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(54) **Phase difference correcting demodulator for a receiver for spread spectrum communication and method of demodulating**

Phasendifferenzkorrekturdemodulator eines Spreizspektrumkommunikationsempfängers und Demodulationsverfahren

Demodulateur correctif de la différence de phase dans un récepteur pour la communication par spectre étalé et procédé de démodulation

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Description

BACKGROUND OF THE INVENTION

1. Field of the Invention

[0001] This invention relates to a data demodulator of a receiving apparatus for spread spectrum communication and more particularly to a data demodulator which removes a phase difference remaining after detection by using pilot signals not data-modulated transmitted from a base station or cell-site for improving signal quality.

2. Description of the Related Art

[0002] A spread spectrum communication system of direkt-sequence technique, which has advantages such as good resistance to interference and a property hard to give interference, is developed as one of communication systems for small capacity communication using communication satellite and mobile communication such as mobile phones, portable phones, or cordless phones.

[0003] Figure 11 shows the schematic configuration of a receiving apparatus of a CDMA (code division multiple access) cellular telephone system disclosed in US Pat. No.5,103,459. The mobile unit CDMA telephone system contains an antenna 1 for connection through a duplexer 2 to an analog receiver 3. The antenna 1 receives spread spectrum communication signals from base stations or cell-sites and feeds the received signals via the duplexer 2 into the analog receiver 3. The analog receiver 3, which contains a down converter and analog-to-digital converter, converts (or detects) the fed signals into base band signals by the down converter and further converts the base band signals to digital signals by the analog-to-digital converter. The base band signals converted to the digital signals are fed into a searcher receiver 5 and digital data receivers (data demodulators) 6 and 7.

[0004] When spread spectrum communication signal arrive at the receiving apparatus through a plurality of paths, a difference occurs in the reception time for each signal of paths. The data demodulators 6 and 7 can select which signal of paths is to be received and tracked on respectively. If two data demodulators are installed as shown in Figure 11, two independent paths can be tracked in parallel.

[0005] On the other hand, in response to a control signal from a control processor 8, the searcher receiver 5 scans the time domain around the nominal time of received pilot signals to detect pilot signals contained in each received multipath signals from cell-sites. The searcher receiver 5 compares the strength of one received pilot signal with that of another, and outputs the strength signal to the control processor 8 to indicate the strongest signal.

[0006] The control processor 8 provides control signals to the data demodulators 6 and 7 for each to process a different one of the strongest signals.

[0007] The function of each of the data demodulators 6 and 7 is to correlate received signals with PN codes used in transmitting part at cell-sites. Figure 12 shows the details of data demodulator disclosed in US Pat. No.5,103,459. Each of the data demodulators 6 and 7 contains PN generators 516 and 518 which generate PN codes PNI(t) and PNQ(t) for the in-phase axis and quadrature axis respectively corresponding to received path signals. The data demodulator 6, 7 also contains a Walsh function generator 520 generating the Walsh function appropriate for the cell-site to communicate with the mobile unit. The Walsh function generator 520 generates a code sequence corresponding to a Walsh function assigned in response to a select signal from the control processor 8. The select signal is transmitted by the cell-site to the mobile unit as a part of a call setup message. Outputs of the PN generators 516 and 518, PN codes PNI(t) and PNQ(t), are input to exclusive-OR gates 522 and 524 respectively. The Walsh function generator 520 supplies its output to tie exclusive-OR gates 522 and 524 where the signals are then exclusive-OR'ed together to generate sequences PNI'(t) and PNQ'(t).

[0008] The sequences PNI'(t) and PNQ'(t) are input to a PN QPSK correlator 526 for processing, and outputs of the PN QPSK correlator 526, I and Q, are fed into accumulators 528 and 530 respectively. The accumulators 528 and 530 integrate (accumulate and add) the input signals over the 1-symbol time. As a result, the correlation between PNI'(t) and in-phase axis received signal and that between PNQ'(t) and quadrature axis received signal are calculated by the PN QPSK correlators 526 and the accumulators. The accumulator outputs are input to a phase rotator 532. The phase rotator 532 also receives a pilot phase signal from the control processor 8. The phase of receive symbol data is rotated according to the phase of the pilot signal. The pilot signal phase is determined by the searcher receiver and the control processor. The output of the phase rotator 532, data on in-phase axis, is supplied to a combiner and decoder circuit.

[0009] With the conventional receiving apparatus, the analog receiver which down converts (or detects) received signals into base band signals and further converts into digital signals processes the signals passed through all paths in common, as described above. However, the received signals passed through the paths have carrier phases independent of each other. If the receive signals are passed through a single path, the phase of the received signal can be controlled by a carrier recovery circuit, but if the received signals are passed through a plurality of paths, their phases cannot be controlled because of plurality of independent carrier phases. Therefore, inevitably the input signals to each digital data receiver include the carrier phase difference between a received path signal and recovered carrier using for down converting (so called phase difference remaining after detection). When the phase differences exist, received signal components of in-phase axis and quadrature axis mixes with each other.

[0010] As with the communication system disclosed in US Pat. No.5,103,459, assume that data modulation and Walsh function modulation for user identification are bi-phase shift keying (BPSK) and spread modulation is quadrature phase shift keying (QPSK). Complex envelope of transmitted signal, S(t), is

$$S(t) = W(t)[PNI(t) + jPNQ(t)]$$

where W(t) is a multiplex signal of the transmit signals and pilot signals to each user. Assuming that modulation data to the ith user is di(t), the Walsh function is Wi(t), and the number of multiplexed signals is N,

$$W(t) = \Sigma di(t)\ Wi(t)$$

where i = 1 to N.

Next, assume that the reception amplitude (envelope) of a received path signal is ρ and the phase difference between the carrier of the received path signal and the recovered carrier multiplied at the analog receiver for down converting (carrier phase difference remaining after detection) is θ. The complex envelope of the received path signal component to be demodulated including in the output of analog receiver is

$$Rx(t) = \rho S(t) * \exp(j\theta)$$

$$= \rho W(t)\ [PNI\ (t) + j\ PNQ\ (t)]\ [\cos\theta + j \sin\theta]$$

$$= \rho W(t)\ [\{PNI(t) \cos\theta - PNQ\ (t) \sin\theta\} + j\ \{PNI\ (t) \sin\theta + PNQ\ (t) \cos\theta\}].$$

That is, the in-phase axis received signal is $\rho W(t)\ \{PNI(t) \cos\theta - PNQ\ (t) \sin\theta\}$, and the quadrature axis received signal is $\rho W(t)\ \{PNI(t) \sin\theta + PNQ(t) \cos\theta\}$. Thus, the in-phase axis received signal and quadrature axis received signal include a different signal component with each other (a component related to PNQ(t) in the in-phase axis and a component related to PNI(t) in the quadrature axis). Therefore, compensation processing is required. Formerly, for example, a PN QPSK correlator as shown in Figure 13 is provided with multipliers which multiply in-phase axis and quadrature axis received signals by PN codes of both in-phase and quadrature axes, and the multiplier outputs are added in a predetermined combination.

[0011] At the PN QPSK correlator in Figure 13, each of the in-phase axis and quadrature axis received signals is multiplied by the PN code PNI(t) for the in-phase axis and the PN code PNQ(t) for the quadrature axis, and the results are added together in combinations shown in Figure 13. That is, output I is

$$I = \rho W(t)\ [PNI'(t)\ \{PNI(t) \cos\theta - PNQ\ (t) \sin\theta\ + PNQ'(t)\ \{PNI\ (t) \sin\theta + PNQ\ (t) \cos\theta\}]$$

Output Q is

$$Q = \rho W(t)\ [-PNQ'(t)\ \{PNI(t) \cos\theta - PNQ\ (t) \sin\theta\ + PNI'(t)\ \{PHI\ (t) \sin\theta + PNQ\ (t) \cos\theta\}]$$

The outputs I and Q are integrated by the accumulators 528 and 530 respectively over the symbol time. Of the integration results, only the component of di(t) multiplied by Wi(t) contained in PNI', PNQ' in the multiplexed signals remains due to orthogonality of the Walsh function. For example, assuming that the symbol time is T, the following relationship becomes true:

$$\int PNI'(t)\ PNI(t)\ W(t)\ dt$$

$$= \int [PNI(t)\ Wi(t)]\ PNI(t)\ W(t)\ dt$$

$$= \int [PNI(t)\ Wi(t)]\ PNI(t)\ [\Sigma di(t)\ Wi(t)]dt$$

$$= \int PNI^2(t)\ Wi(t)\ [\Sigma di(t)\ Wi(t)]dt$$

$$= \int Wi(t)\ [\Sigma di(t)\ Wi(t)]dt$$

$$= Tki \cdot di(t)$$

where ki is a ratio constant related to the power allocation percentage of the multiplex signal. Therefore, the outputs of the accumulators 528 and 530 become $2\rho ki \cdot di(t) \cos \Theta$ and $2\rho ki \cdot di(t) \sin \Theta$ respectively. This assumes that the correlation processing timing is given by a timing recovery circuit and that the cross-correlation value between PNI(t) and PNQ(t) is sufficiently small due to one of the PN code characteristics and may be ignored by correlation processing. Now in-phase axis and quadrature axis received signals are separated efficiently, but the effect of $\cos \Theta$ remains in the output of the accumulator 528 and that of $\sin \theta$ remains in the output of the accumulator 530. To remove these effects, for example, calculation of $\theta = \tan^{-1}$ (Q/I) is executed and phase rotation operation is performed in response to the resultant θ, thereby providing $2\rho\ ki \cdot di(t)$. However, complicated steps of calculation of $\tan^{-1}$ to estimate θ and the phase rotation operation are required.

[0012] The data demodulator requires a timing recovery circuit (not shown in the conventional example) to provide timing to correlation processing. Generally the timing recovery circuit is constructed with DLL (delay-locked loop), etc.,; the correlation pulse level corresponding to the correlation processing timing must be obtained at the DLL. To obtain the correlation pulse level from the circuit configuration in Figure 13, the square sum of the outputs of the accumulators 528 and 530 is required for removing uncertainties of phase difference Θ and data di(t). With such operation,

$$8\rho^2 ki^2 \cdot di^2(t)\ [\cos^2\theta + \sin^2\theta] = 8\rho^2 ki^2 \cdot di^2(t)$$

is obtained, and by integrating over the data demodulation interval time, a component corresponding to the power of correlation pulse is obtained. However, in this method, noise contained separately in both the in-phase axis and quadrature axis received signals mix with each other by the square operation and the noise effect becomes greater and it degrades the timing recovery characteristic. In order to avoid square sum operation, correlation pulses of pilot signals which are not data-modulated may be used after the effect of phase difference is removed.

[0013] However, in the conventional configuration, complicated processing of calculation of $\tan^{-1}$ to estimate Θ and phase rotation operation is required. To use the correlation pulses of pilot signals at the DLL, the sequences PNI'(t) and PNQ'(t) used at the PN QPSK correlator must be generated from the Walsh function corresponding to pilot signal and PN code; another PN QPSK correlator for the DLL is required in addition to the PN QPSK correlator for data demodulation. Further, since correlation processing must be performed at the timings slightly shifted before and after from the data demodulation timing at the DLL, additional two systems for such complicated processing are required in addition to the data demodulation system; an enormous amount of operations must be performed.

