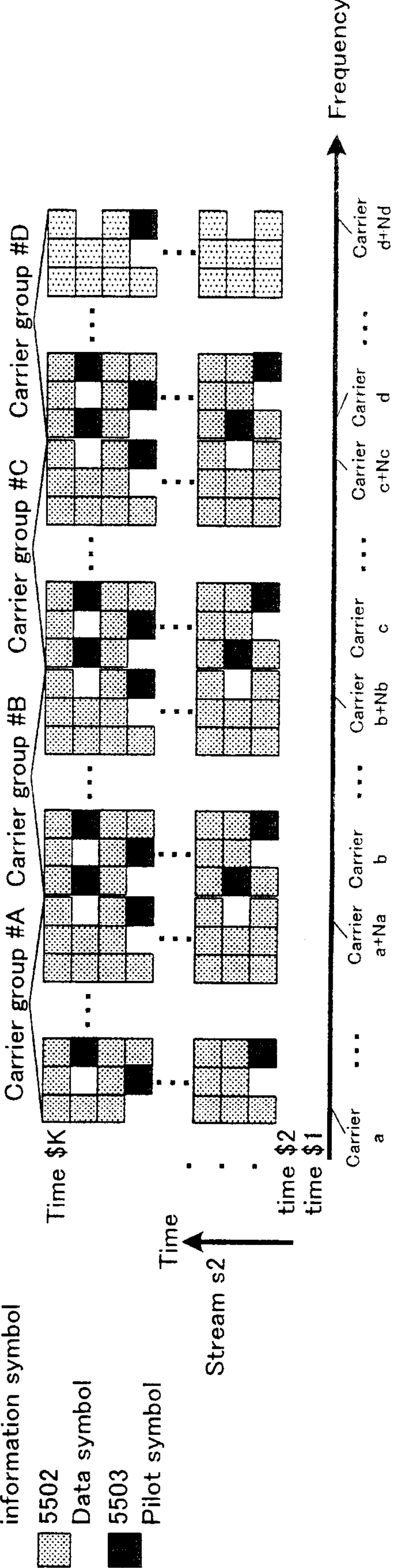
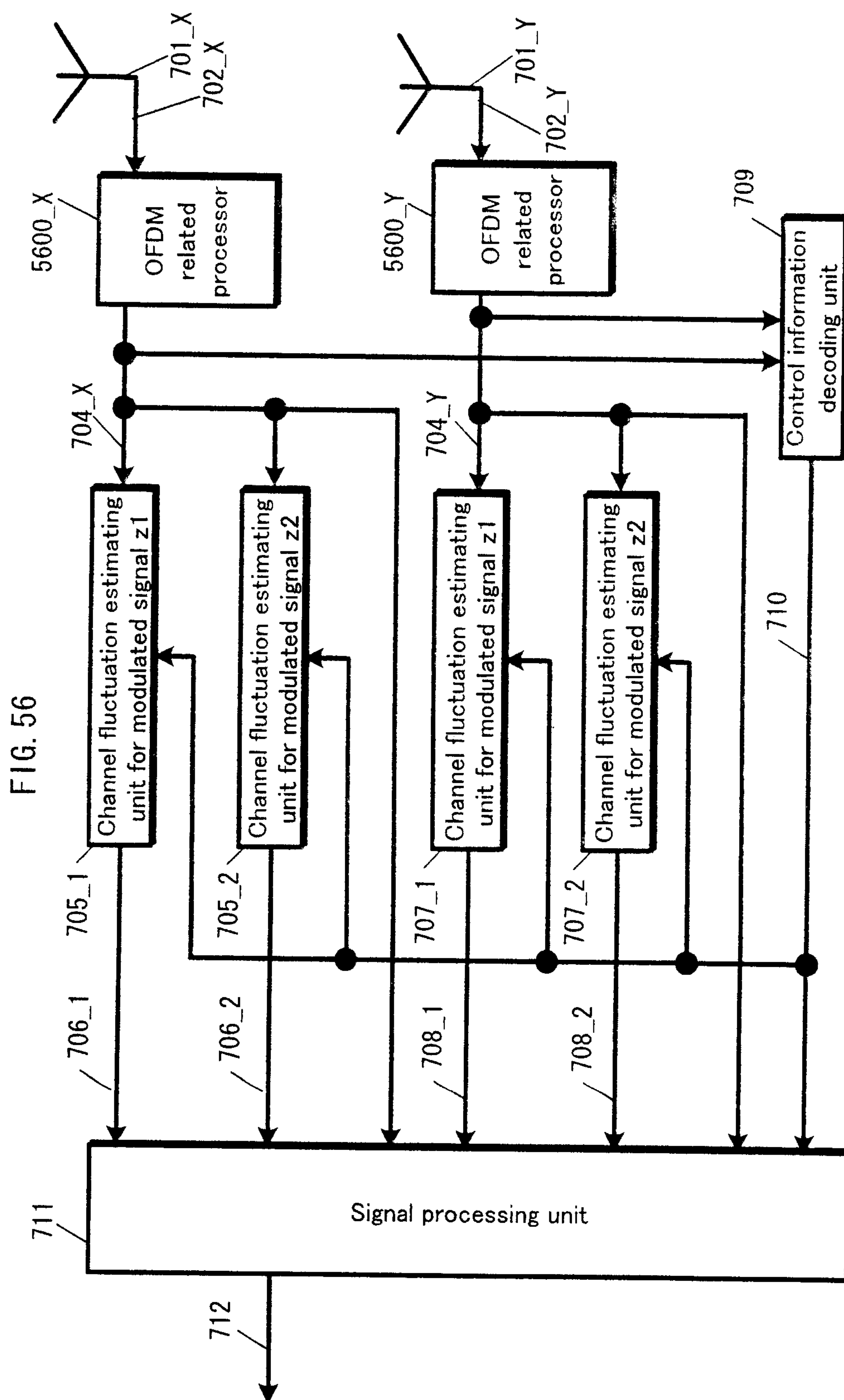