[0014] Thus, the data demodulator of the conventional receiving apparatus for spread spectrum communication has a problem of complicated processing required to remove the effect of the phase difference remaining after detection. Timing reproduction also requires either the processing of the square sum of PN QPSK correlator output or the processing of phase correction; if the square sum is processed, the noise effect degrades the timing recovery characteristic or if phase correction is processed, complicated operation is required.

SUMMARY OF THE INVENTION

[0015] This object is solved by the method with the features of claim 1 and by a demodulator using the method of claim 1. Using the method of the invention the effect of the phase difference contained in in-phase axis and quadrature axis received signals is removed, and the received signal from which the phase difference effect is removed is used for data demodulation. At timing recovery, the correlation pulse component of pilot signal is also provided by a simple configuration without increasing the noise effect.

BRIEF DESCRIPTION OF THE DRAWINGS

[0016] In the accompanying drawings:

Figure 1 is a block diagram of a first embodiment of the phase difference compensation circuit being part of the data demodulator of the invention;

Figure 2 is a block diagram of a second embodiment of the phase difference compensation circuit being part of the data demodulation of the invention;

Figure 3 is a block diagram of a third embodiment of the phase difference compensation circuit being part of the data demodulator of the invention;

Figure 4 s a block diagram of a fourth embodiment of the phase difference compensation circuit being part of the data demodulator of the invention;

Figure 5 is a block diagram of a fifth embodiment of the phase difference compensation circuit being part of the data demodulator of the invention;

Figure 6 is a block diagram of each averaging section in the first to fifth embodiments of the phase difference compensation circuit;

Figure 7 is a block diagram of a first embodiment of the DLL being part of the data demodulator of the invention;

Figure 8 is a block diagram of another embodiment of the DLL being part of the data demodulator of the invention;

Figure 9 is a block diagram of a data demodulation section according to a first embodiment;

Figure 10 is a block diagram of a data demodulation section according to another embodiment;

Figure 11 is a block diagram of a conventional receiving apparatus;

Figure 12 is a block diagram of a data demodulator in the conventional receiving apparatus; and

Figure 13 is an illustration showing data demodulation principle in the conventional receiving apparatus.

DESCRIPTION OF THE PREFERRED EMBODIMENTS

[0017] Referring now to the accompanying drawings, there are shown preferred embodiments of a data demodulator of a receiving apparatus for spread spectrum communication according to the invention.

[0018] Figure 1 shows the configuration of a phase difference compensation circuit according to a first embodiment. The received signals converted into a digital signal by an A/D converter of analog receiver 3 is fed into the phase difference compensation circuit according to the embodiment. The phase difference compensation circuit is provided with a correlation calculation section which calculates the correlation between an in-phase axis received signal and in-phase axis PN code and that between a quadrature axis received signal and quadrature axis PN code. That is, an in-phase axis received signal given from an in-phase axis A/D converter is multiplied by in-phase axis PN code PNI(t) and averaged by an averaging section meanA 20 for calculating the correlation. The result is then multiplied by in-phase axis received signal given from the in-phase axis A/D converter. In-phase axis received signal given from the in-phase axis A/D converter is multiplied by quadrature axis PN code PNQ(t) and averaged by an averaging section meanA 22, then is multiplied by in-phase axis received signal given from the in-phase axis A/D converter. On the other hand, a quadrature axis received signal given from a quadrature axis A/D converter, like the in-phase axis signal described above, is multiplied by PNI(t), PNQ(t) and averaged by averaging section meanA 24, 26, then is multiplied by the original quadrature axis received signal for output. The output of the multiplier 21 and the output of the multiplier 30 are added together by an adder 32 for output. The output of the multiplier 23 and the output of the multiplier 28 are added together by an adder 34 for output. Then, phase difference compensation is executed.

[0019] The phase difference compensation circuit according to the embodiment has the configuration described above. The operation of the phase difference compensation circuit is as follows. The received path signal component supplied from the analog receiver 3 is

$$\rho W(t)\,(PNI(t)\cos\Theta - PNQ(t)\sin\Theta) + j\rho w(t)\,(PNQ(t)\cos\Theta + PNI(t)\sin\Theta)$$

where in-phase axis received signal is the first term on the right-hand side of the equation and quadrature axis received signal is the second term on the right-hand side of the equation. Therefore, when the in-phase axis received signal $\rho W(t)\,(PNI(t)\cos\Theta - PNQ(t)\sin\Theta)$ is multiplied by PNI(t) and further averaged by the averaging section meanA 20, output is

$$(1/T)\int W(t)\,(PNI(t)\cdot PNI(t)\cos\Theta - PNI(t)\,PNQ(t)\sin\Theta)\,dt = \rho k0\cos\Theta$$

because of orthogonality of the Walsh function, in which $(1/T)\int (PNI(t)\,PNI(t))\,dt = 1$, and $(1/T)\int (PNI(t)\,PNQ(t))\,dt$ may be sufficiently small by averaging processing. Although the analog receiver output often contains another received path signal component, the timings of the PN codes multiplied them have different timing, thus it may be sufficiently small by averaging processing. k0 is a proportional constant corresponding to the power allocation percentage to the pilot channel (W0(t): All 1).

[0020] When in-phase axis received signal is multiplied by PNQ(t) and further averaged by the averaging section meanA 22, the resultant signal is

$$(1/T) \int \rho W(t) (PNQ(t)\, PNI(t) \cos \Theta - PNQ(t)\, PNQ(t) \sin \Theta)\, dt$$

$$= -\rho k0 \sin \Theta$$

where $(1/T) \int (PNQ(t)\, PNQ(t))\, dt=1$. Likewise, for the quadrature axis received signal, output from the averaging section meanA 24 is

$$(1/T) \int \rho W(t) (PNI(t)\, PNQ(t) \cos \Theta + PNI(t)\, PNI(t) \sin \Theta)\, dt$$

$$= \rho k0 \sin \Theta$$

Output from the averaging section meanA 26 is

$$(1/T) \int \rho W(t) (PNQ(t)\, PNQ(t) \cos \Theta + PNQ(t)\, PNI(t) \sin \Theta)\, dt$$

$$= \rho k0 \cos \Theta$$

In-phase and quadrature axis received signals are multiplied by the $\rho k0 \cos \Theta$, $-\rho k0 \sin \Theta$, $\rho k0 \sin \Theta$, and $\rho k0 \cos \Theta$ by the multipliers 21, 28, 30, and 23 respectively. As described above, if phase difference Θ exists, receive signal is $\rho W(t)\, [PNI(t) + j\, PNQ(t)] \exp [j \Theta]$. Thus, this signal may be multiplied by $\exp [-j \Theta]$ to remove the phase difference. That is,

$$\rho W(t)\, [(PNI(t) \cos \Theta - PNQ(t) \sin \Theta) + j\, (PNQ(t) \cos \Theta + PNI(t) \sin \Theta)]\, (\cos \Theta - j \sin \Theta)$$

$$= \rho W(t)\, [\cos \Theta\, (PNI(t) \cos \Theta - PNQ(t) \sin \Theta) + \sin \Theta\, (PNQ(t) \cos \Theta + PNI(t) \sin \Theta)]$$

$$+ j\, \rho W(t)\, [(-\sin \Theta\, (PNI(t) \cos \Theta - PNQ(t) \sin \Theta) + \cos \Theta\, (PNQ(t) \cos \Theta + PNI(t) \sin \Theta)]$$

$$= \rho W(t)\, PNI(t) + j \rho W(t)\, PNQ(t)$$

[0021] Considering the right-hand side of the equation, the first term is the sum of the term of multiplying in-phase axis received signal by $\cos \Theta$ and the term of multiplying quadrature axis received signal by $\sin \Theta$, and the second term is the sum of the term of multiplying in-phase axis received signal by $\sin \Theta$ and the term of multiplying quadrature axis received signal by $\cos \Theta$. On the other hand, as described above, meanA 20, 22, 24, and 26 output $\rho k0 \cos \Theta$, $-\rho k0 \sin \Theta$, $\rho k0 \sin \Theta$, and $\rho k0 \cos \Theta$ respectively. Therefore, in-phase and quadrature axis received signals are multiplied by the outputs from the averaging sections meanA 20, 22, 24, and 26, and then added appropriately so as to satisfy the above-mentioned equation, thereby removing the phase difference Θ.

[0022] That is, in-phase axis received signal is multiplied by the output $\rho k0 \cos \Theta$ from meanA 20 by the multiplier 21 and quadrature axis received signal is multiplied by the output $\rho k0 \sin \Theta$ from meanA 24 by the multiplier 30, then the outputs of the multipliers 21 and 30 are added together by the adder 32, thereby enabling signal processing equivalent to the first term on the right of the equation.

[0023] Likewise, quadrature axis received signal is multiplied by the output $\rho k0 \cos \Theta$ from meanA 26 by the multiplier 23 and in-phased axis received signal is multiplied by the output $-\rho k0 \sin \Theta$ from meanA 22 by the multiplier 28, then the outputs of the multipliers 23 and 28 are added together by the adder 34, thereby enabling signal processing equivalent to the second term on the right of the equation. Thus, the in-phase and quadrature axis signals with no phase difference, $\rho k0 \cdot \rho W(t)\, PNI(t)$ and $\rho k0 \cdot \rho W(t)\, PNQ(t)$, can be obtained from the in-phase and quadrature axis received signals.

[0024] Each of this is the value multiplied by ρko to the each of desired value, where k0 is constant and ρ is useful for maximal ratio combining at the combiner and decoder.

[0025] Figures 2 to 6 show phase compensation circuits according to other embodiments. Each of outputs of the averaging sections 20 and 26 of the phase compensation circuit shown in Figure 1 contains $\rho k0 \cos \Theta$. On the other hand, outputs of the averaging sections 22 and 24 contain $-\rho k0 \sin \Theta$ and $\rho k0 \sin \Theta$. Therefore, the phase compensation function is provided by one system which finds the $\cos \Theta$ component and one system which finds the $\sin \Theta$ component considering the polarity.

[0026] Figure 2 shows a configuration in which $\cos \Theta$ and $\sin \Theta$ components are found from an in-phase axis received signal and phase compensation is executed. Figure 3 shows a configuration in which $\cos \Theta$ and $\sin \Theta$ components are found from a quadrature axis received signal and phase compensation is executed. Figure 4 shows a configuration in which the $\cos \Theta$ component is found from an in-phase axis received signal and the $\sin \Theta$ component is found from a quadrature axis received signal. Figure 5 shows a configuration in which the $\cos \Theta$ component is found from a quadrature axis received signal and the $\sin \Theta$ component is found from an in-phase axis received signal. To consider the polarity of the $\sin \Theta$ component, one input to the adder 32 has the negative polarity in Figures 2 and 5 and one input to the adder 34 has the negative polarity in Figures 3 and 4.

[0027] Figure 6 shows a preferred example of the averaging sections 20, 22, 24, and 26 in Figures 1 to 5. An input is first fed into an accumulator 201 and integrated (accumulated and added) over the 1-symbol time, then the integration result is output every 1-symbol time. The accumulator output is fed into a recursive adding section which consists of multipliers 202 and 205, an adder 203, and a delay circuit 204 where recursive addition (accumulative addition with weighting) is performed to remove the noise effect. The delay time of the delay circuit is the 1-symbol time T and r input to the multiplier 205 ($0 \leq r < 1$), which is a weight, indicates the averaging degree by recursive addition and is set properly depending on the link condition. 1-r input to the multiplier 202 is a normalization constant used to keep the same power between input and output of the recursive adding section.

[0028] Figure 7 shows the configuration of a DLL (delay lock loop) according to the first embodiment Both in-phase and quadrature axis received signals from which the phase difference is removed by the phase difference compensation circuit described above are fed into the DLL. The in-phase axis received signal is multiplied by codes provided with timing shift of PNI(t) before and after by Δ, PNI $(t - \Delta)$ and PHI $(t + \Delta)$. Likewise, the quadrature axis received signal is multiplied by codes provided with timing shift of PNQ(t) before and after by Δ, PNQ $(t - \Delta)$ and PNQ $(t + \Delta)$. Then, the multiplication results are added together with the polarities shown in Figure 7 and further averaged by an averaging section meanB, then input to a timing controller. Where averaging section meanB is constructed by Fig. 6 or some kind of loop filter. The timing controller outputs a timing signal so that the signal from the averaging section meanB becomes zero. The timing signal is used for the generation timing of PN codes in Figure 1 and for a symbol clock of a data demodulator (described below) via a divider, etc. In addition, this timing signal is also supplied to the control processor 8 for comparison with the timing of a strength signal given from the searcher receiver 5 for control of a plurality of data demodulators to always executing demodulation to the optimum path (strong signal path); it is also used for the diversity combining timing at the combiner and decoder circuit 9.

[0029] Figure 8 shows a DLL according to another embodiment. This DLL differs from the DLL shown in Figure 7 in that it has a multiplier between an adder and averaging section B. The phase difference effect is removed from the in-phase and quadrature axis components by the phase compensation circuit and at the same time, the amplitude is multiplied by $\rho k0$, and further is multiplied by $\rho k0$ by the averaging operation of the averaging section B. Thus, the multiplier multiplies an adder output by $1/(\rho k0)^2$, thereby compensating fluctuation of the input level of the DLL due to fluctuation of the amplitude ρ of a reception path caused by fading. Since the input level fluctuation of the DLL becomes fluctuation of loop gains, stable operation is enabled by compensation of the input level fluctuation by the multiplier.

[0030] The factor of $(\rho k0)^2$ is a coefficient corresponding to reception power of pilot signal, and can be provided by the data demodulation section described below.

[0031] Figure 9 shows the configuration of a data demodulation section according to the first embodiment of the invention. The received base band signals from which the phase difference is removed by the phase difference compensation circuit described above are fed into the data demodulation section. The supplied in-phase and quadrature axis signals $\rho k0 \cdot \rho W(t)$ PNI(t) and $\rho k0 \cdot \rho W(t)$ PNQ(t) are sent to multipliers 40 and 42 respectively for multiplication by PNI (t) and PNQ (t) respectively, and the effect of PN codes are removed, then both becomes to be $\rho k0 \rho W(t)$. That is, since they are multiplied at the same timing, the PN code effect is removed. The subsequent data demodulation circuit operation is to solve Walsh functions for data demodulation. Outputs from the multipliers 40 and 42 are sent to an adder 44 for addition of the in-phase and quadrature axis signals and output of the result. This step is performed to combine the same signals appearing on both channels when the effect of PN codes are removed. To improve the DLL resolution, the signals from the analog receiver may be oversampled to the chip rate. That is, a single chip may be transmitted consecutively a given number of times, such as four. To deal with such oversampling, the data demodulation section according to the embodiment is provided with a 1/4 serial-parallel converter 46 and an adder 48 for restoring the overlapping chip samples to the original one chip symbol which is then sent to a 1/64 serial-parallel converter 50 at the following stage. Here, the method of adding a sample value for oversampling, but a method of extracting only one sample every four samples is also possible. The 1/64 serial-parallel converter 50 converts the input signal into parallel data of 64 chip symbols in response to symbol clock and sends the results to an FHT processor 52. The FHT processor 52 fast Hadamard transforms the received 64 chip symbol data for channel separation, then outputs correlation signals for Walsh codes W0 to W63 to a selector 54. The selector 54 selects correlation signal $2\rho^2k0kidi(t)$ related to desired Walsh code Wi in response to the select signal supplied from the control processor 3, and sends the correlation signal to the diversity circuit, etc., for data demodulation. In the data demodulation section, synchronous tracking is performed by the pilot signal as described above, and the FHT processor 52 can be operated separately from the synchronous tracking system. Thus, it needs to be operated only at the data timing, and consumption power can be reduced. While the FHT processor 52 is outputting the correlation signals, the correlation signal for W0 becomes $2\rho^2k0^2$ which can be used as an input to the multiplier in Figure 8.

[0032] Without using the FHT processor 52 at the data demodulation section according to the embodiment, data demodulation can also be executed by using a correlator which uses Walsh function generated by a Walsh function generator according to a select signal supplied from the control processor as a reference sequence as shown in Figure 10. An output from an adder 48 is fed into a multiplier 58 for multiplication by Walsh code Wi (t) assigned at the Walsh function generator. The result is accumulated and added by an accumulator 60 to provide $2\rho^2k0kidi(t)$ which is then fed into the diversity circuit. This configuration enables power to be less consumed as compared with the use of the FHT processor 52.

[0033] An output of an accumulator 56 becomes the correlation signal for W0, $2\rho^2k0^2$, which can be used as an input to the multiplier in Figure 8.

[0034] The data demodulator of the receiving apparatus for spread spectrum communication can adopt a simple configuration to remove the phase difference effect for improvement of reception S/N ratio and low power comsumption.

Claims

1. Method of demodulating data for spread spectrum communication in which spread spectrum modulated signals for an in-phase axis and a quadrature axis in a direct spread system with in-phase axis and quadrature axis pseudonoise codes are received and data from the received signals are recovered, whereby the received signals are correlated with the pseudonoise codes,
characterized in that an in-phase axis and/or quadrature axis received signal are multiplied by a pseudonoise code corresponding to a pilot signal transmitted from a base station and multiplication results are averaged for calculating a correlation containing an information of phase difference remaining after detection and a phase difference contained in in-phase axis and quadrature axis received signals is compensated using the phase difference information obtained by the correlation process, and that the in-phase axis and quadrature axis received signals for which the phase difference is compensated are multiplied by in-phase axis and quadrature axis pseudonoise codes, respectively, corresponding to the pilot signals transmitted from the base station and a signal provided by adding the multiplication results together is demodulated.

2. Method as claimed in claim 1 **characterized in that** the signal is demodulated using a signals processing by fast Hadamard transform (FHT).

3. Method as claimed in claim 1 further **characterized by** the steps of multiplying the inphase axis received signal for which the phase difference is compensated by in-phase axis pseudonoise codes provided by shifting the timing of an in-phase axis pseudonoise code slightly before and after;

   multiplying the in-phase axis received signal for which the phase difference is compensated by in-phase axis pseudonoise codes provided by shifting the timing of an in-phase axis pseudonoise code slightly before and after;
   multiplying the quadrature axis received signal for which the phase difference is compensated by quadrature axis pseudonoise codes provided by shifting the timing of a quadrature axis pseudonoise code slightly before and after;
   adding the in-phase axis and quadrature axis received signals multiplied by the pseudonoise codes;
   averaging the added signals; whereby responsive to the operation result of the averaging procedure a timing signal is output.

4. Method as claimed in claim 3, **characterized by** further multiplying the resultant signals of the addition procedure by a coefficient equivalent to a reciprocal of reception power of received pilot correlation signals and outputting the multiplication result for averaging.

5. Method as claimed in claim 3 or 4, **characterized by** generating in-phase axis and quadrature axis pseudonoise codes in response to the timing signals.

6. Method as claimed in claim 3 or 4, **characterized in that** the demodulation procedure of the received signals is responsive to the timing signal.

7. Method according to one of claims 1 to 6, **characterized in that** the in-phase axis and quadrature axis received signals are multiplied by in-phase axis and quadrature axis pseudonoise codes.

8. Method according to one of claims 1 to 6, **characterized in that** the inphase-axis received signal is multiplied by in-phase axis and quadrature axis pseudonoise codes.

9. Method according to one of claims 1 to 6, **characterized in that** the said quadrature axis received signal is multiplied by in-phase axis and quadrature axis pseudonoise codes.

10. A data demodulator of a receiving apparatus carrying out the method of claim 1 including means adapted to receive an in-phase axis and quadrature axis received signal, correlation calculating means comprising first multiplication means adapted to multiply the in-phase axis and/or quadrature axis received signals by a pseudonoise code corresponding to a pilot signals transmitted from a base station and averaging means (20, 22, 24, 26) adapted to average the results of the multiplication for calculating a correlation containing an information of phase difference remaining after detection and a phase difference contained in in-phase axis and quadrature axis received signals, the demodulator further comprising phase difference compensation means (21, 23, 28, 30, 32, 34) for compensating an effect of a phase difference contained in in-phase and quadrature axis received signal using the phase difference and data demodulation means (40, 42, 44, 46, 48, 50, 52, 54, 56, 58, 60, 520) comprising second multiplication means (40, 42) adapted to multiply the in-phase axis and quadrature axis received signals, for which the effect of the phase difference is compensated by said phase difference compensation means, by in-phase axis and quadrature axis pseudonoise codes, respectively, corresponding to the pilot signal transmitted from the base station and addition means (44) adapted to demodulate a signal provided by adding the multiplication results together.

11. A data demodulator according to claim 10 for carrying out the method according to claim 3 further including third multiplication means adapted to multiply the phase difference compensated in-phase axis received signals by the time shifted in-phase axis pseudonoise codes,

fourth multiplication means adapted to multiply the phase difference compensated quadrature axis received signal by the time shifted quadrature axis pseudonoise codes;
further addition means adapted to add the results of the third and fourth multiplication means;
further averaging means for averaging the added signals;
timing control means for generating the timing signal.

12. A data demodulator according to claim 11 for carrying out the method of claim 4 including fifth multiplication means being disposed between said further addition means and said further averaging means.

Patentansprüche

1. Verfahren zum Demodulieren von Daten für die Spreizspektrumkommunikation, bei welchem spreizspektrummodulierte Signale für eine gleichphasige Achse und eine Querachse in einem direkten Spreizsystem mit Pseudorauschcodes für die gleichphasige Achse und die Querachse empfangen und Daten aus den empfangenen Signalen wiedergewonnen werden, wodurch die empfangenen Signale mit Pseudorauschcodes korreliert werden, **dadurch gekennzeichnet, daß** ein empfangenes Signal für eine gleichphasige Achse oder eine Querachse multipliziert wird mit einem Pseudorauschcode entsprechend einem von einer Basisstation übertragenen Pilotsignal und die Multiplikationsergebnisse einer Durchschnittswertbildung unterzogen werden zum Berechnen einer Korrelation enthaltend eine Information über die Phasendifferenz, die nach der Erfassung verbleibt, und eine Phasendifferenz, die in den empfangenen Signalen für eine gleichphasige Achse und eine Querachse enthalten ist, kompensiert wird unter Verwendung der durch den Korrelationsvorgang erhaltenen Phasendifferenzinformation, und daß die empfangenen Signale für eine gleichphasige Achse und eine Querachse, für welche die Phasendifferenz kompensiert ist, mit Pseudorauschcodes für eine gleichphasige Achse bzw. eine Querachse multipliziert werden, entsprechend den von der Basisstation übertragenen Pilotsignalen, und ein durch Addieren der Multiplikationsergebnisse miteinander erhaltenes Signal demoduliert wird.