FIG. 55B



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FIG. 57

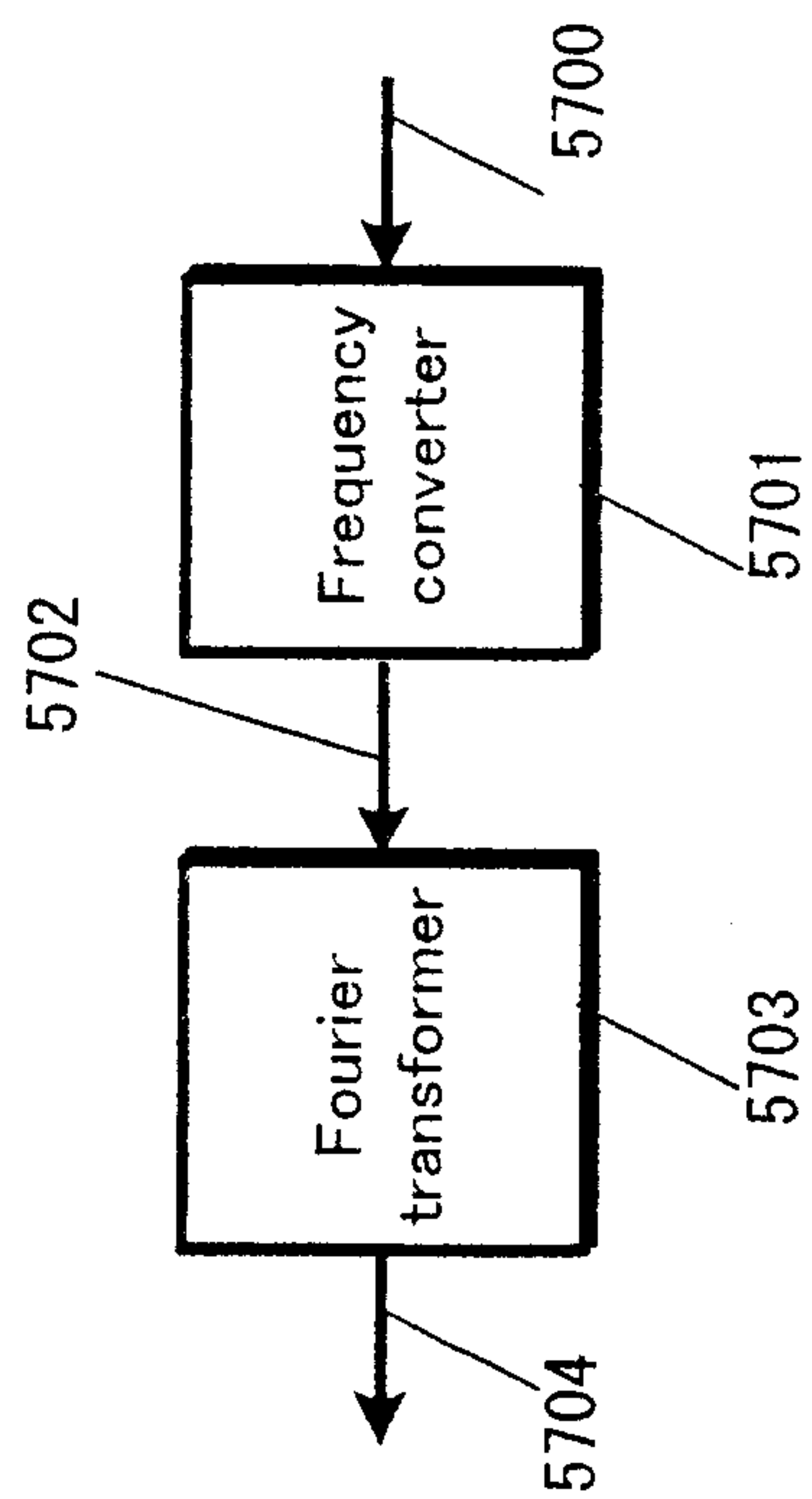


FIG. 58A

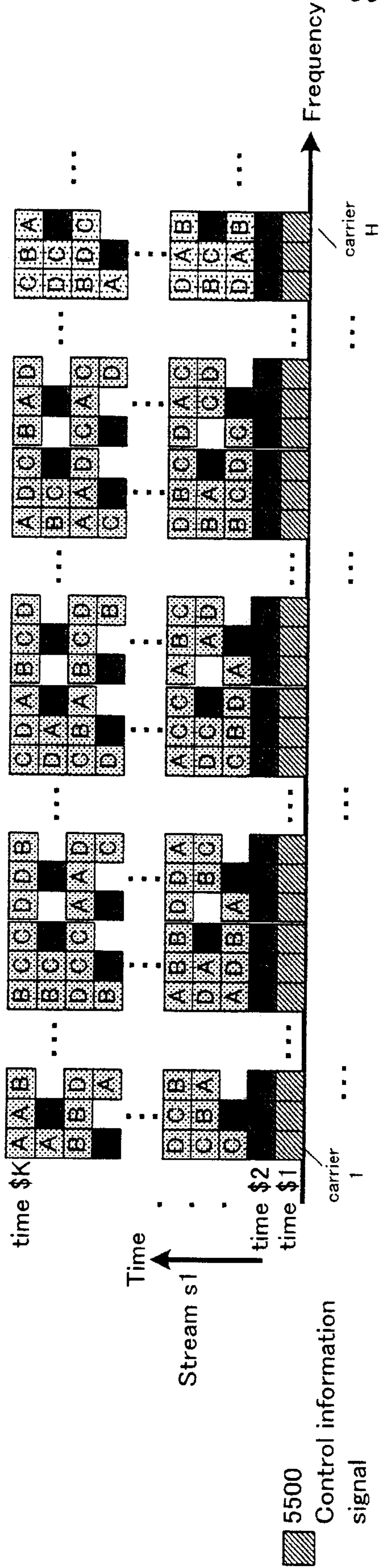


FIG. 58B

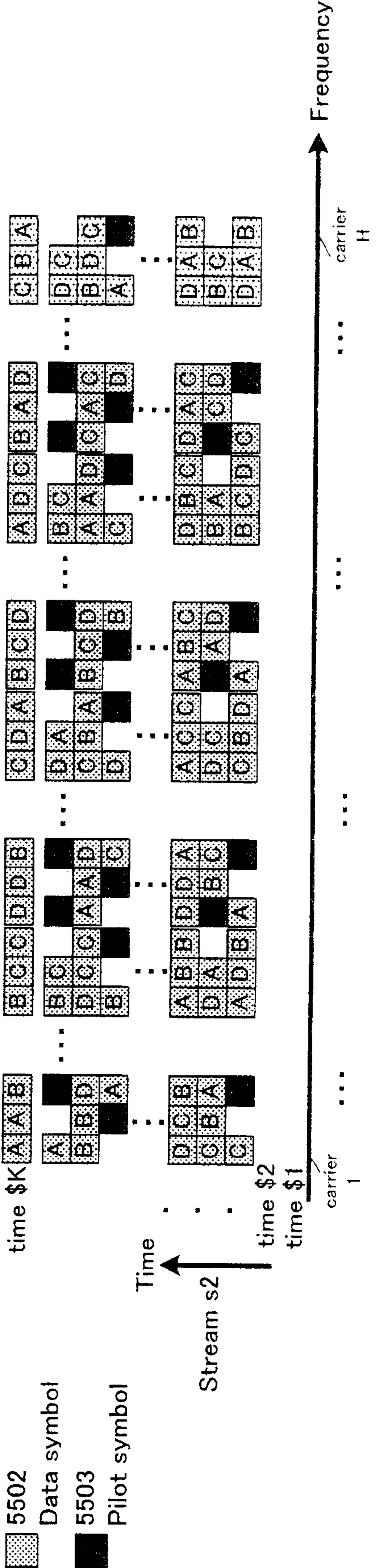


FIG. 59

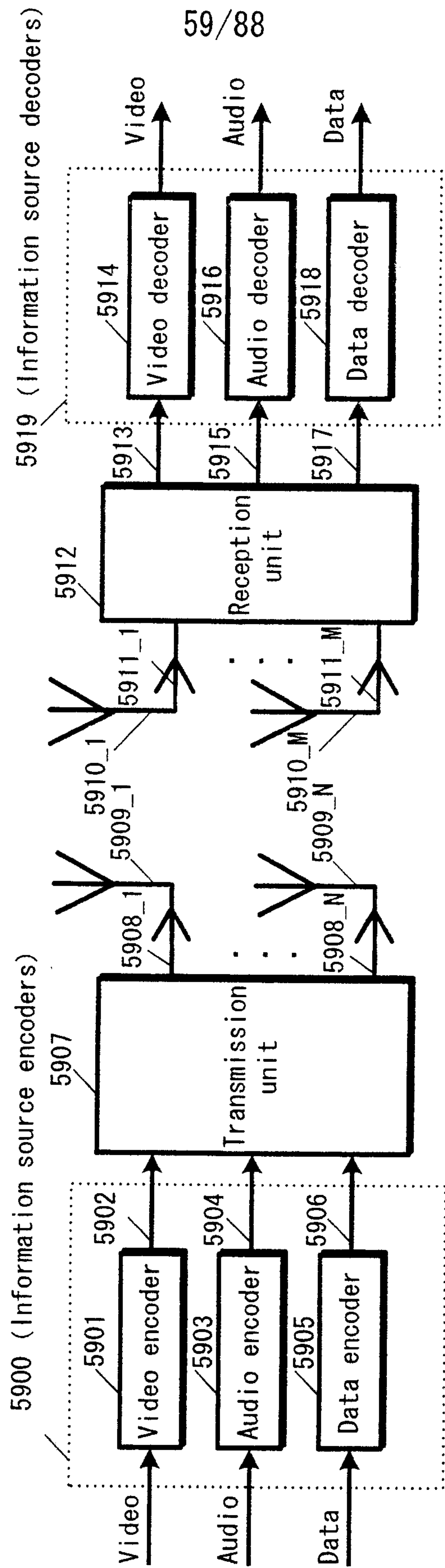
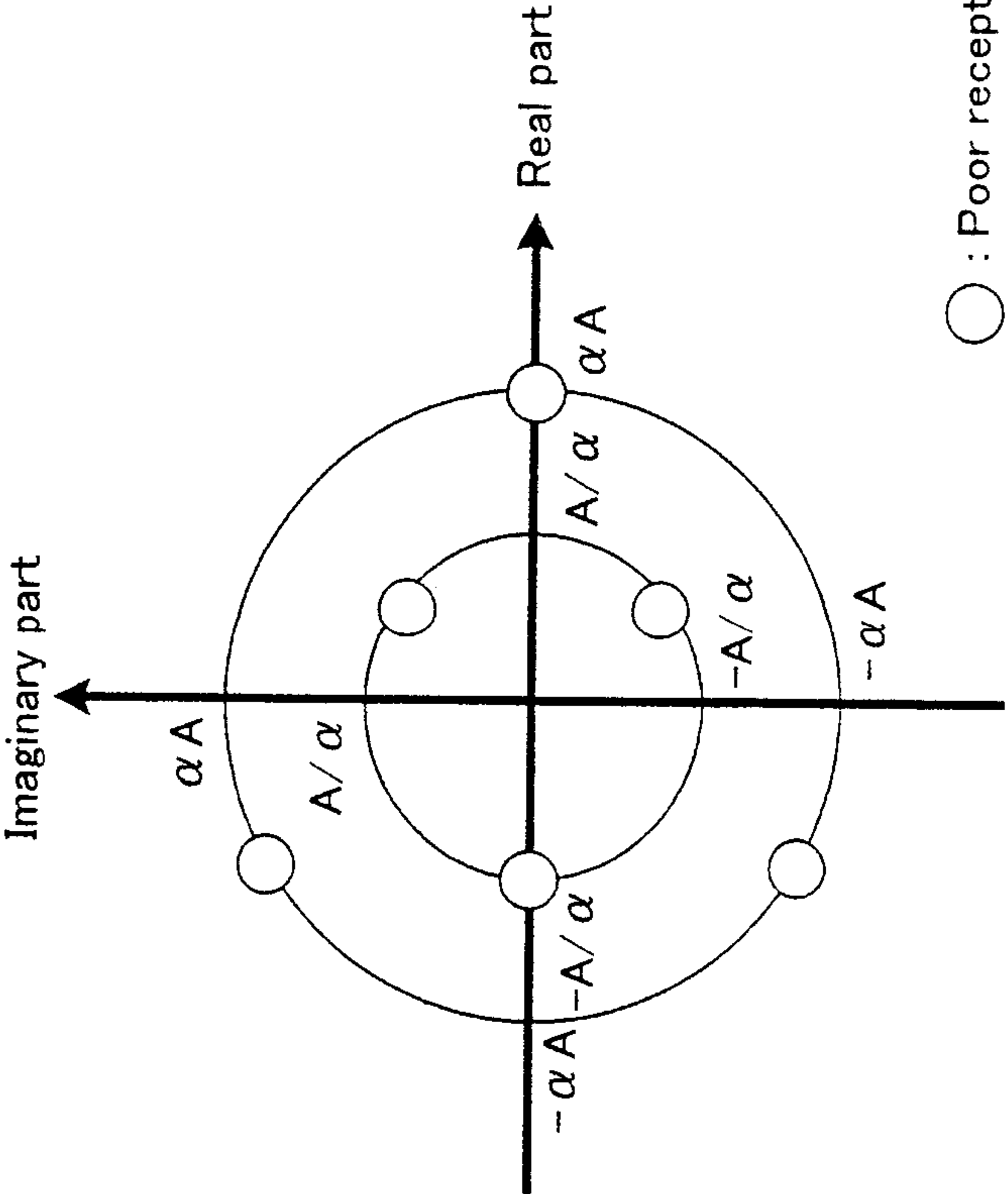
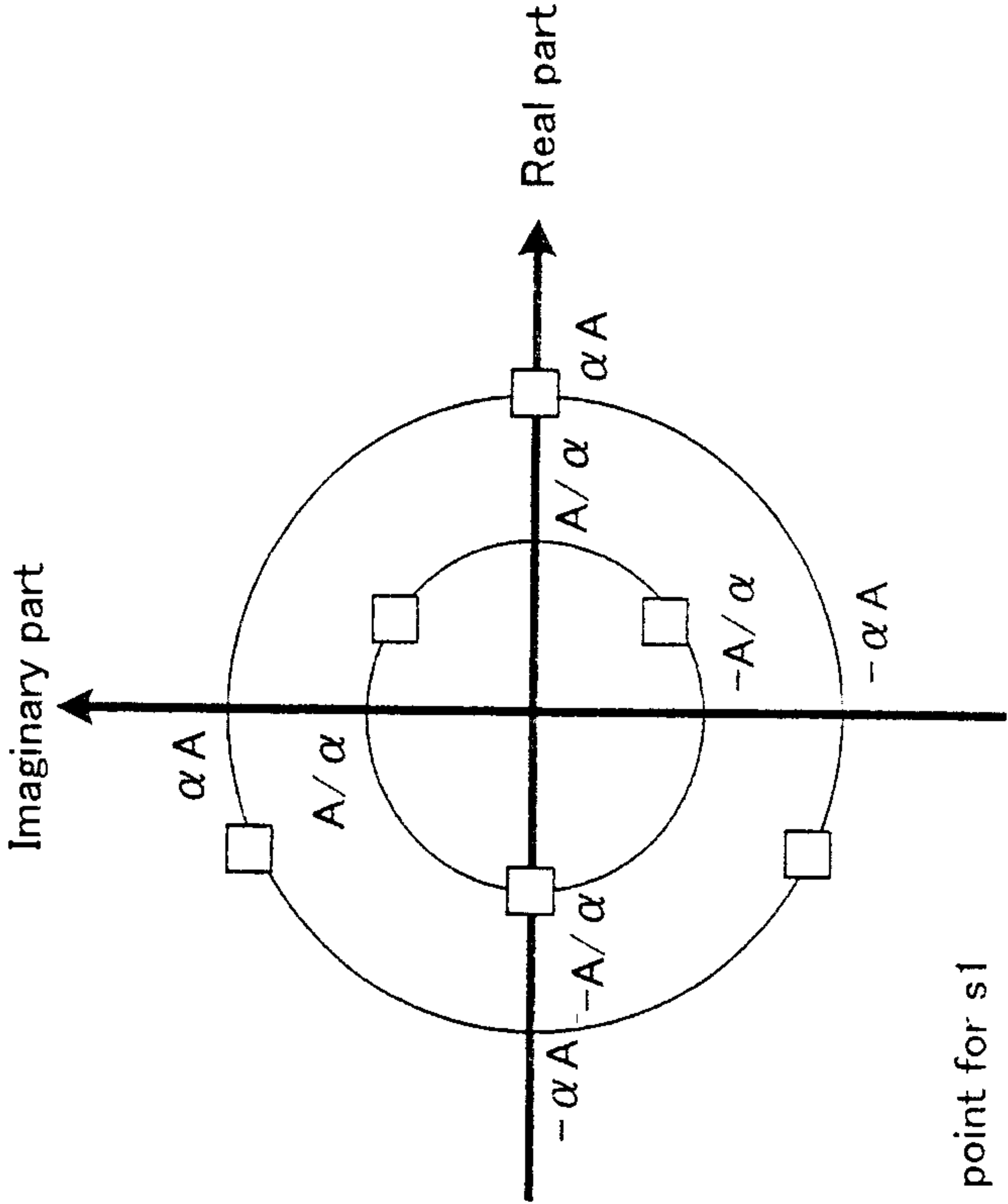


FIG. 60A



$\alpha > 1, N=3$

FIG. 60B



$\alpha > 1, N=3$

- : Poor reception point for s1
- : Poor reception point for s2

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FIG. 61

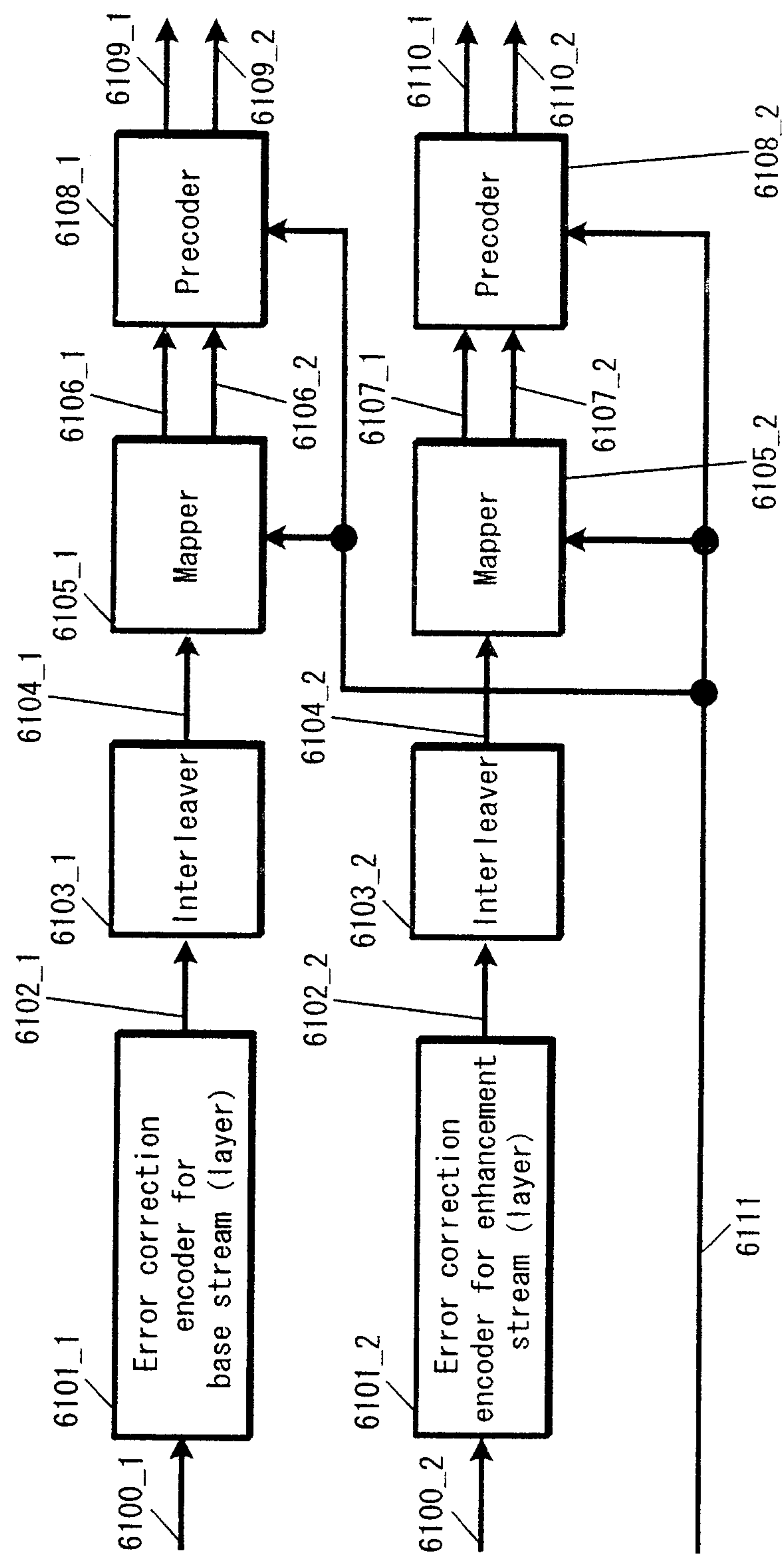


FIG. 62

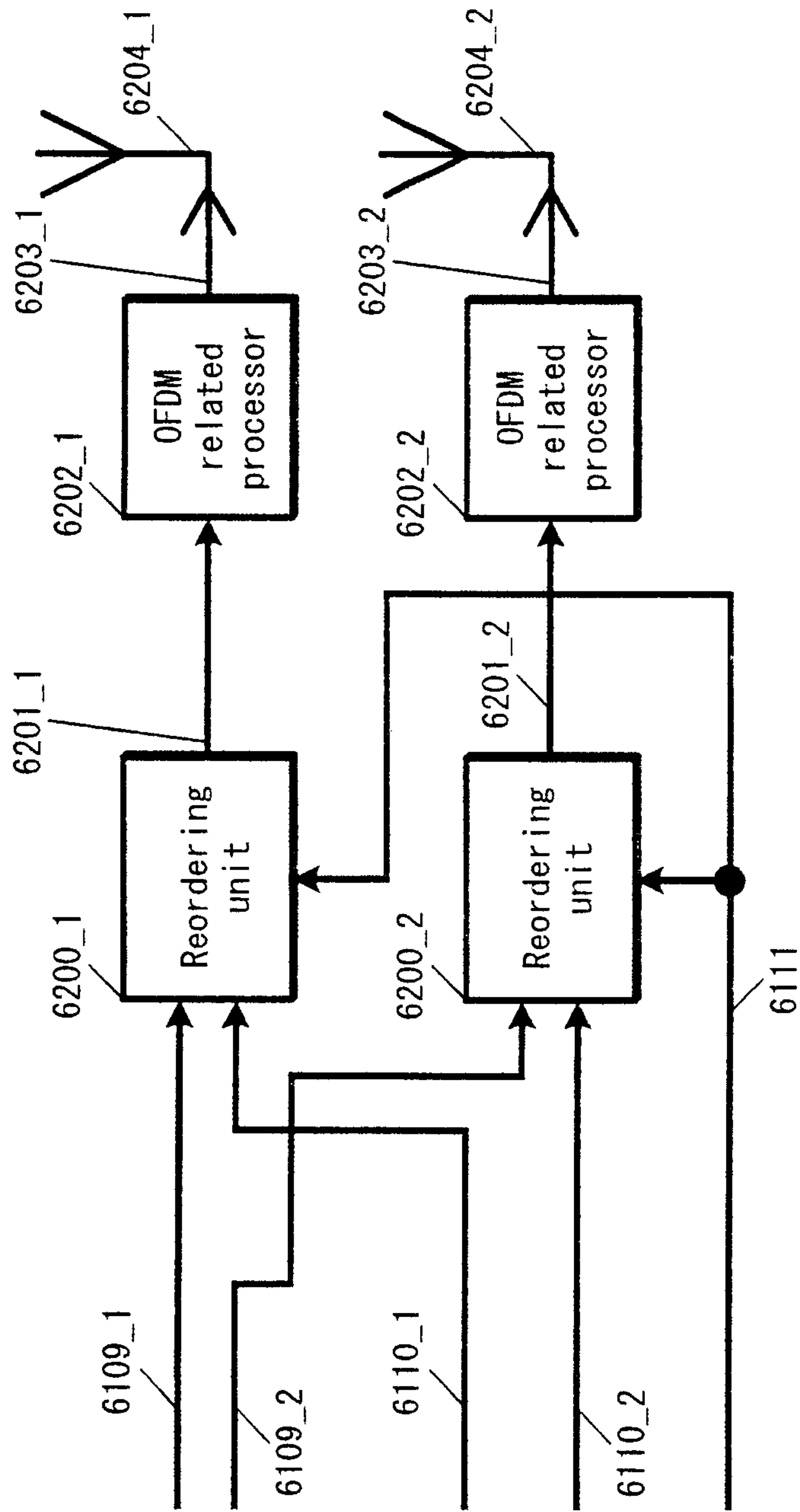


FIG. 63

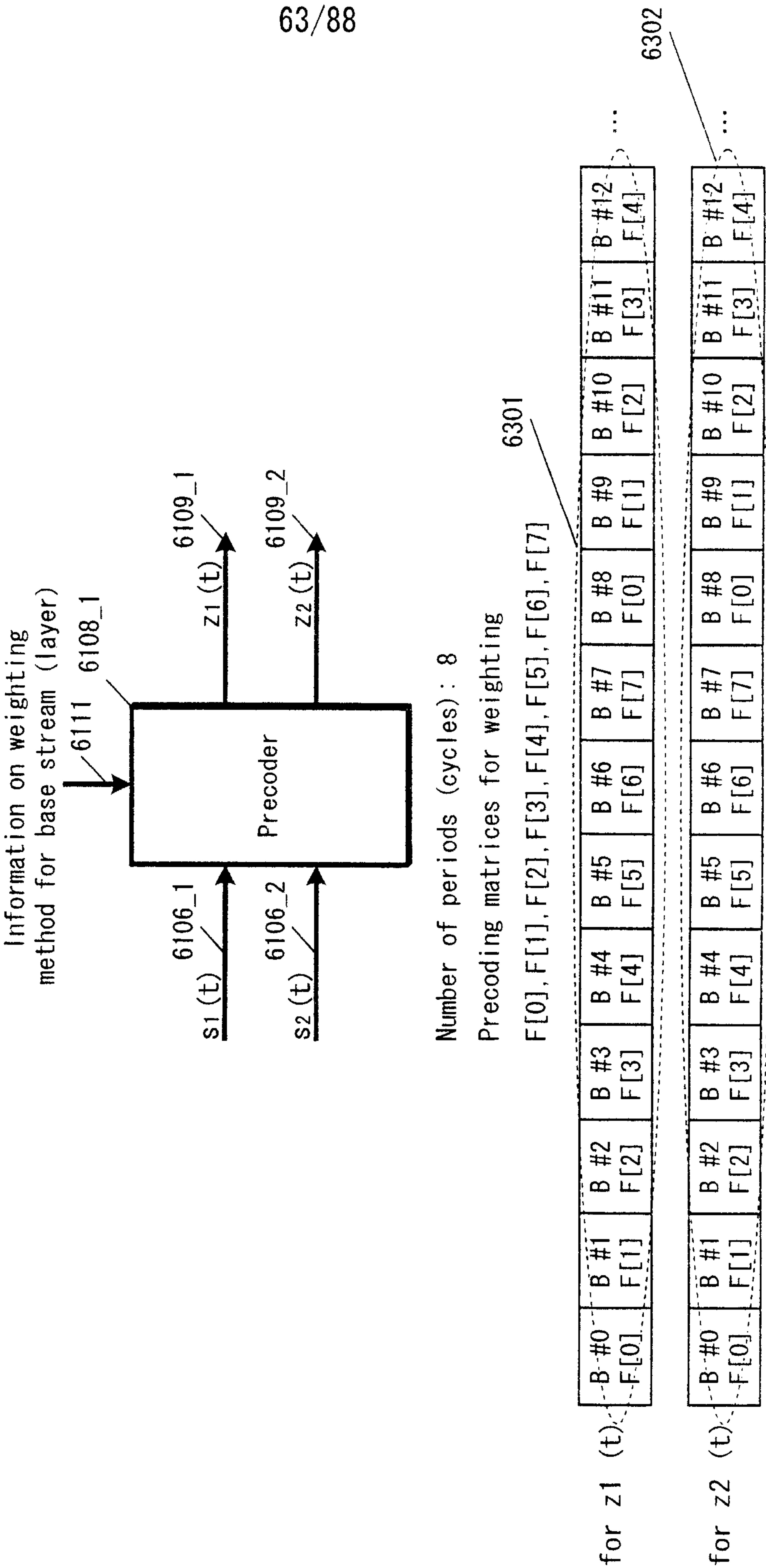


FIG. 64

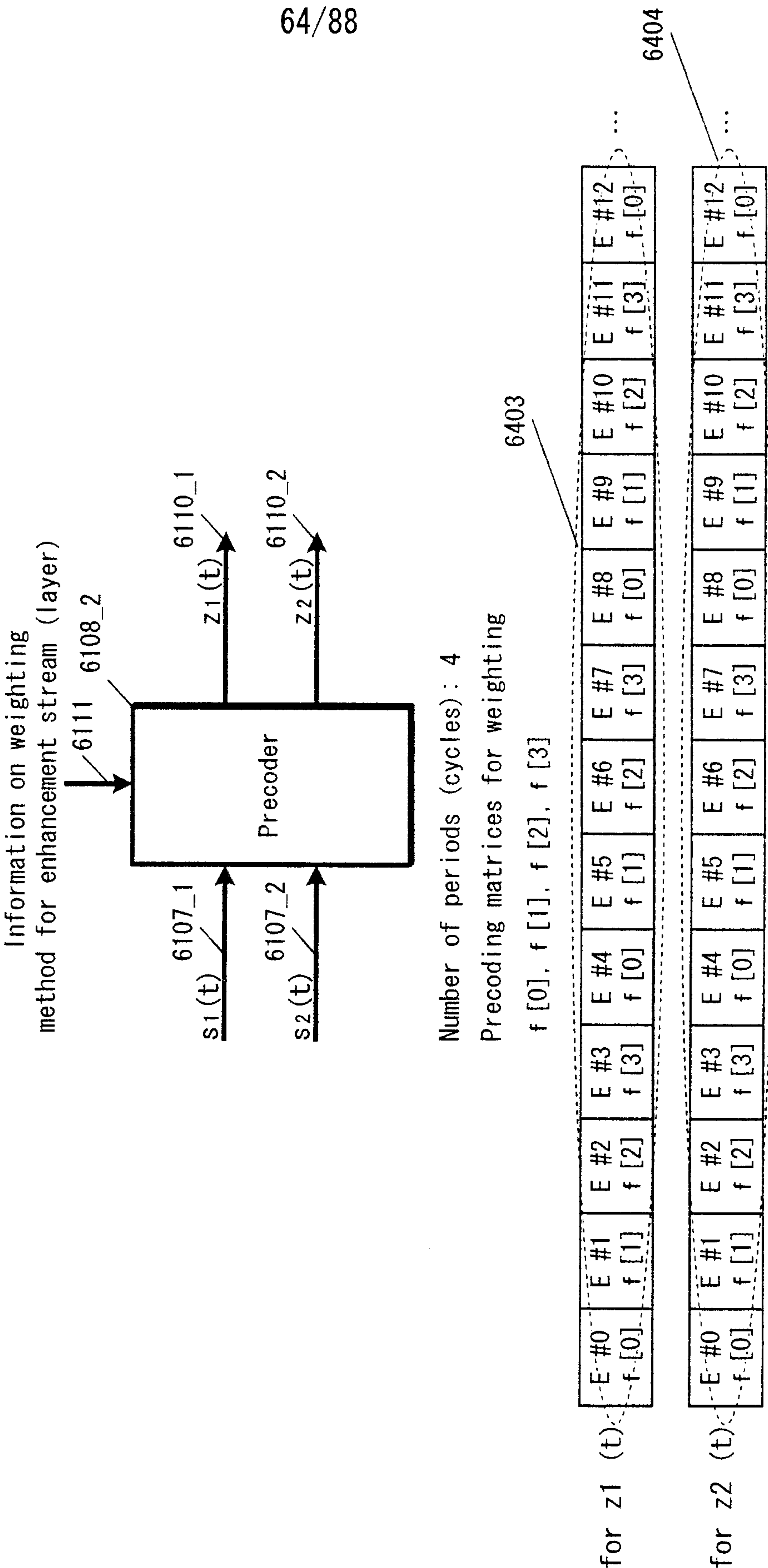


FIG. 65A

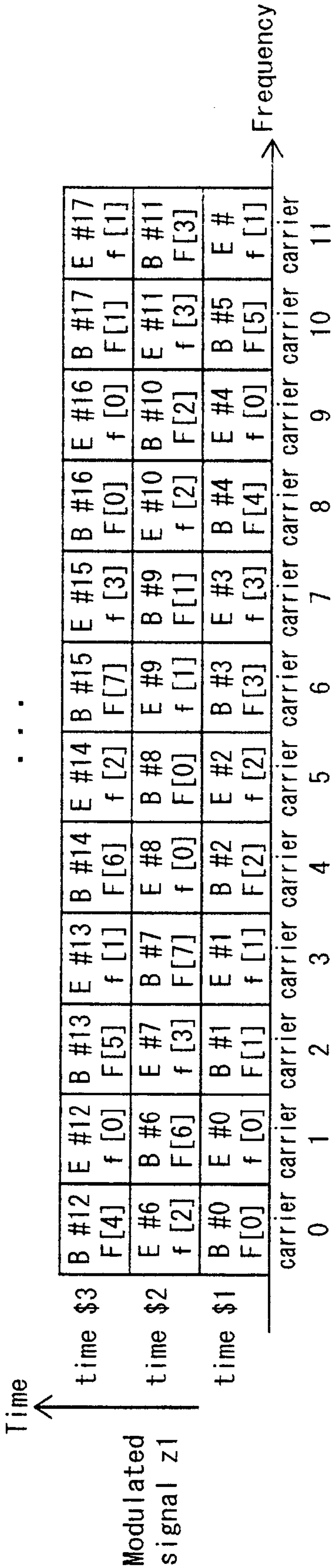
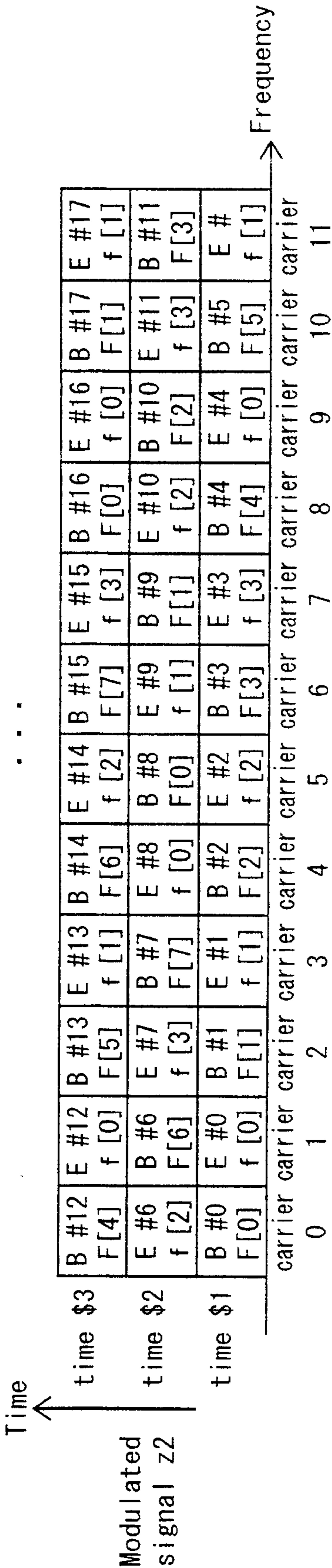
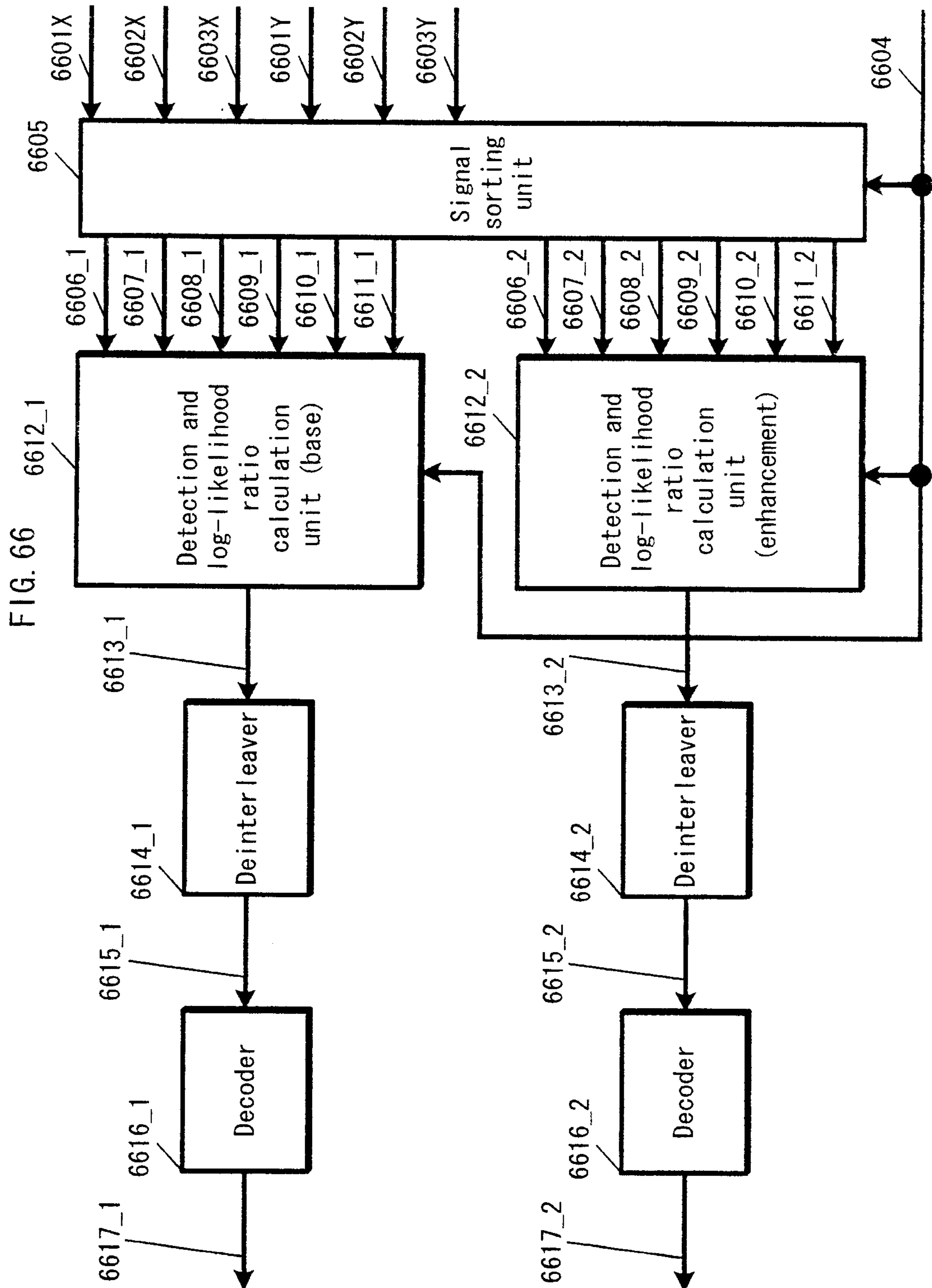