2. Verfahren nach Anspruch 1, **dadurch gekennzeichnet, daß** das Signal unter Verwendung einer Signalverarbeitung durch schnelle Hadamard-Transformation (FHT) demoduliert wird.

3. Verfahren nach Anspruch 1, weiterhin **gekennzeichnet durch** die Schritte des Multiplizierens des empfangenen Signals für die gleichphasige Achse, für welches die Phasendifferenz kompensiert ist, mit Pseudorauschcodes für die gleichphasige Achse, die **durch** Verschieben des Zeitpunktes eines Pseudorauschcodes für die gleichphasige Achse leicht nach vorn und zurück erhalten wurden;

   Multiplizieren des empfangenen Signals für die gleichphasige Achse, für welches die Phasendifferenz kompensiert ist, mit Pseudorauschcodes für die gleichphasige Achse, die **durch** Verschieben des Zeitpunktes eines Pseudorauschcodes für die gleichphasige Achse leicht nach vorn und zurück erhalten wurden;
   Multiplizieren des empfangenen Signals für die Querachse, für welches die Phasendifferenz kompensiert ist, mit Pseudorauschcodes für die Querachse, die **durch** Verschieben des Zeitpunktes eines Pseudorauschcodes für die Querachse etwas nach vorn und zurück erhalten wurden;
   Addieren der mit den Pseudorauschcodes multiplizierten empfangenen Signale für die gleichphasige Achse und die Querachse;
   Durchschnittswertbildung für die addierten Signale;
   wodurch in Abhängigkeit von dem Operationsergebnis des Durchschnittswertbildungsvorgangs ein Zeitsignal ausgegeben wird.

4. Verfahren nach Anspruch 3, **gekennzeichnet durch** weiteres Multiplizieren der sich **durch** den Additionsvorgang ergebenenden Signale mit einem Koeffizienten, der äquivalent ist einem reziproken Wert der Empfangsleistung von empfangenen Pilotkorrelationssignalen und **durch** Ausgeben des Multiplikationsergebnisses für die Durchschnittswertbildung.

5. Verfahren nach Anspruch 3 oder 4, **gekennzeichnet durch** Erzeugen von Pseudorauschcodes für die gleichphasige Achse und die Querachse in Abhängigkeit von den Zeitsignalen.

6. Verfahren nach Anspruch 3 oder 4, **dadurch gekennzeichnet, daß** der Demodulationsvorgang der empfangenen Signale von dem Zeitsignal abhängt.

7. Verfahren nach einem der Ansprüche 1 bis 6, **dadurch gekennzeichnet, daß** die empfangenen Signale für die gleichphasige Achse und die Querachse mit Pseudorauschcodes für die gleichphasige Achse und die Querachse multipliziert werden.

8. Verfahren nach einem der Ansprüche 1 bis 6, **dadurch gekennzeichnet, daß** das empfangene Signal für die gleichphasige Achse mit Pseudorauschcodes für die gleichphasige Achse und die Querachse multipliziert wird.

9. Verfahren nach einem der Ansprüche 1 bis 6, **dadurch gekennzeichnet, daß** das empfangene Signal für die Querachse mit Pseudorauschcodes für die gleichphasige Achse und die Querachse multipliziert wird.

10. Datendemodulator einer Empfangsvorrichtung, der das Verfahren nach Anspruch 1 durchführt, enthaltend Mittel, die zum Empfang eines empfangenen Signals für die gleichphasige Achse und die Querachse geeignet sind, Korrelationsberechnungsmittel aufweisend erste Multiplikationsmittel, die zum Multiplizieren der empfangenen Signale für die gleichphasige Achse und/oder die Querachse mit einem Pseudorauschcode entsprechend einem von einer Basisstation übertragenen Pilotsignal geeignet sind, und Durchschnittswertbildungsmittel (20, 22, 24, 26), die zur Durchschnittswertbildung der Ergebnisse der Multiplikation zum Berechnen einer Korrelation, die eine Information über die nach der Erfassung verbleibende Phasendifferenz und eine Phasendifferenz enthalten in empfangenen Signalen für die gleichphasige Achse und die Querachse geeignet sind, welcher Demodulator weiterhin aufweist Phasendifferenzkompensationsmittel (21, 23, 28, 30, 32, 34) zum Kompensieren einer Wirkung einer Phasendifferenz, die in dem empfangenen Signal für die gleichphasige Achse und die Querachse enthalten ist, unter Verwendung der Phasendifferenz, und Datendemodulationsmittel (40, 42, 44, 46, 48, 50, 52, 54, 56, 58, 60, 520) aufweisen zweite Multiplikationsmittel (40, 42), die zum Multiplizieren der empfangenen Signale für die gleichphasige Achse und die Querachse geeignet sind, für welche die Wirkung der Phasendifferenz kompensiert wird durch die Phasendifferenzkompensationsmittel durch Pseudorauschcodes für die gleichphasige Achse bzw. die Querachse, entsprechend dem von der Basisstation übertragenen Pilotsignal, und Additionsmittel (44), die zum Demodulieren eines durch Addieren der Multiplikationsergebnisse miteinander erhaltenen Signals geeignet sind.

11. Datendemodulator nach Anspruch 10 zum Durchführen des Verfahrens nach Anspruch 3, weiterhin enthaltend dritte Multiplikationsmittel, die zum Multiplizieren der empfangenen Signale für die gleichphasige Achse mit kompensierter Phasendifferenz mit den zeitverschobenen Pseudorauschcodes für die gleichphasige Achse geeignet sind,

   vierte Multiplikationsmittel, die zum Multiplizieren des empfangenen Signals für die Querachse mit kompensierter Phasendifferenz mit den zeitverschobenen Pseudorauschcodes für die Querachse geeignet sind;
   weiterhin Additionsmittel, die zum Addieren der Ergebnisse der dritten und der vierten Multiplikationsmittel geeignet sind;
   weiterhin Durchschnittswertbildungsmittel für die Durchschnittswertbildung der addierten Signale;
   Zeitsteuermittel zum Erzeugen des Zeitsignals.

12. Datendemodulator nach Anspruch 11 zur Durchführung des Verfahrens nach Anspruch 4, enthaltend fünfte Multiplikationsmittel, die zwischen den weiteren Additionsmitteln und den weiteren Durchschnittswertbildungsmitteln angeordnet sind.

Revendications

1. Procédé pour démoduler des données pour une communication à spectre étalé, selon lequel des signaux modulés selon un spectre étalé pour un axe en phase et un axe en quadrature dans un système d'étalement direct avec des codes de pseudo-bruit d'axe en phase et d'axe en quadrature, sont reçus et des données provenant des signaux reçus sont récupérées, ce qui a pour effet que les signaux reçus sont corrélés aux signaux de pseudo-bruit, **caractérisé en ce qu'**un signal d'axe en phase et/ou un signal d'axe en quadrature reçus sont multipliés par un code de pseudo-bruit correspondant à un signal pilote émis par un poste de base, et la moyenne des résultats de la multiplication est calculée pour le calcul d'une corrélation contenant une information de différence de phase qui subsiste après la détection et une différence de phase contenue dans les signaux d'axe en phase et d'axe en quadrature reçus est compensée moyennant l'utilisation de l'information de différence de phase obtenue au moyen du processus de corrélation, et que les signaux d'axe en phase et d'axe en quadrature reçus, pour lesquels la différence de phase est compensée, sont multipliés respectivement par des codes de pseudo-bruit d'axe en phase et d'axe en quadrature, qui correspondent aux signaux pilotes émis par le poste de base, et un signal obtenu par addition des résultats de multiplication entre eux est démodulé.

2. Procédé selon la revendication 1, **caractérisé en ce que** le signal est démodulé moyennant l'utilisation d'un traitement du signal au moyen d'une transformée de Hadamard rapide (FHT).

3. Procédé selon la revendication 1, **caractérisé en outre par** les étapes consistant à multiplier le signal d'axe en phase reçu, pour lequel la différence de phase est compensée par des codes de pseudo-bruit d'axe en phase, obtenus par un décalage d'avance et de retard d'un code de pseudo-bruit d'axe en phase;

   multiplier le signal d'axe en phase reçu, pour lequel la différence de phase est compensée, par des codes de pseudo-bruit d'axe en phase obtenus par Un léger décalage d'avance et de retard de l'instant d'un code de pseudo-bruit d'axe en phase;
   multiplier le signal reçu d'axe en quadrature reçu, pour lequel la différence de phase est compensée, par les codes de pseudo-bruit d'axe en quadrature obtenus par un léger décalage d'avance et de retard de l'instant d'un code de pseudo-bruit d'axe en quadrature;
   additionner les signaux d'axe en phase et d'axe en quadrature reçus, multipliés par les codes de pseudo-bruit;
   calculer la moyenne des signaux additionnés; ce qui a pour effet qu'en réponse au résultat de l'opération de la procédure de calcul de la moyenne, un signal de cadencement est délivré.

4. Procédé selon la revendication 3, **caractérisé en outre par** la multiplication des signaux résultant de la procédure d'addition par un coefficient équivalent à l'inverse de la puissance de réception des signaux de corrélation pilotes reçus et la délivrance du résultat de multiplication pour la formation de la moyenne.

5. Procédé selon la revendication 3 ou 4, **caractérisé par** la production de codes de pseudo-bruit d'axe en phase et d'axe en quadrature, en réponse aux signaux de cadencement.

6. Procédé selon la revendication 3 ou 4, **caractérisé en ce que** la procédure de modulation des signaux reçus a

été appliquée en réponse au signal de cadencement.

7. Procédé selon l'une des revendications 1 à 6, **caractérisé en ce que** les signaux d'axe en phase et d'axe en quadrature reçus sont multipliés par des codes de pseudo-bruit d'axe en phase et d'axe en quadrature.

8. Procédé selon l'une des revendications 1 à 6, **caractérisé en ce que** le signal d'axe en phase reçu est multiplié par des codes de pseudo-bruit d'axe en phase et d'axe en quadrature.

9. Procédé selon l'une des revendications 1 a 6, **caractérisé en ce que** le signal d'axe en quadrature reçu est multiplié par des codes de pseudo-bruit d'axe en phase et d'axe en quadrature.

10. Démodulateur de données d'un appareil de réception, mettant en oeuvre le procédé selon la revendication 1, comprenant des moyens adaptés pour recevoir un signal d'axe en phase reçu et un signal d'axe en quadrature reçu, des moyens de calcul de corrélation comprenant des premiers moyens de multiplication adaptés pour multiplier les signaux d'axe en phase et/ou d'axe en quadrature reçus par un code de pseudo-bruit correspondant à un signal pilote émis par un poste de base, et des moyens de calcul de la moyenne (20,22,24,26) adaptés pour former la moyenne des résultats de la multiplication pour le calcul d'une' corrélation contenant une information de différence de phase qui subsiste après la détection, et d'une différence de phase contenue dans les signaux d'axe en phase et d'axe en quadrature reçus; le démodulateur comprenant en outre des moyens (21,23,28,30,32,34) de compensation de la différence de phase, servant à compenser un effet d'une différence de phase obtenue dans un signal d'axe en phase et un signal d'axe en quadrature reçus moyennant l'utilisation de la différence de phase, et des moyens de démodulation de données (40,42,44,46,48,50,52,54, 56,58,60,520) comprenant des seconds moyens de multiplication (40,41) adaptés pour multiplier les signaux d'axe en phase et d'axe en quadrature reçus, pour lesquels l'effet de la différence de phase est compensé par lesdits moyens de compensation de différence de phase, respectivement par des codes de pseudo-bruit d'axe en phase et d'axe en quadrature correspondant au signal pilote émis par le poste de base et des moyens d'addition (44) adaptés pour démoduler un signal produit par addition des résultats de multiplication entre eux.

11. Démodulateur de données selon la revendication 10, pour la mise en oeuvre du procédé selon la revendication 3, comprenant en outre des troisièmes moyens de multiplication adaptés pour multiplier les signaux d'axe en phase reçus, dont la différence de phase est compensée, par les codes de pseudo-bruit d'axe en phase décalés dans le temps;

des quatrième moyens de multiplication adaptés pour multiplier le signal d'axe en quadrature reçu, dont la différence de phase est compensée, par des codes de pseudo-bruit d'axe en quadrature, décalés dans le temps;
d'autres moyens d'addition adaptés pour additionner les résultats des troisième et quatrième moyens de multiplication;
d'autres moyens de calcul de la moyenne pour calculer la moyenne des signaux additionnés;
des moyens de commande de cadencement pour produire le signal de cadencement.

12. Démodulateur de données selon la revendication 11, pour la mise en oeuvre du procédé selon la revendication 4, comprenant des cinquièmes moyens de multiplication disposés entre lesdits autres moyens d'addition et lesdits autres moyens de calcul de la moyenne.

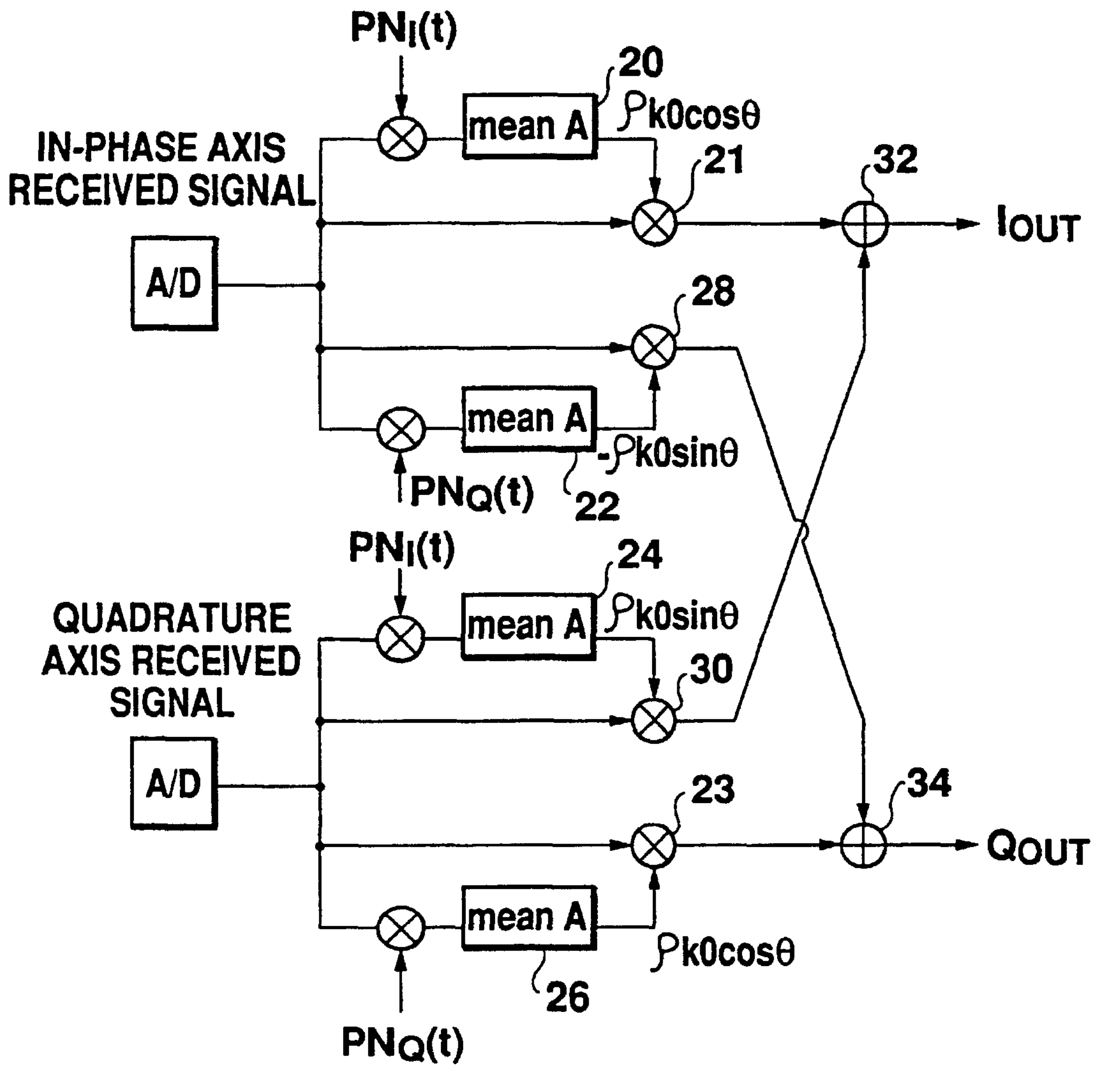
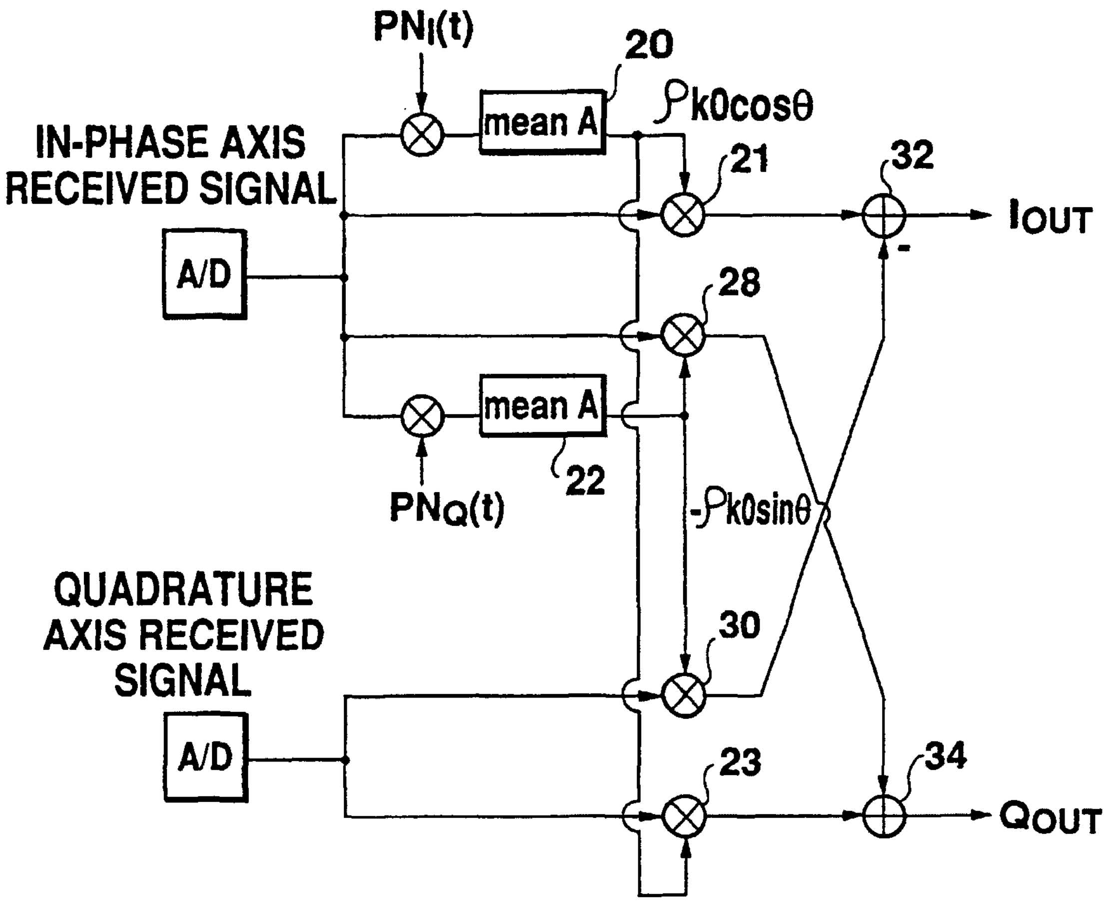

PNI(t)
20
ρk0cosθ
mean A
IN-PHASE AXIS RECEIVED SIGNAL
21
32
IOUT
A/D
28
mean A
-ρk0sinθ
PNQ(t)
22
PNI(t)
24
mean A
ρk0sinθ
QUADRATURE AXIS RECEIVED SIGNAL
30
A/D
23
34
QOUT
mean A
ρk0cosθ
26
PNQ(t)

Fig. 1



$PN_I(t)$
20
mean A
$\rho k0\cos\theta$
IN-PHASE AXIS
RECEIVED SIGNAL
21
32
I_{OUT}
A/D
-
28
mean A
22
$PN_Q(t)$
$-\rho k0\sin\theta$
QUADRATURE
AXIS RECEIVED
SIGNAL
30
A/D
23
34
Q_{OUT}