FIG. 65B



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FIG. 67

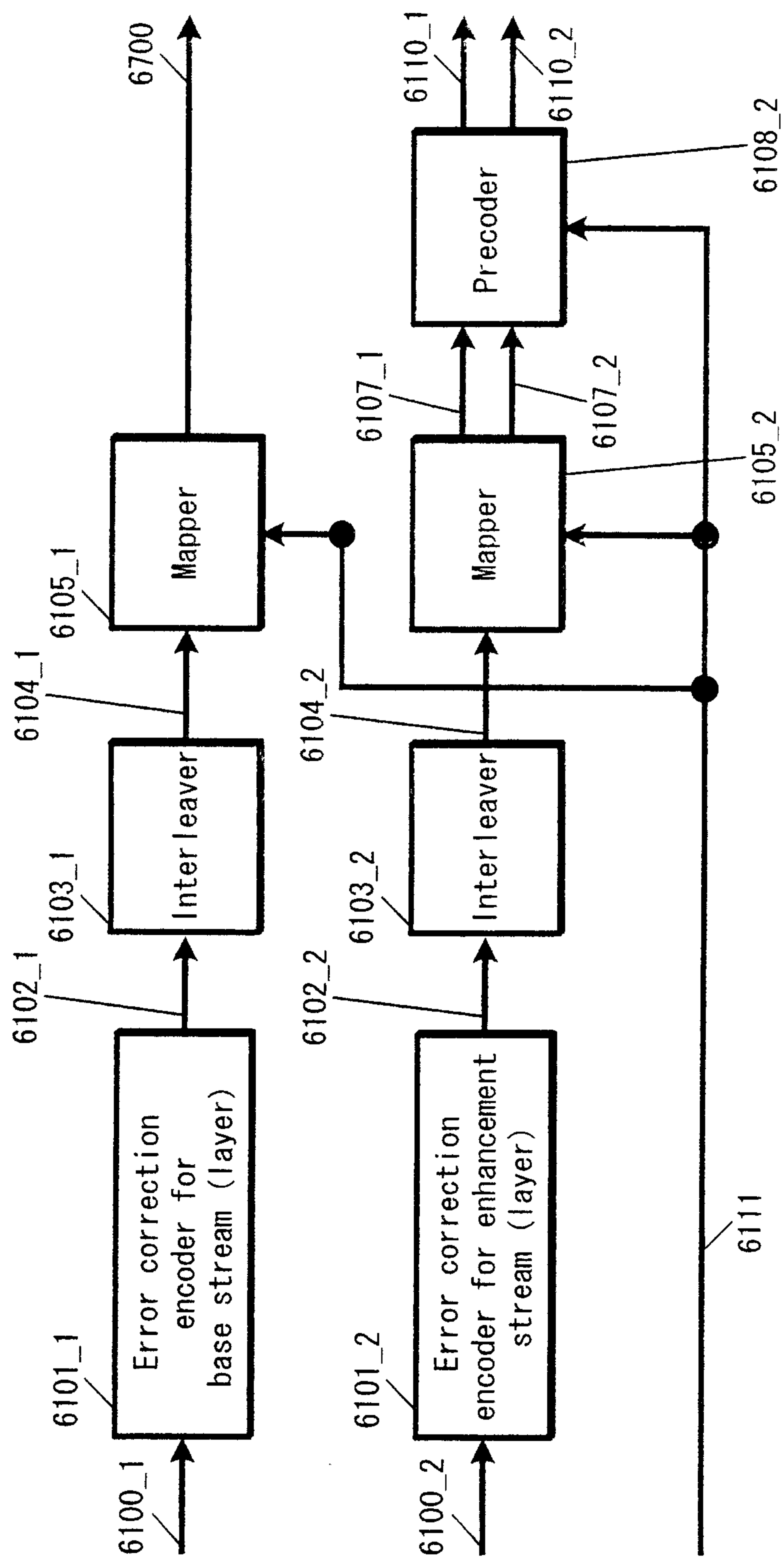


FIG. 68

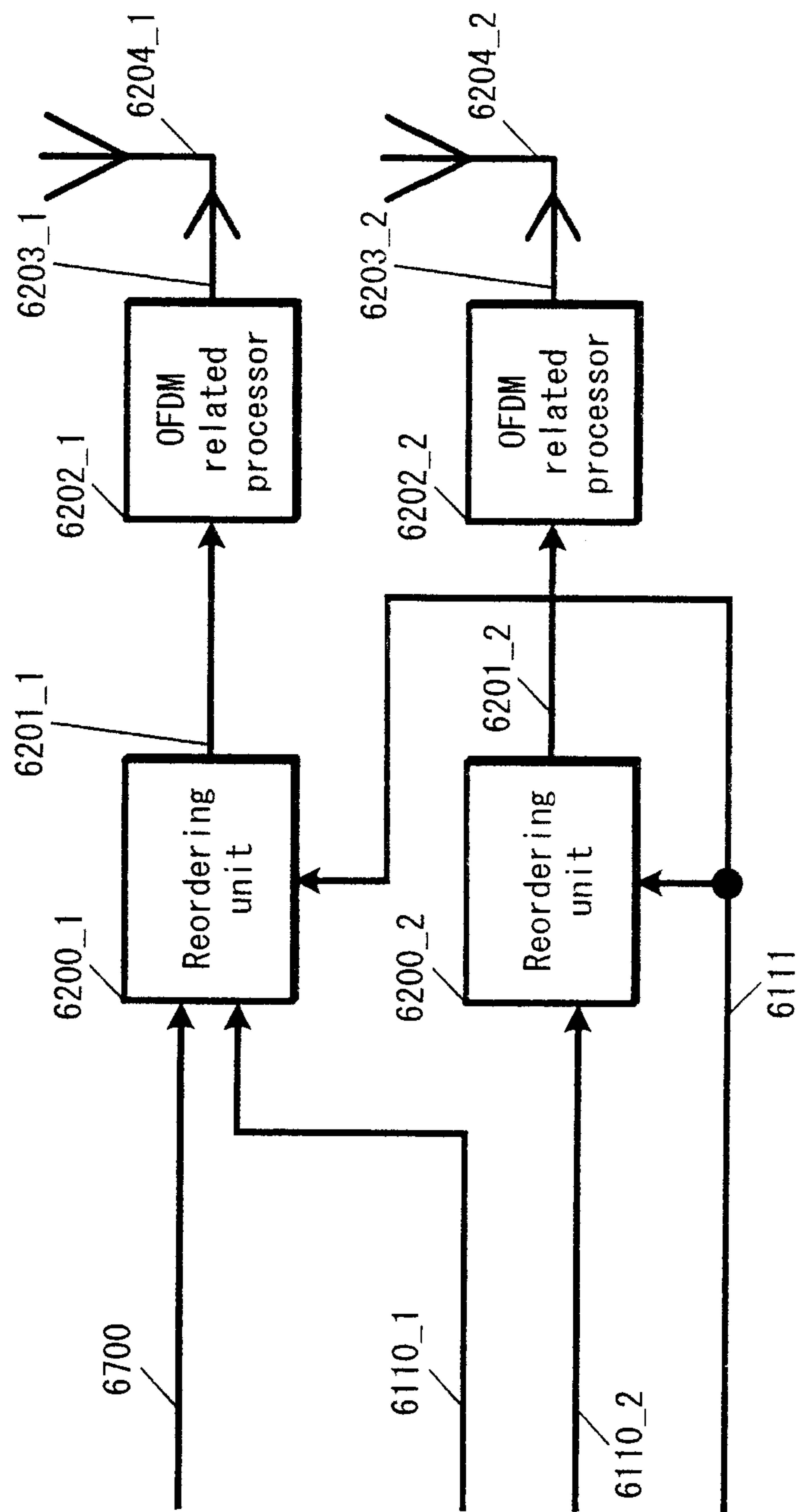


FIG. 69

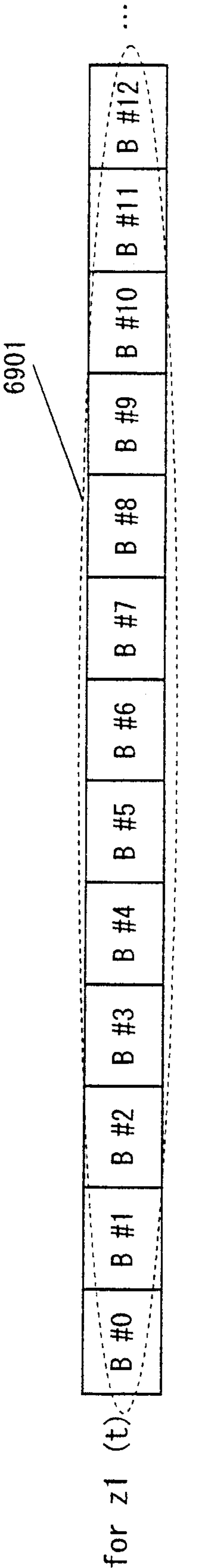


FIG. 70A

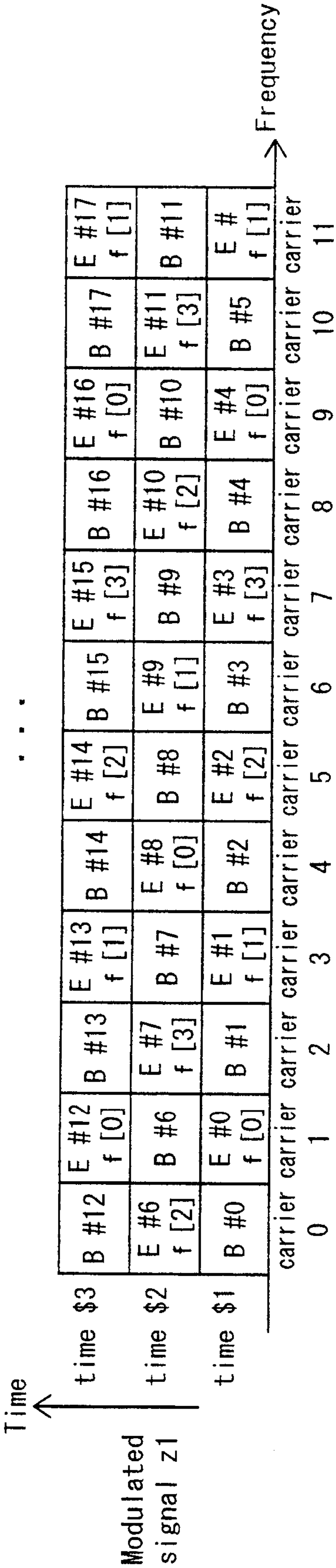
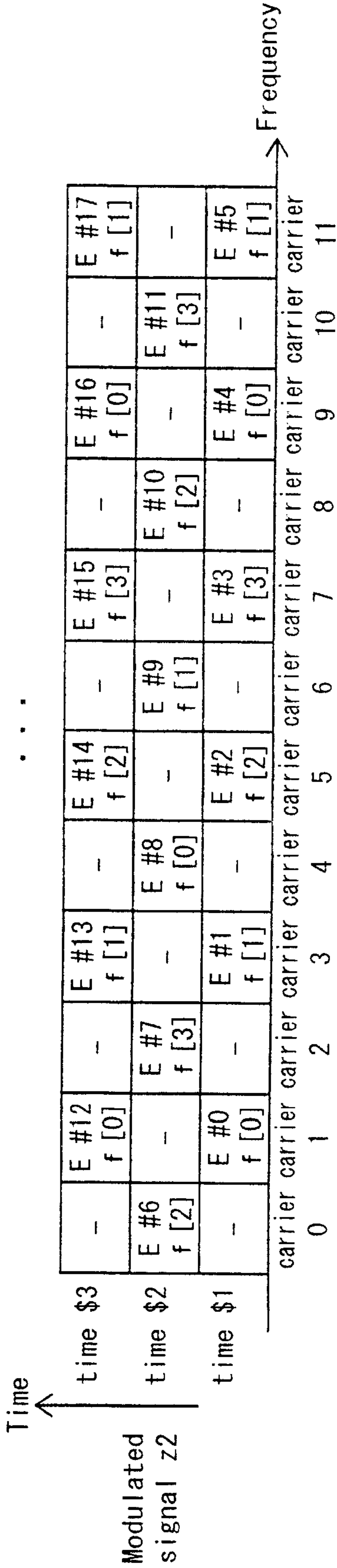