Fig. 2

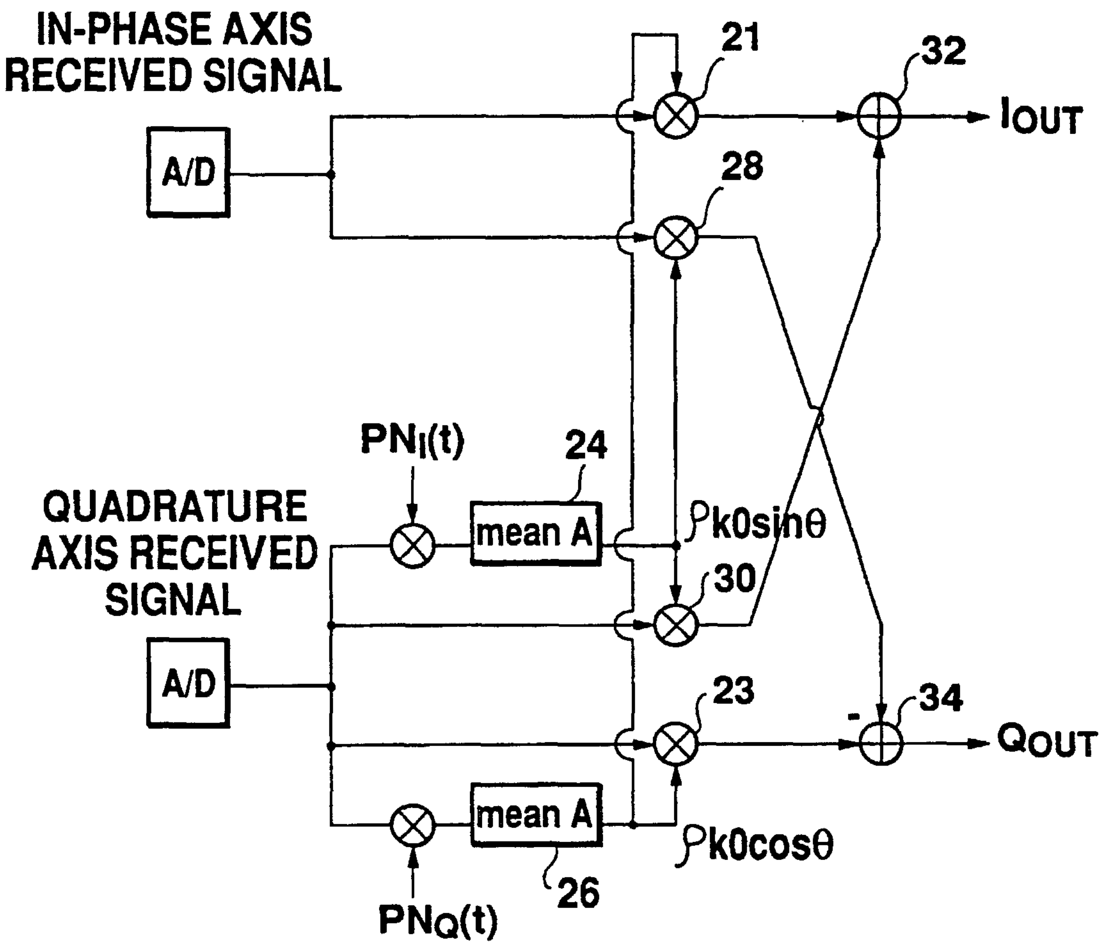
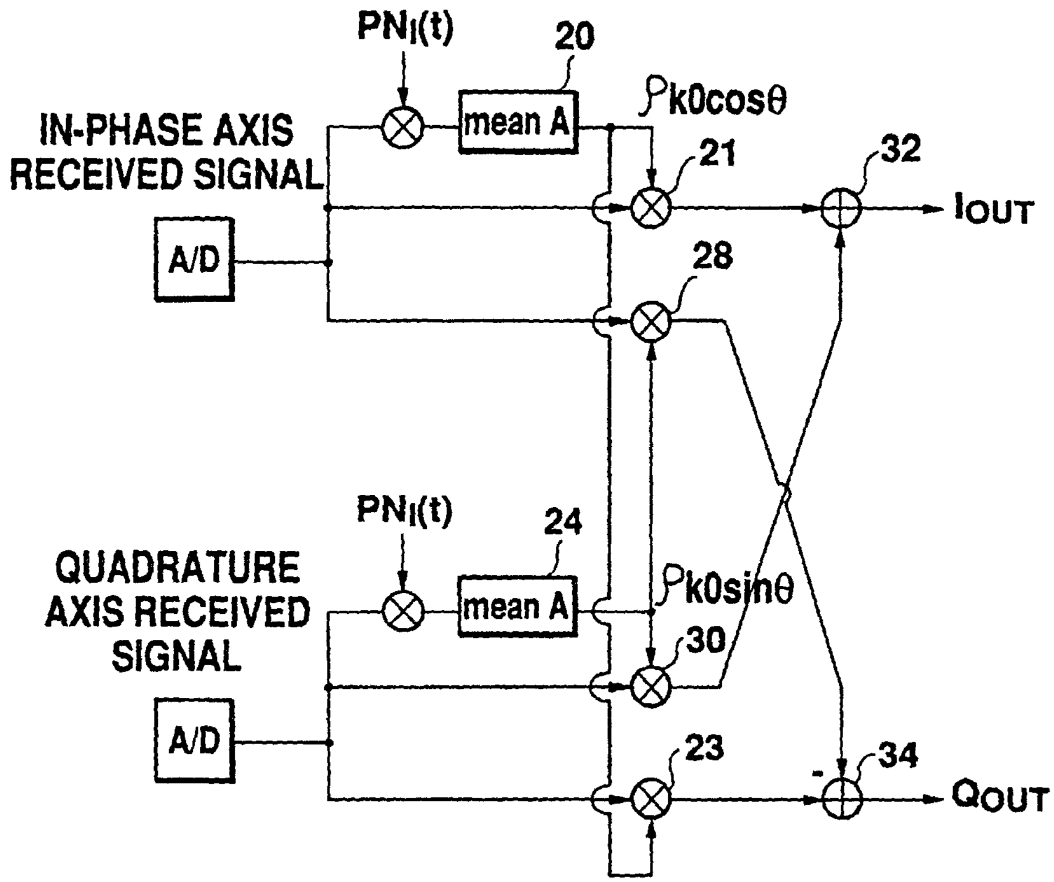

IN-PHASE AXIS RECEIVED SIGNAL
A/D
21
32
I_{OUT}
28
$PN_I(t)$
24
QUADRATURE AXIS RECEIVED SIGNAL
mean A
$\rho k0\sin\theta$
30
A/D
23
34
-
Q_{OUT}
mean A
$\rho k0\cos\theta$
26
$PN_Q(t)$

Fig. 3



$PN_I(t)$
20
mean A
$\rho k0\cos\theta$
IN-PHASE AXIS
RECEIVED SIGNAL
21
32
I_{OUT}
A/D
28
$PN_I(t)$
24
QUADRATURE
AXIS RECEIVED
SIGNAL
mean A
$\rho k0\sin\theta$
30
A/D
23
34
-
Q_{OUT}

Fig. 4

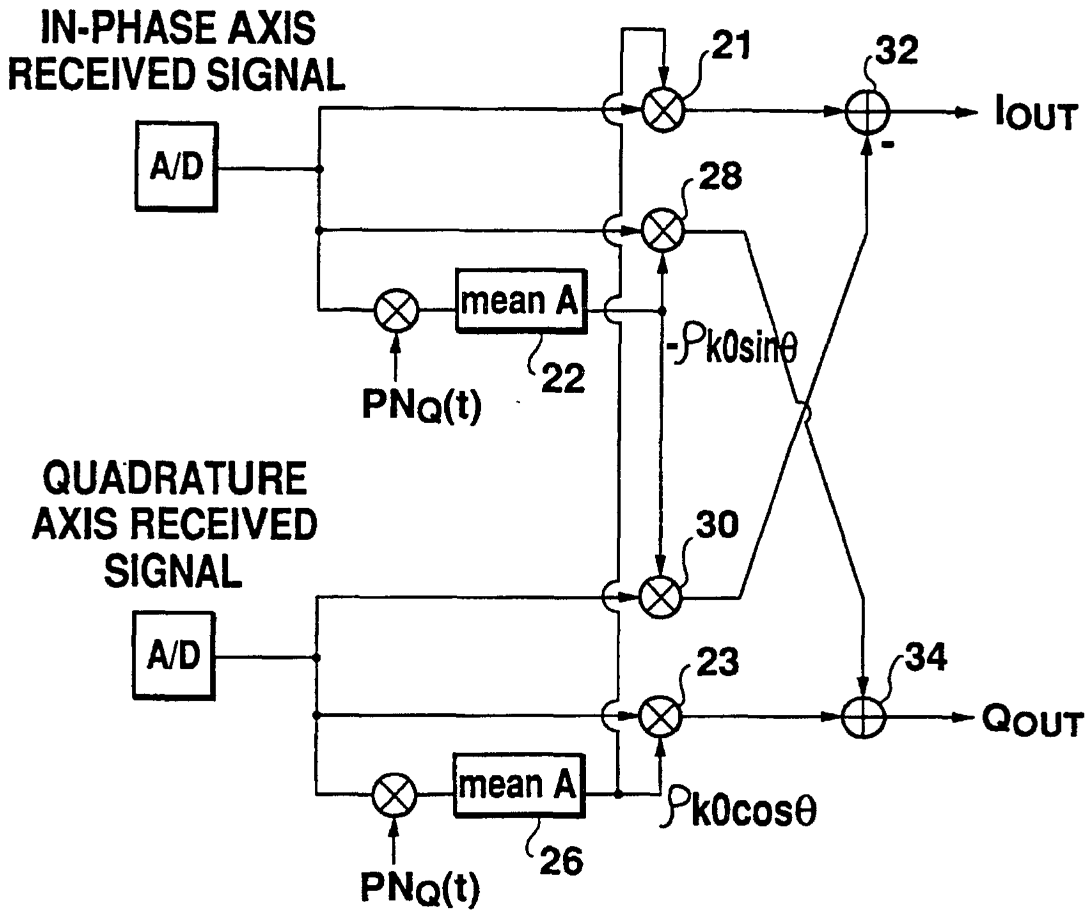

IN-PHASE AXIS RECEIVED SIGNAL
A/D
21
32
IOUT
28
mean A
-ρk0sinθ
22
PNQ(t)
QUADRATURE AXIS RECEIVED SIGNAL
30
A/D
23
34
QOUT
mean A
ρk0cosθ
26
PNQ(t)

Fig. 5

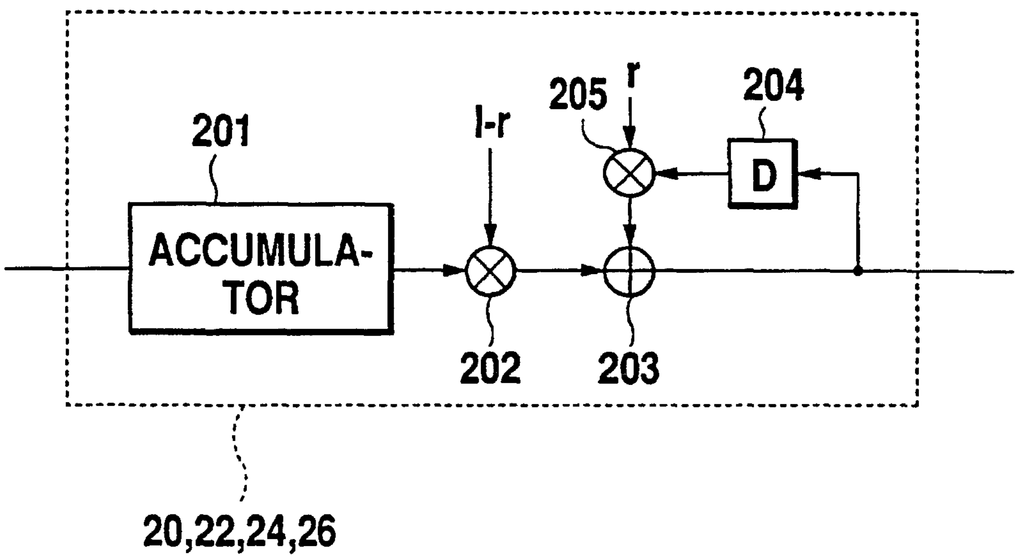
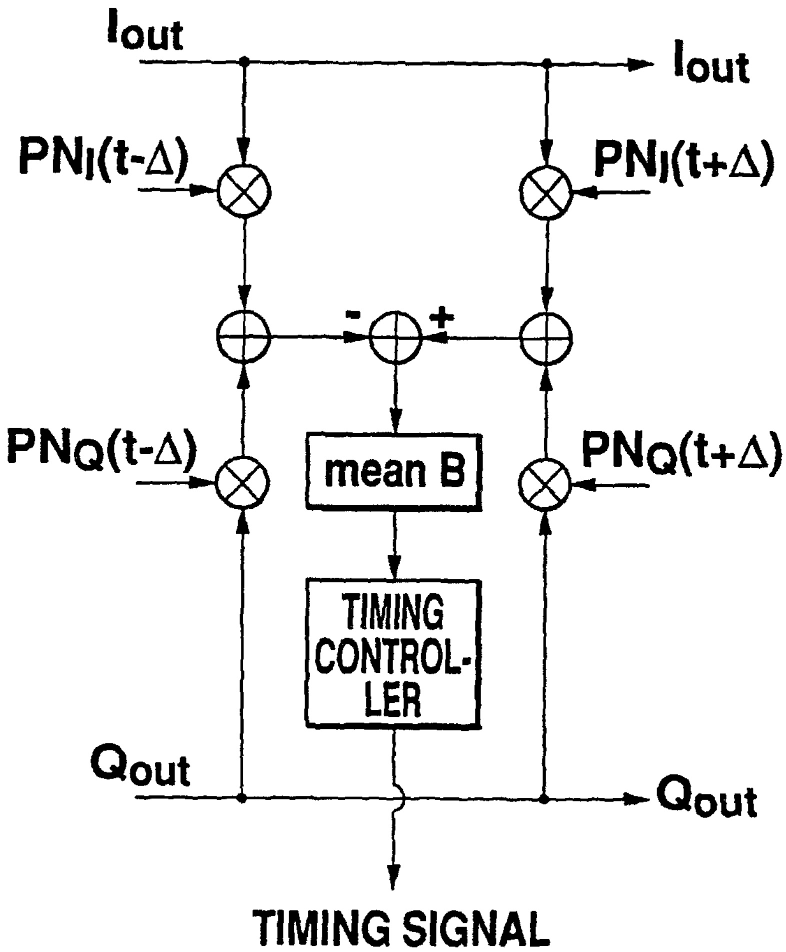


Fig. 6



I_{out}
I_{out}
$PN_I(t-\Delta)$
$PN_I(t+\Delta)$
-
+
$PN_Q(t-\Delta)$
mean B
$PN_Q(t+\Delta)$
TIMING CONTROL-LER
Q_{out}
Q_{out}
TIMING SIGNAL

Fig. 7

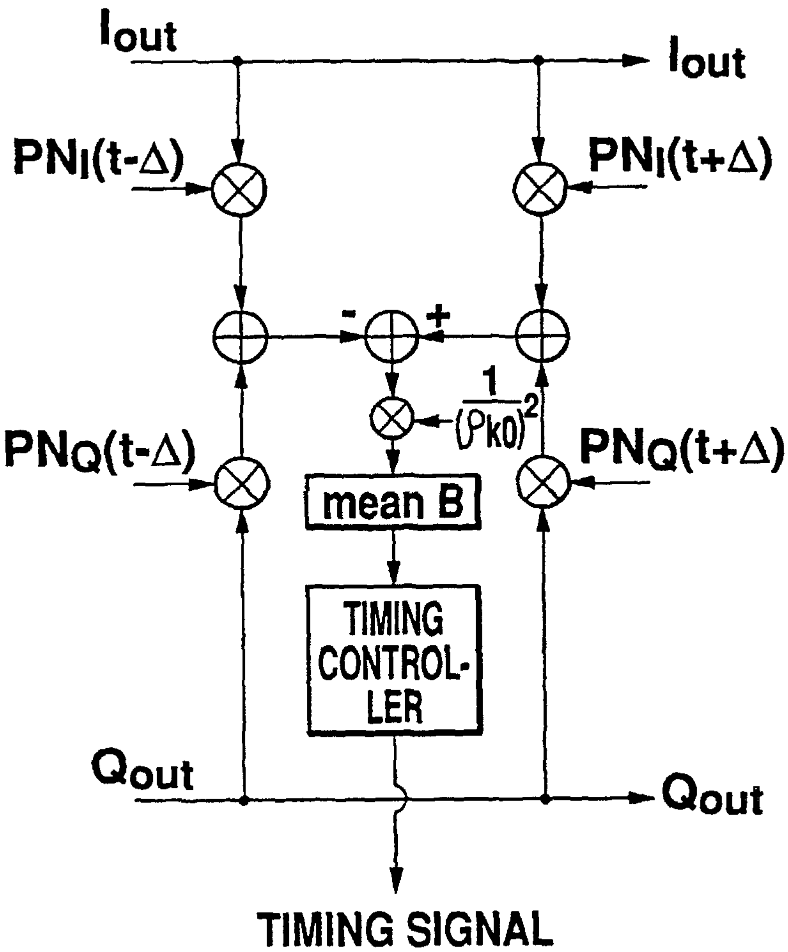

I_{out}
I_{out}
$PN_I(t-\Delta)$
$PN_I(t+\Delta)$
-
+
$\frac{1}{(\rho_{k0})^2}$
$PN_Q(t-\Delta)$
$PN_Q(t+\Delta)$
mean B
TIMING CONTROL-LER
Q_{out}
Q_{out}
TIMING SIGNAL

Fig. 8

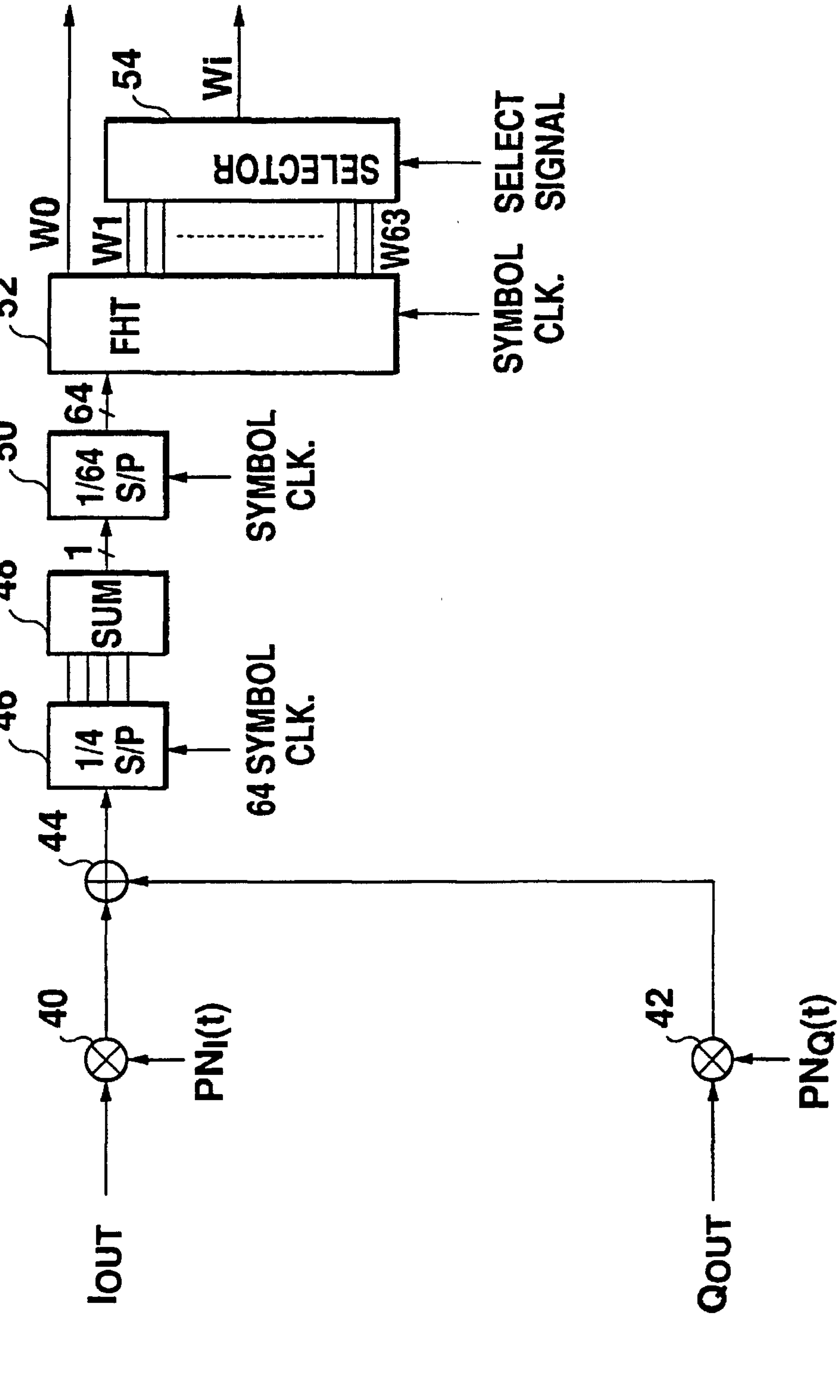

IOUT
40
PNI(t)
QOUT
42
PNQ(t)
44
46
1/4 S/P
64 SYMBOL CLK.
48
SUM
1
50
1/64 S/P
SYMBOL CLK.
64
52
FHT
SYMBOL CLK.
W0
W1
W63
SELECTOR
SELECT SIGNAL
54
Wi