FIG. 70B



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FIG. 71

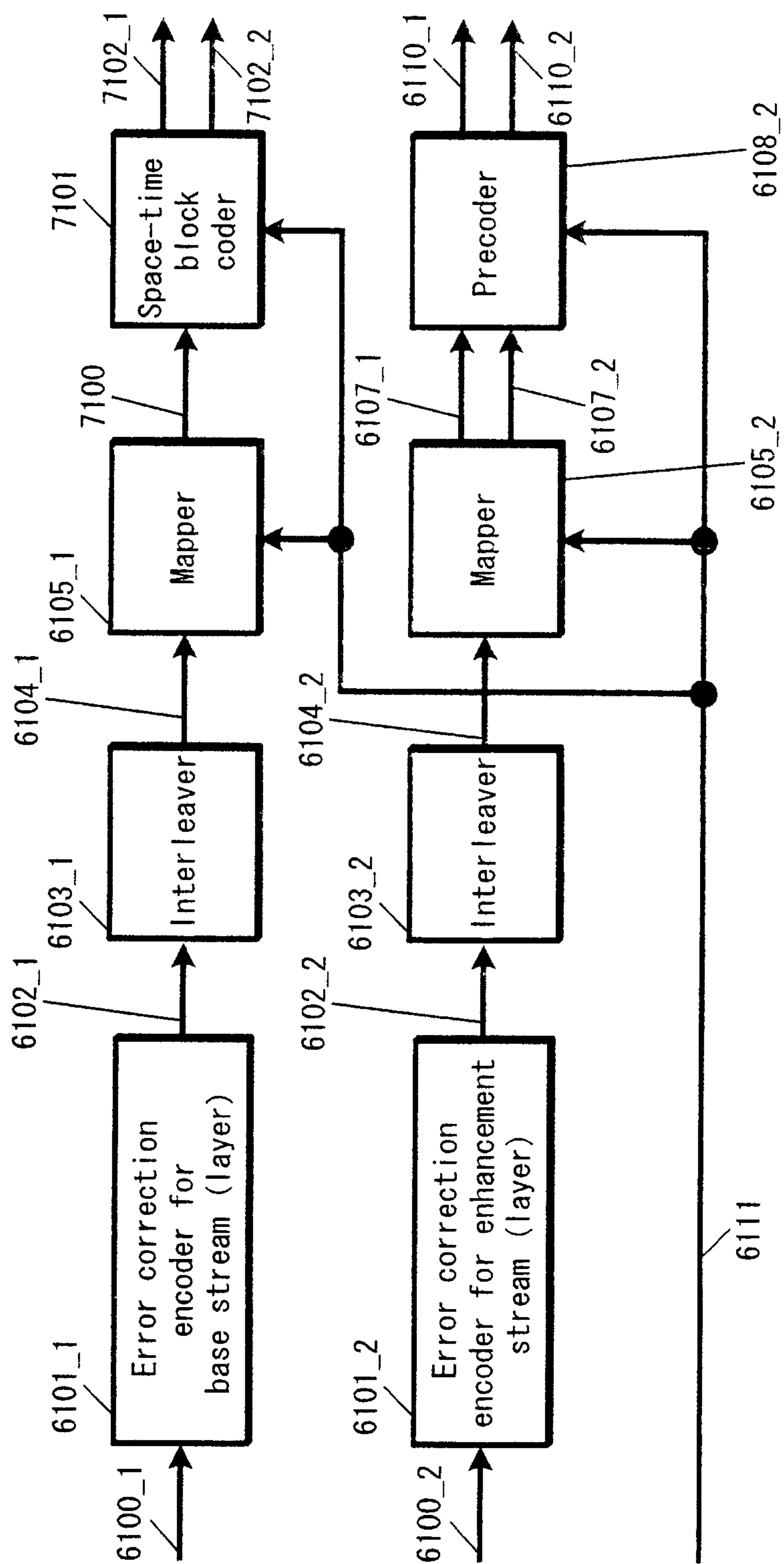


FIG. 72

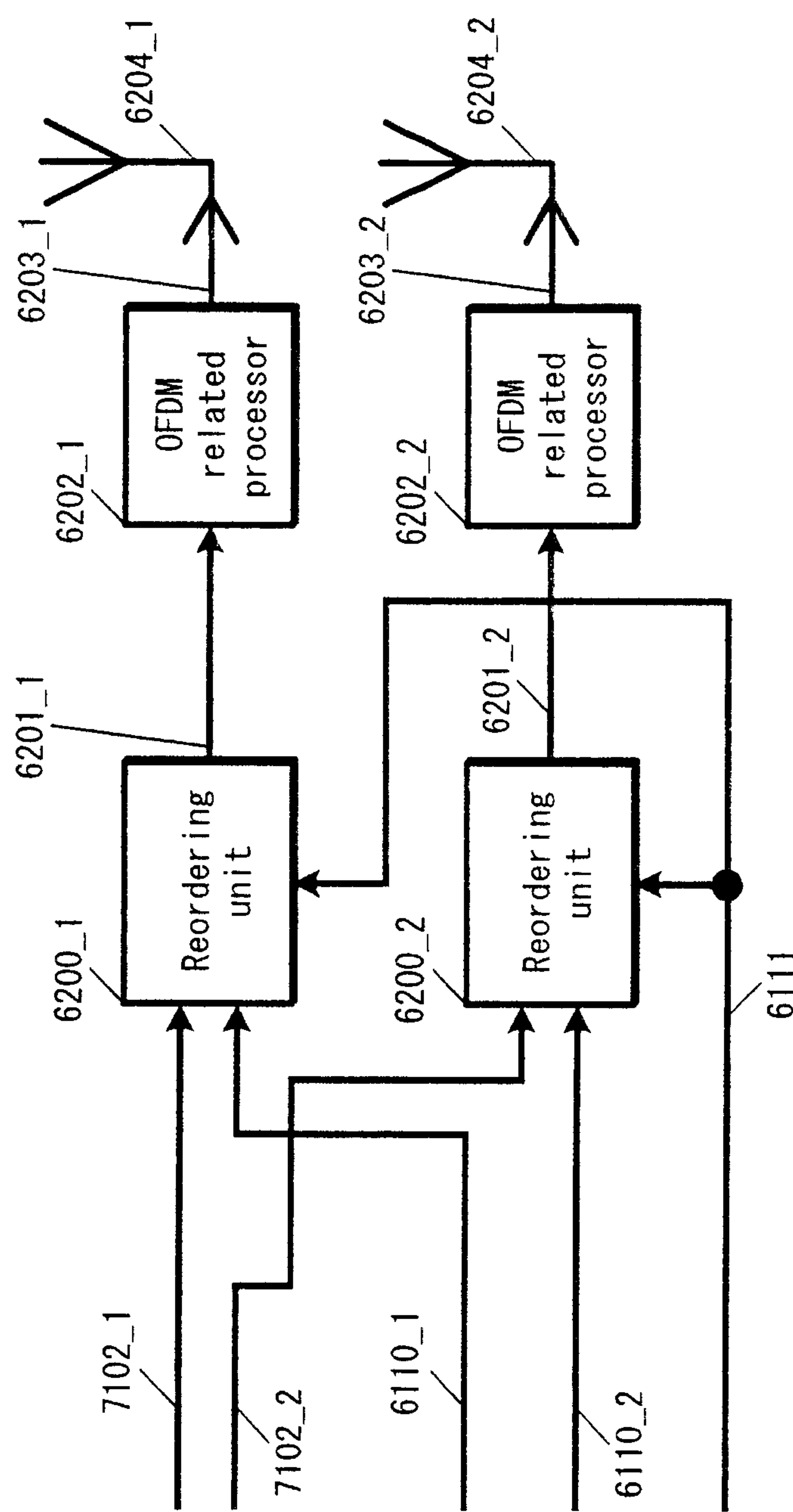


FIG. 73

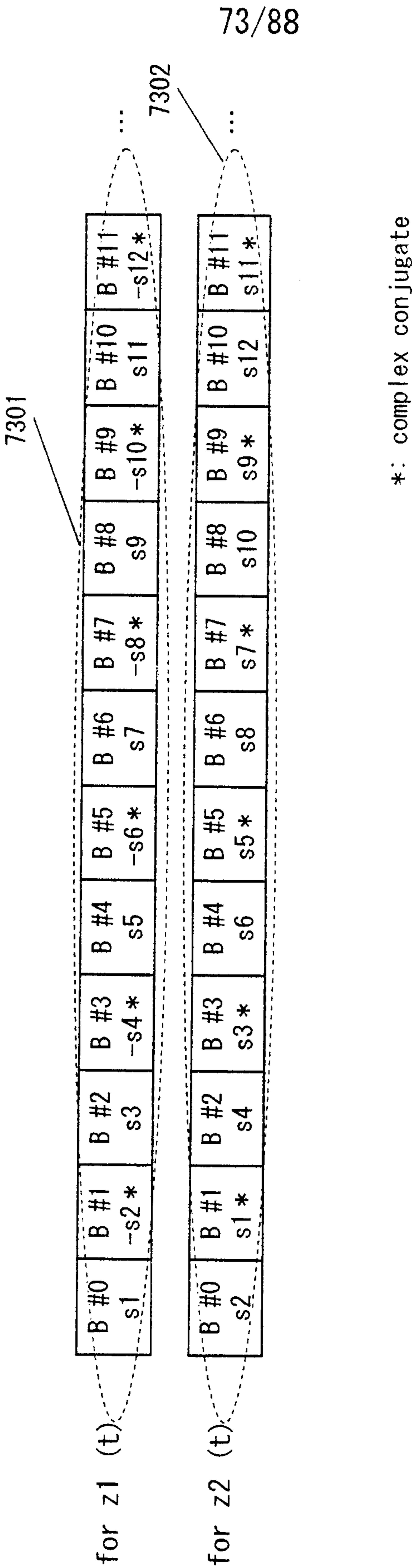


FIG. 74A

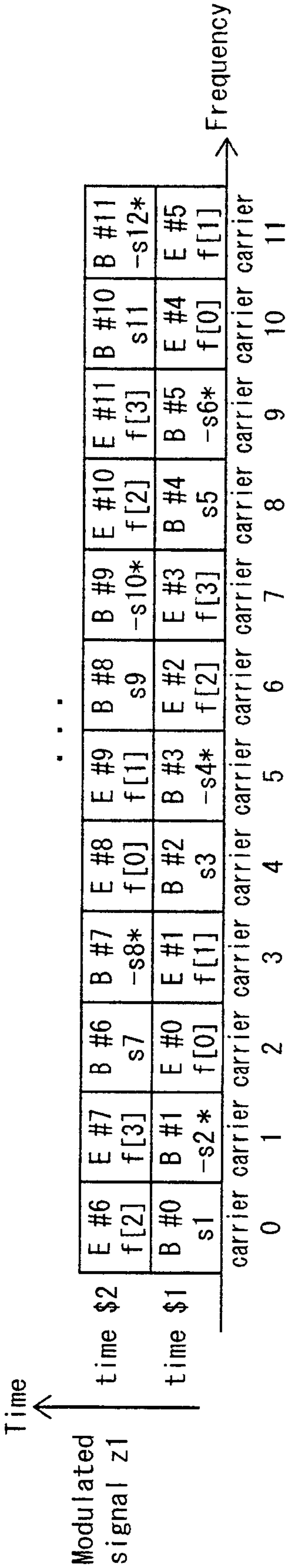


FIG. 74B

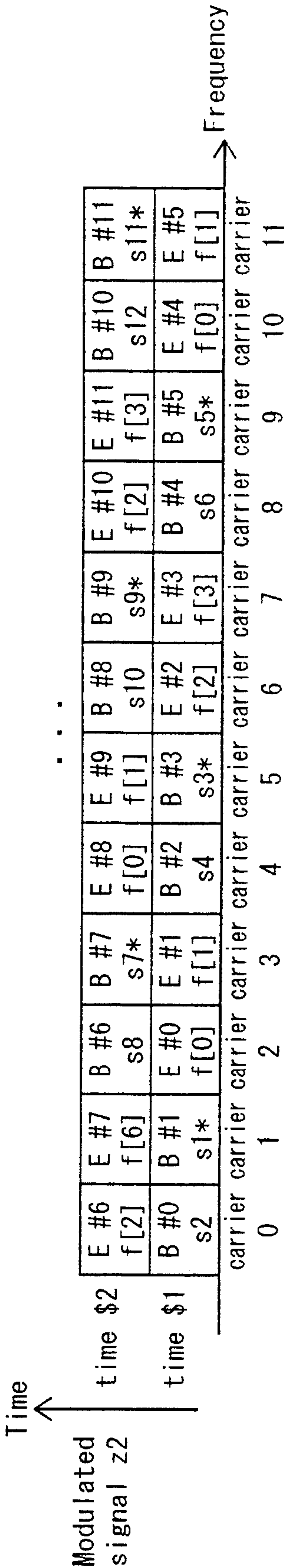


FIG. 75A

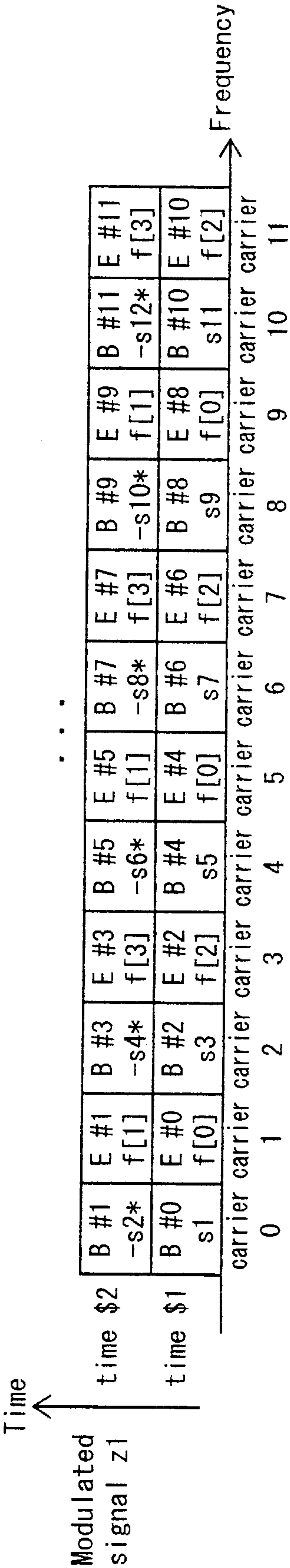


FIG. 75B

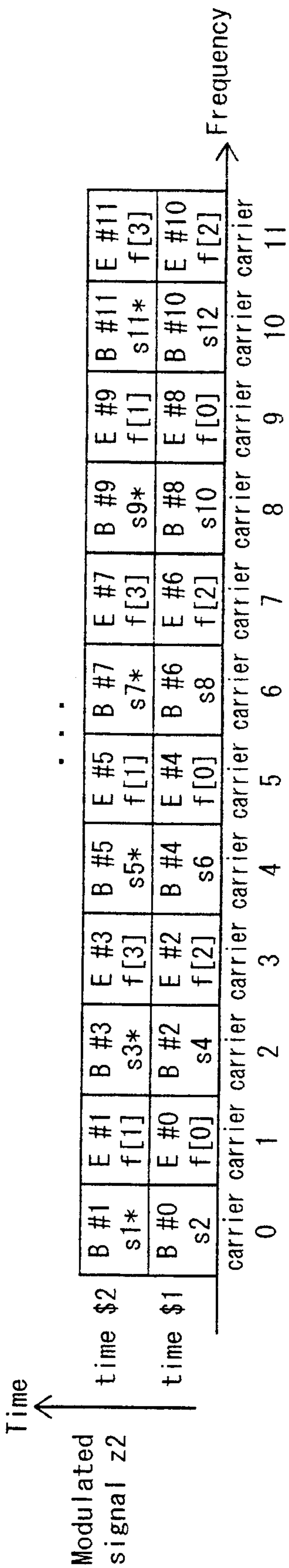


FIG. 76

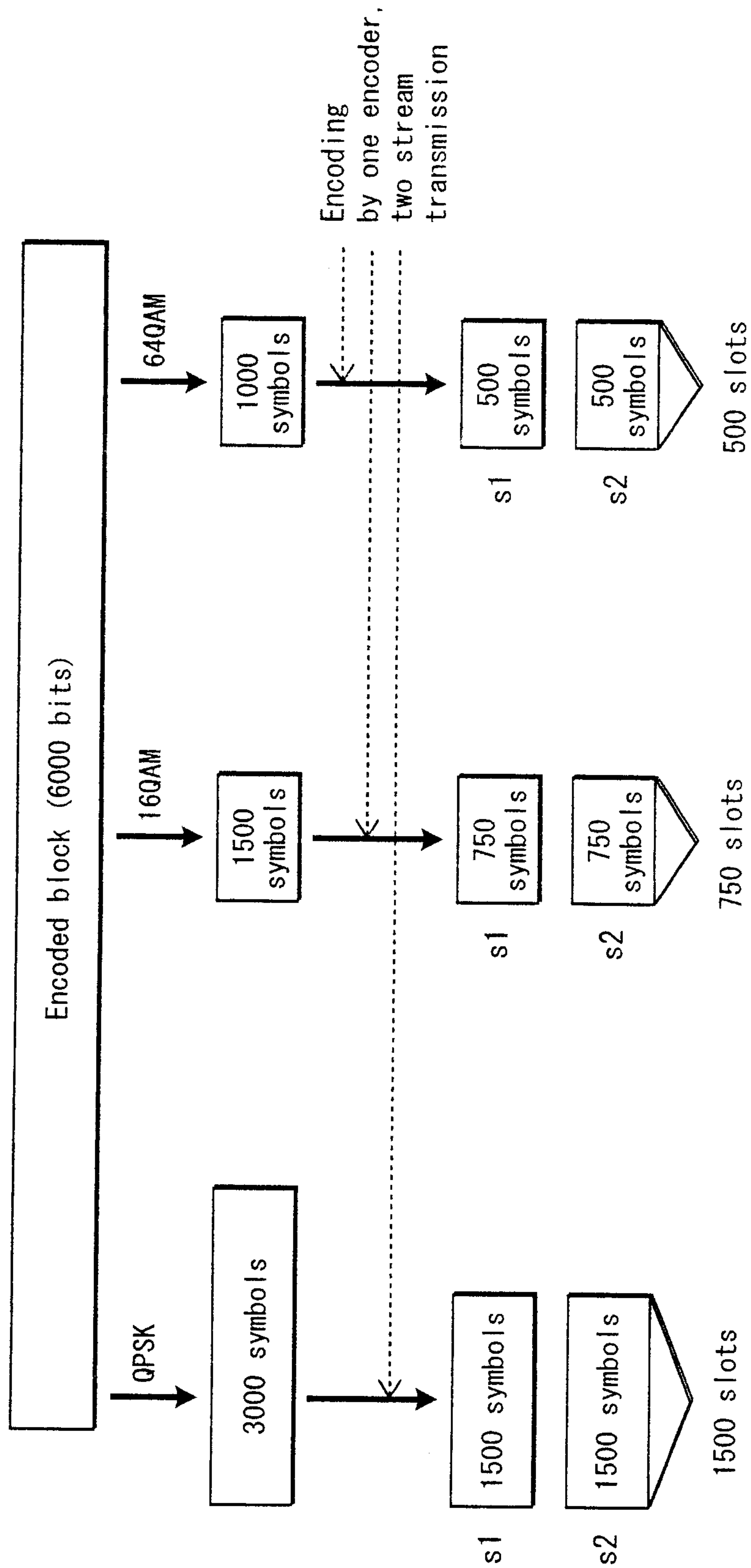
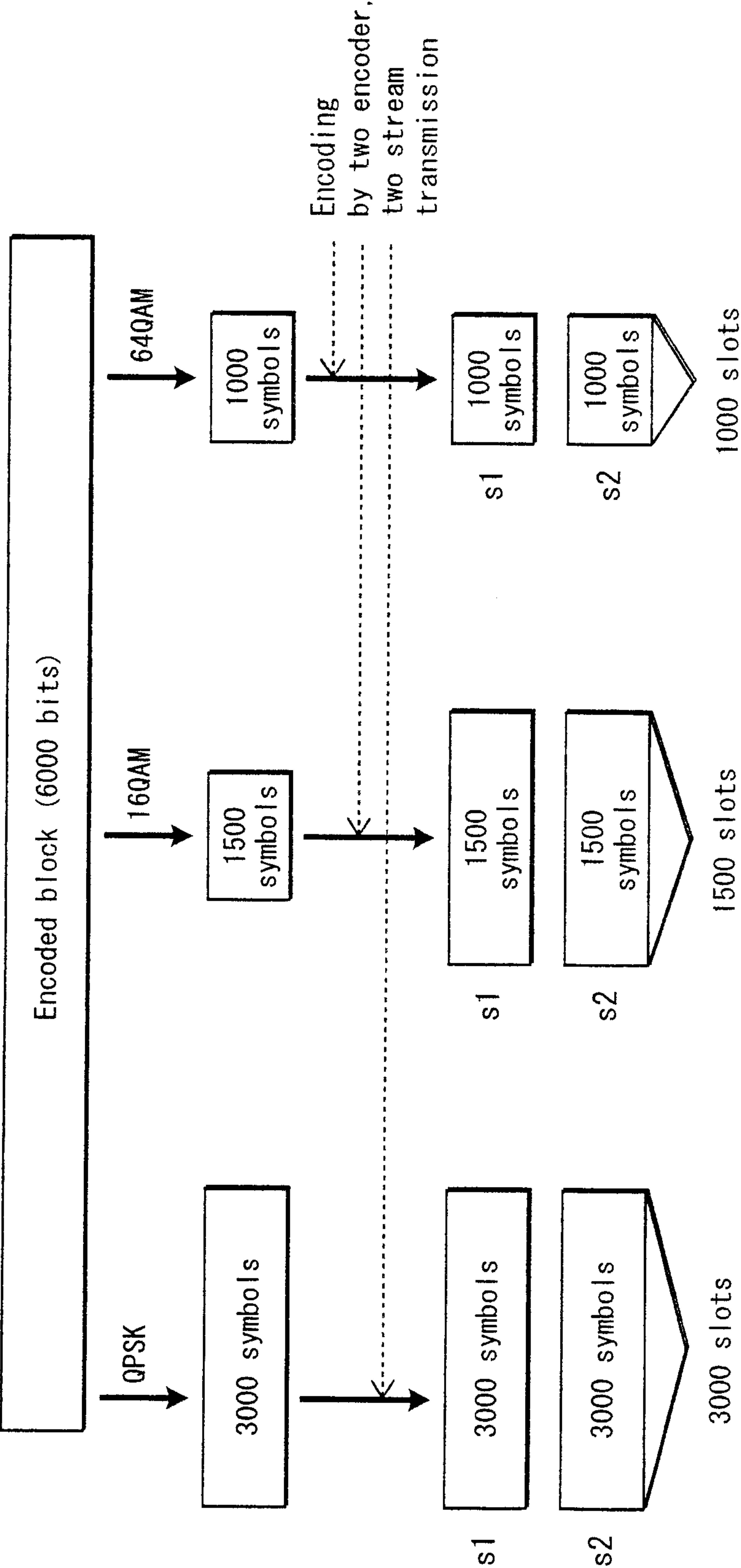
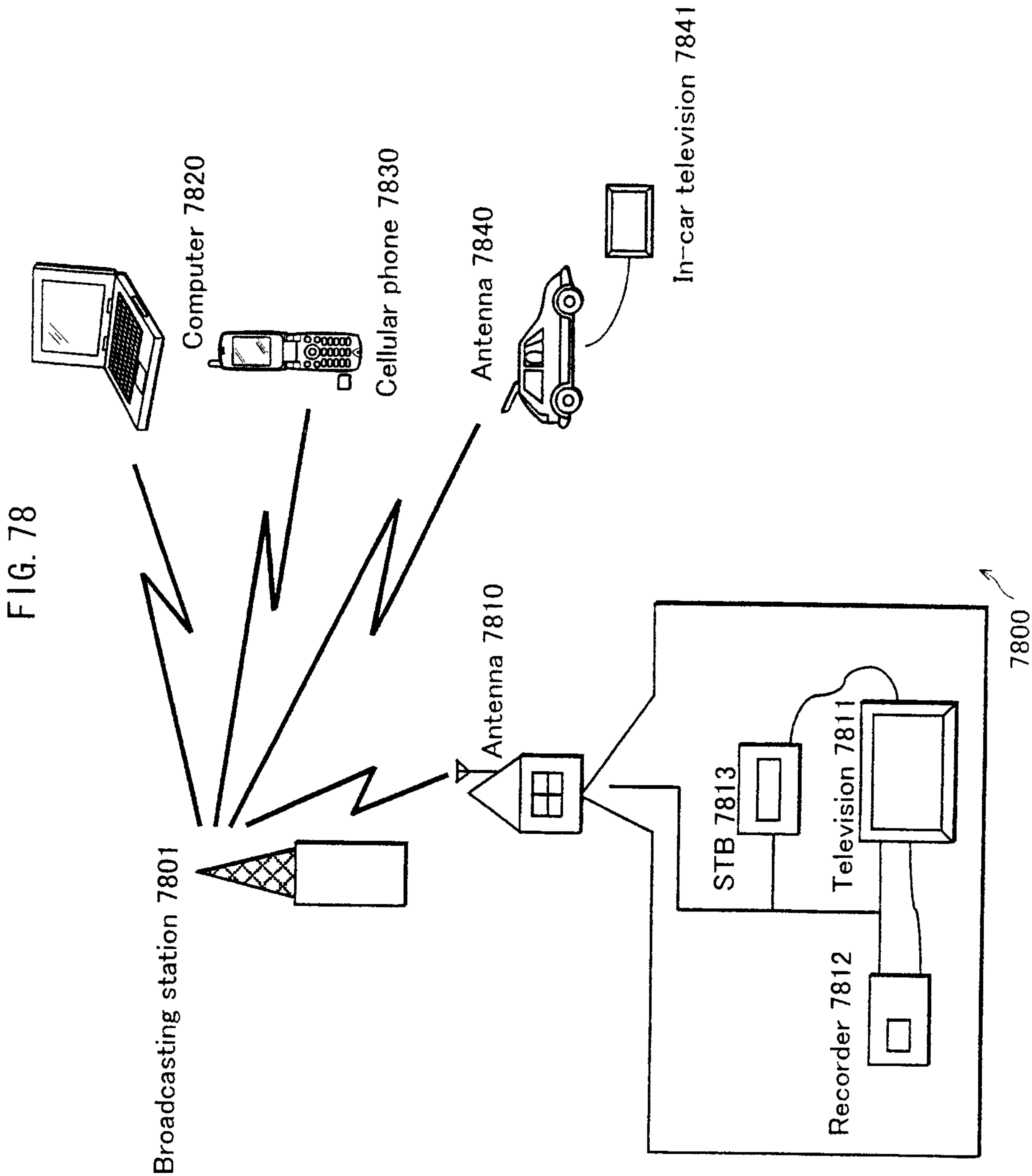


FIG. 77





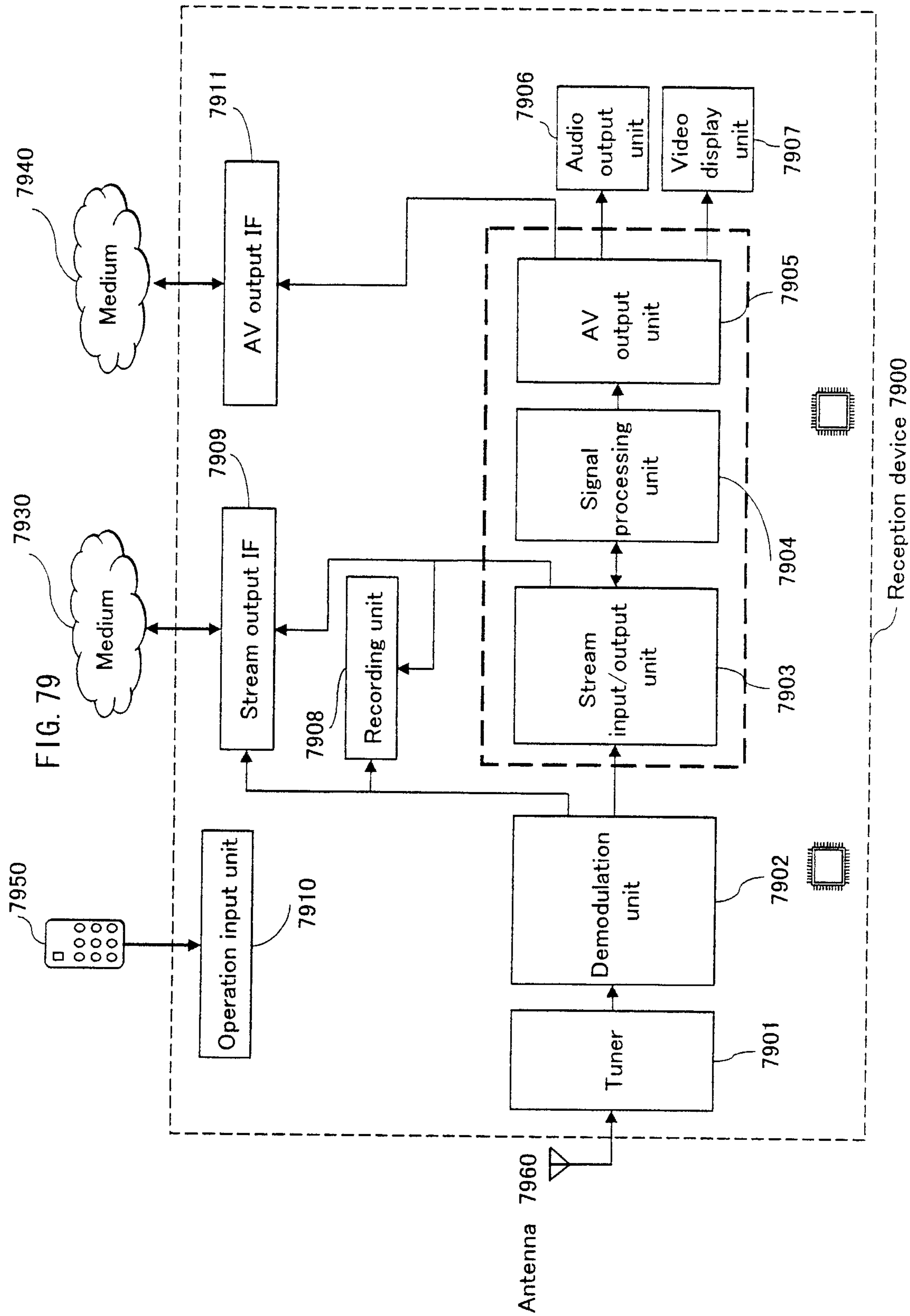


FIG. 80

Video stream (PID=0x1011 primary video)
Audio stream (PID=0x1100)
Audio stream (PID=0x1101)
Presentation graphics stream (PID=0x1200)
Presentation graphics stream (PID=0x1201)
Interactive graphics stream (PID=0x1400)
Video stream (PID=0x1B00 secondary video)
Video stream (PID=0x1B01 secondary video)

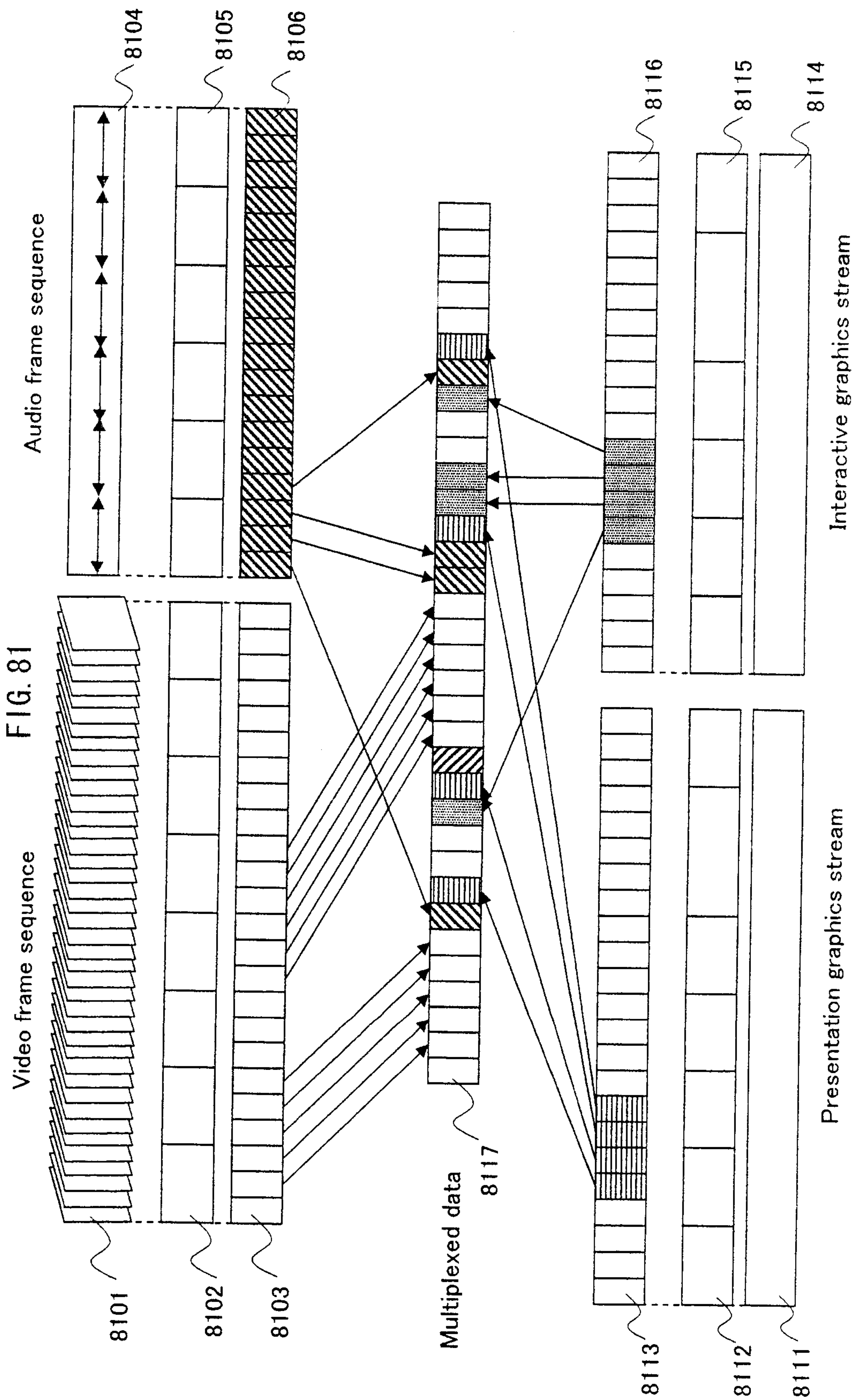


FIG. 82

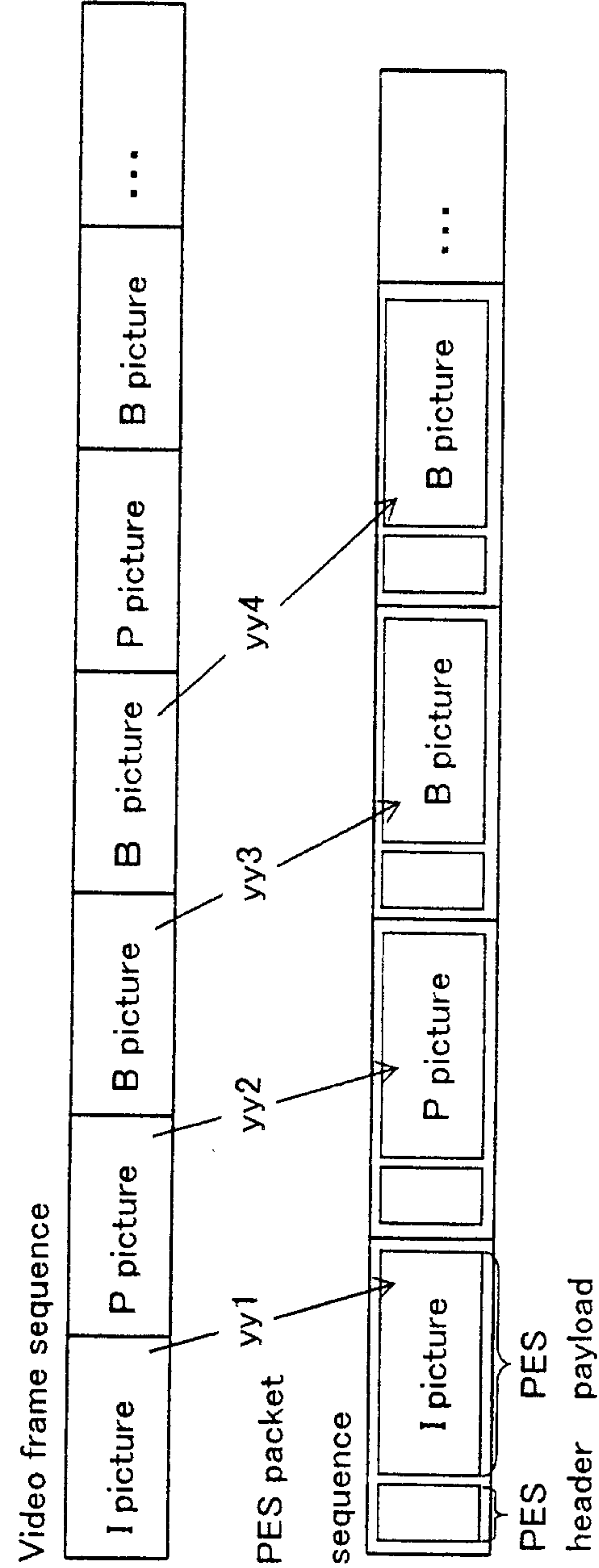


FIG. 83

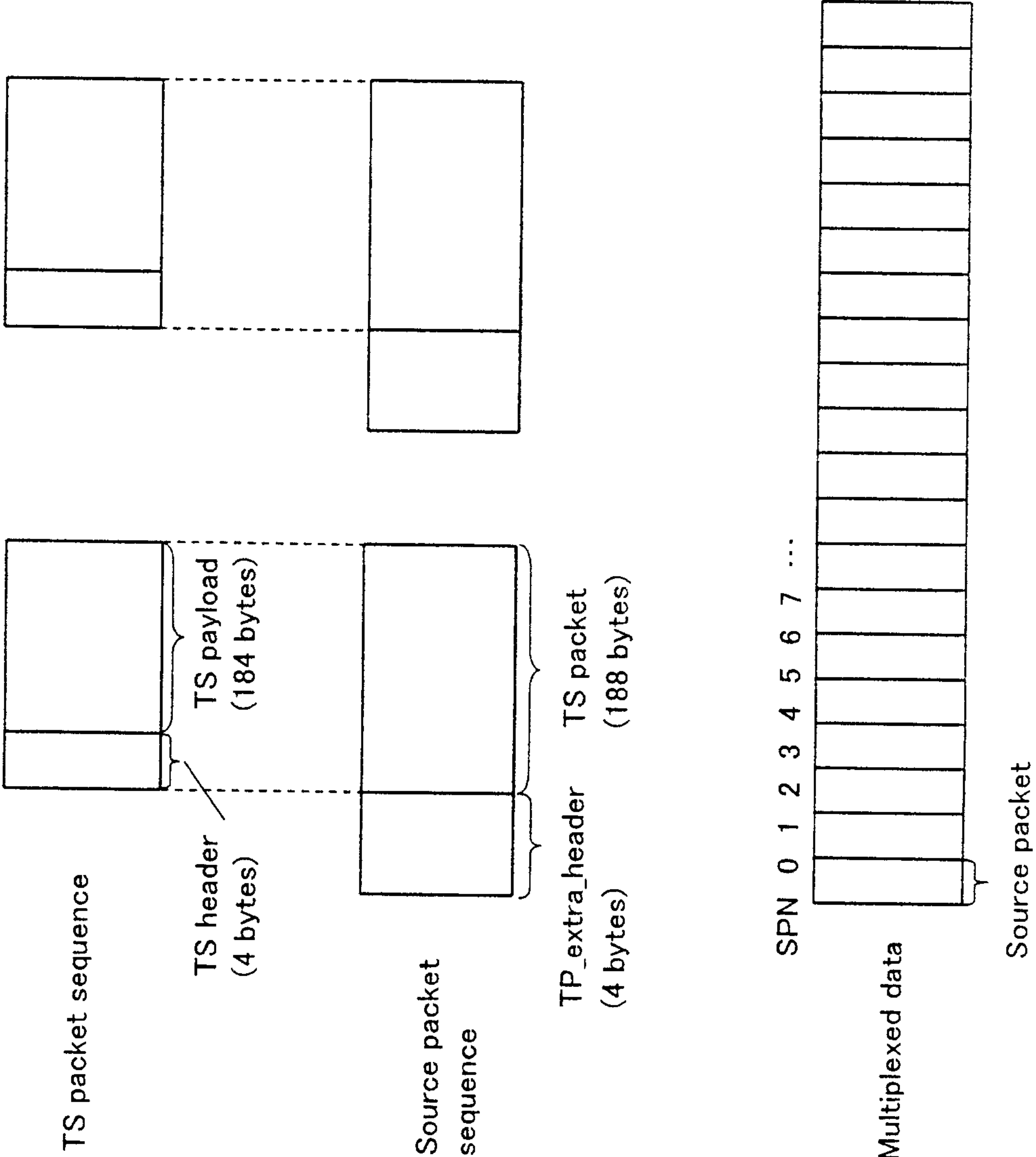


FIG. 84

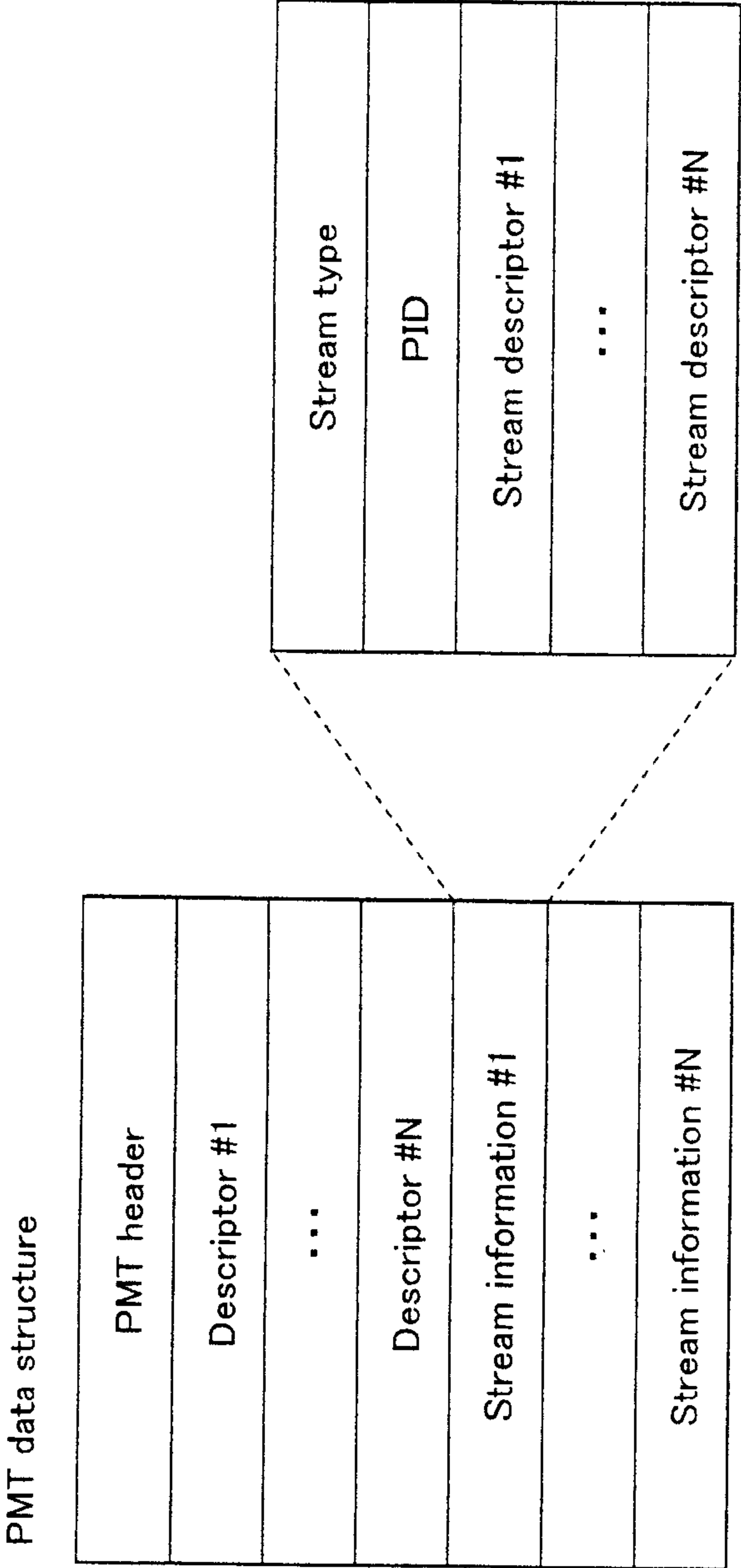


FIG. 85

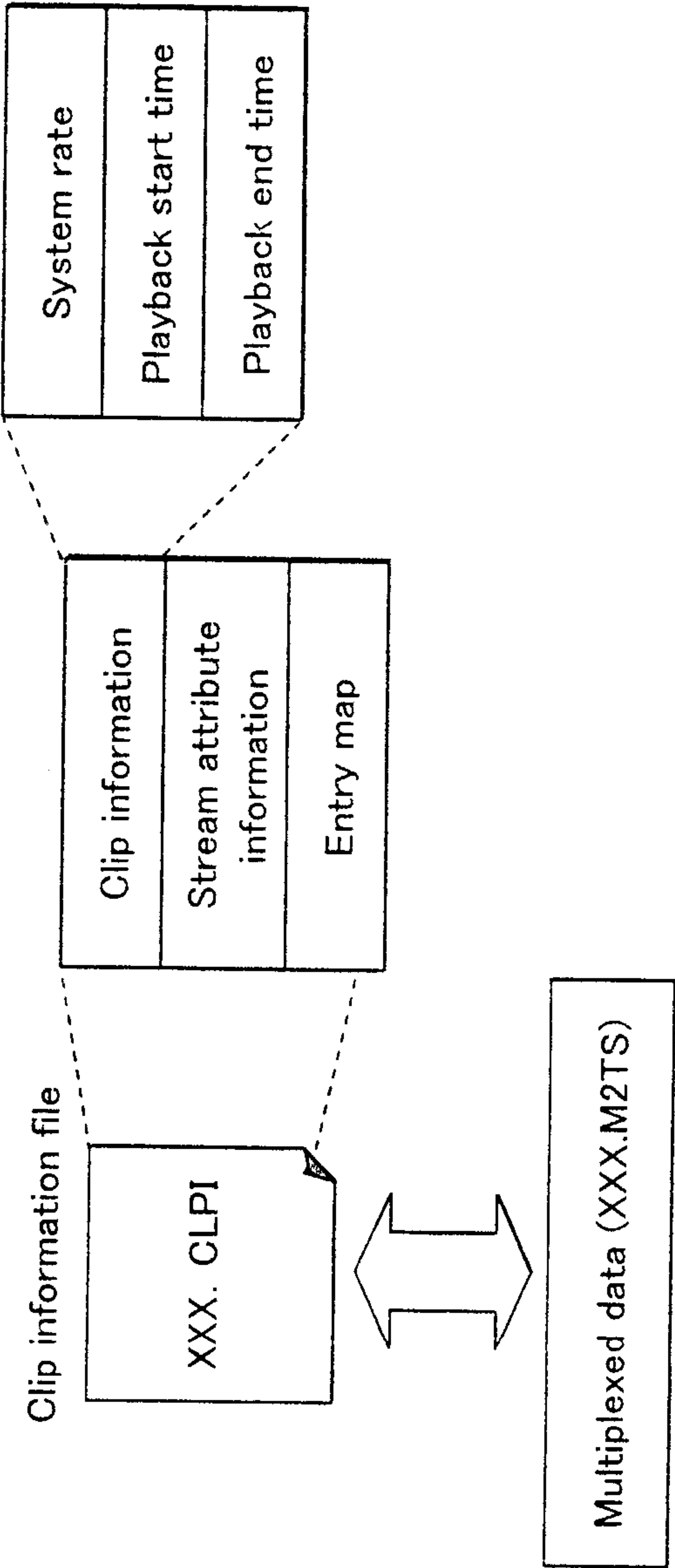
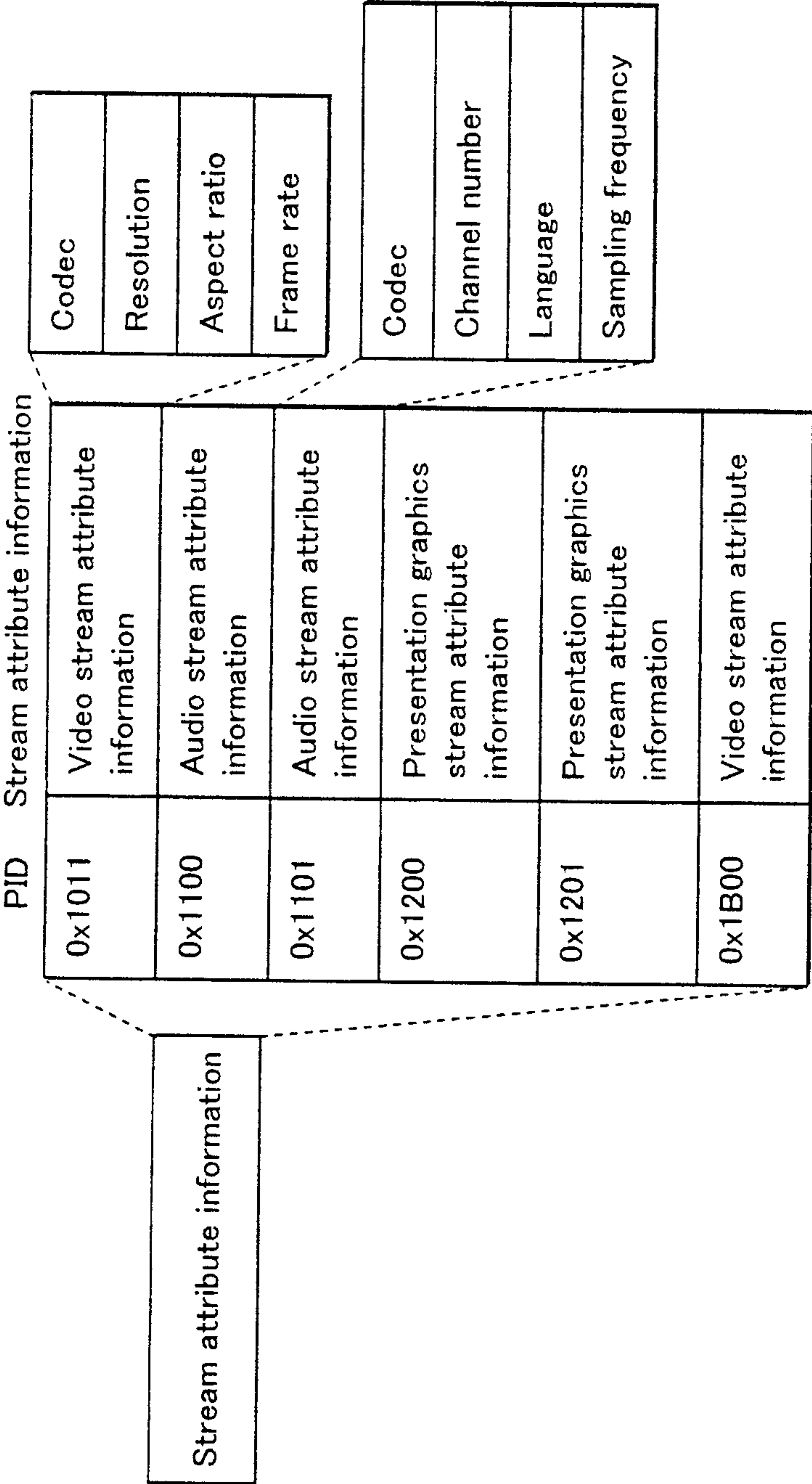
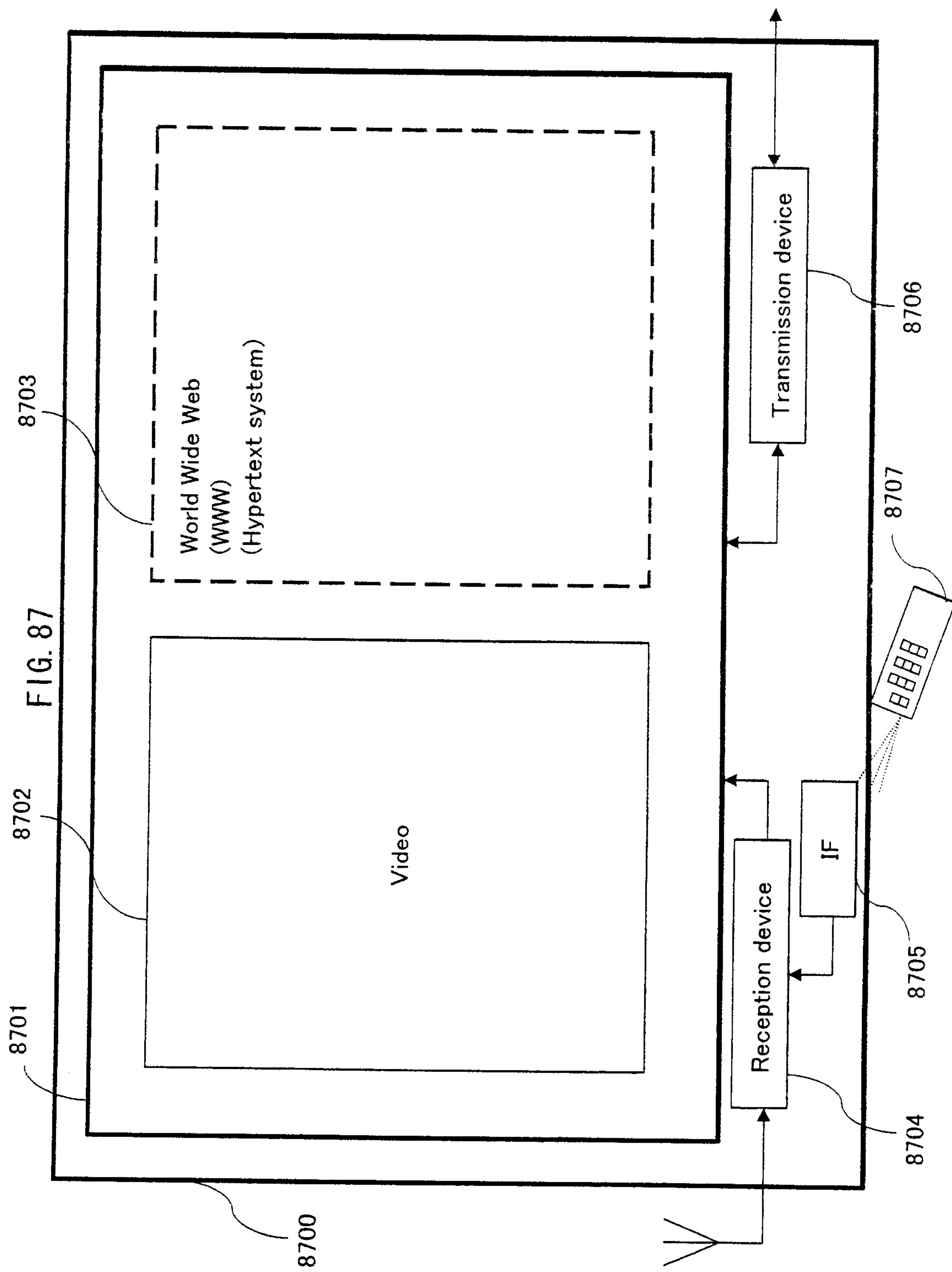


FIG. 86





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FIG. 88