Fig. 9

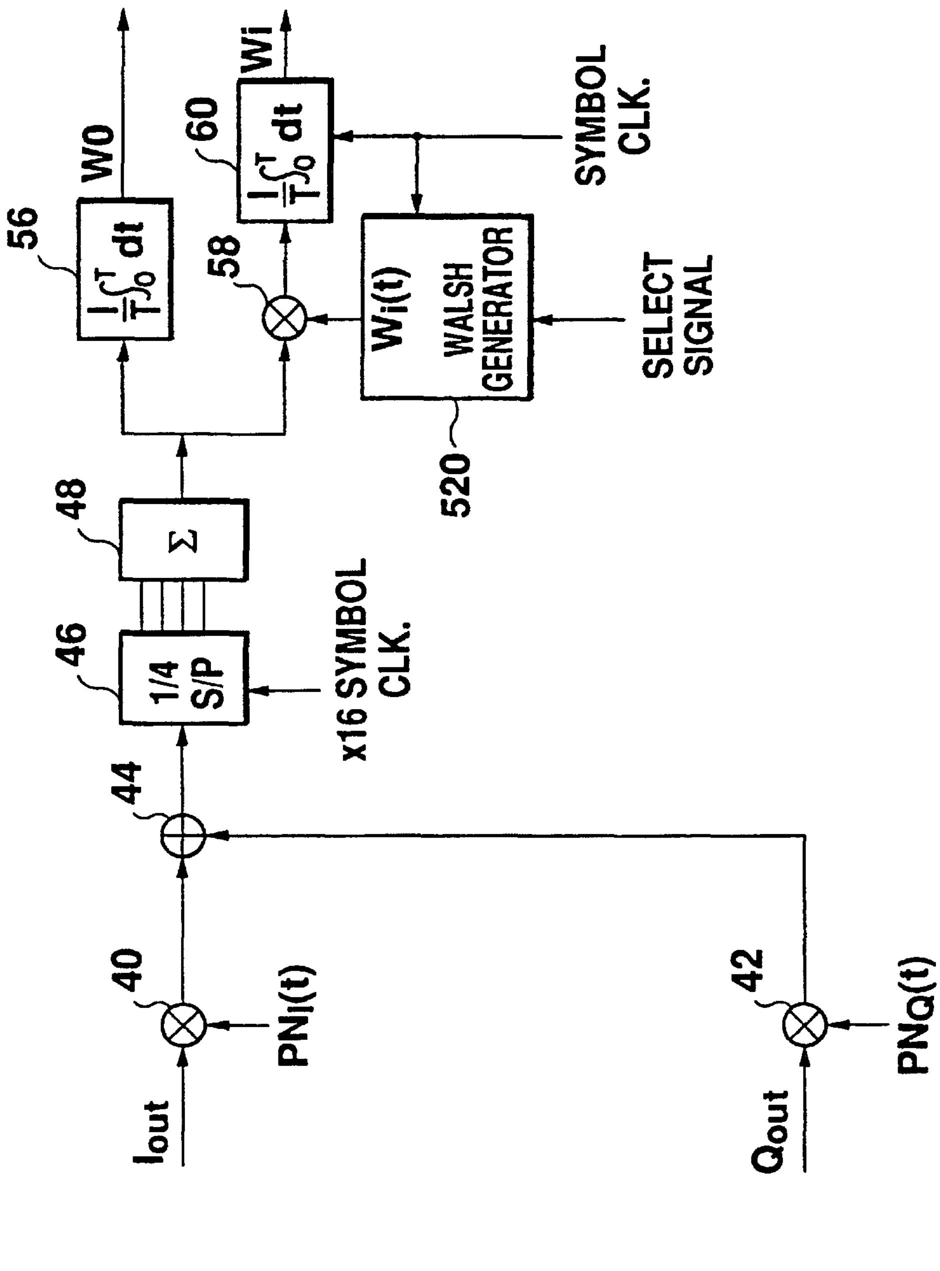

Iout
40
PNI(t)
Qout
42
PNQ(t)
44
46
1/4
S/P
x16 SYMBOL
CLK.
48
Σ
56
W0
58
60
Wi
Wi(t)
WALSH
GENERATOR
520
SYMBOL
CLK.
SELECT
SIGNAL

Fig. 10

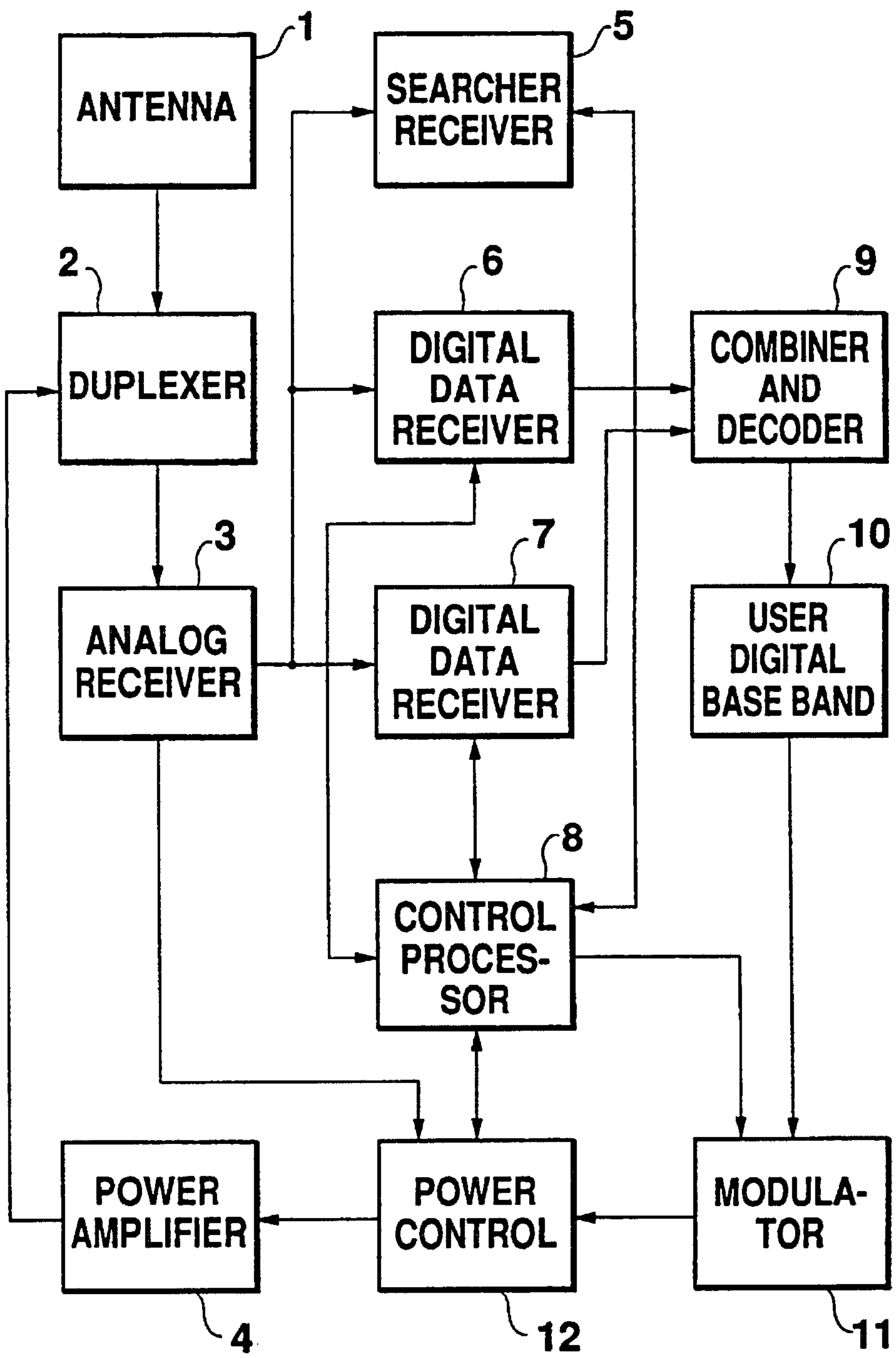

ANTENNA
1
SEARCHER RECEIVER
5
2
DUPLEXER
DIGITAL DATA RECEIVER
6
COMBINER AND DECODER
9
ANALOG RECEIVER
3
DIGITAL DATA RECEIVER
7
USER DIGITAL BASE BAND
10
CONTROL PROCES-SOR
8
POWER AMPLIFIER
4
POWER CONTROL
12
MODULA-TOR
11

Fig. 11

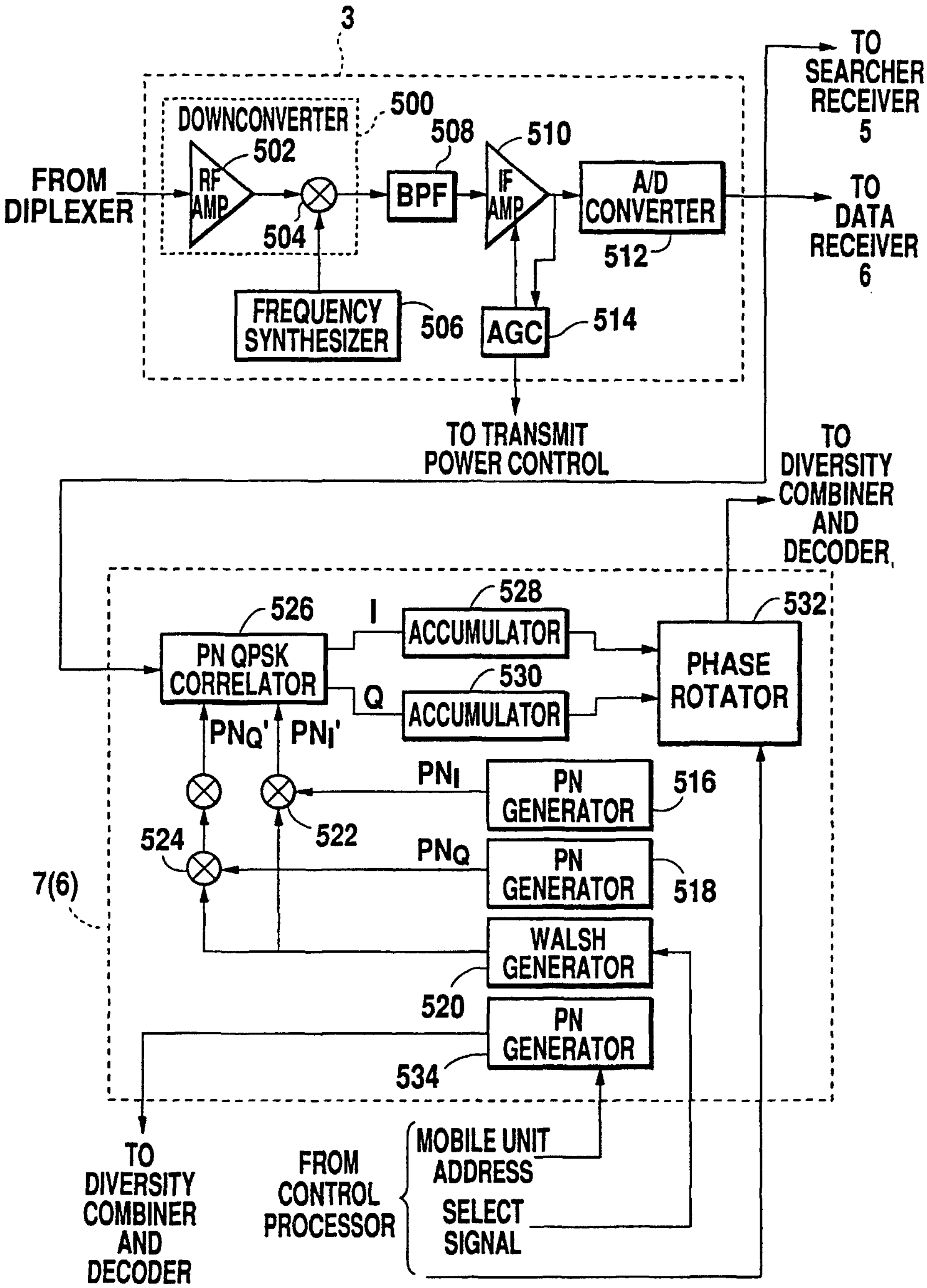

3
DOWNCONVERTER
500
502
RF AMP
504
FROM DIPLEXER
508
BPF
510
IF AMP
A/D CONVERTER
512
TO SEARCHER RECEIVER 5
TO DATA RECEIVER 6
FREQUENCY SYNTHESIZER
506
AGC
514
TO TRANSMIT POWER CONTROL
TO DIVERSITY COMBINER AND DECODER
526
PN QPSK CORRELATOR
I
Q
528
ACCUMULATOR
530
ACCUMULATOR
532
PHASE ROTATOR
PNQ'
PNI'
PNI
PN GENERATOR
516
522
524
PNQ
PN GENERATOR
518
7(6)
WALSH GENERATOR
520
PN GENERATOR
534
TO DIVERSITY COMBINER AND DECODER
FROM CONTROL PROCESSOR
MOBILE UNIT ADDRESS
SELECT SIGNAL

Fig. 12

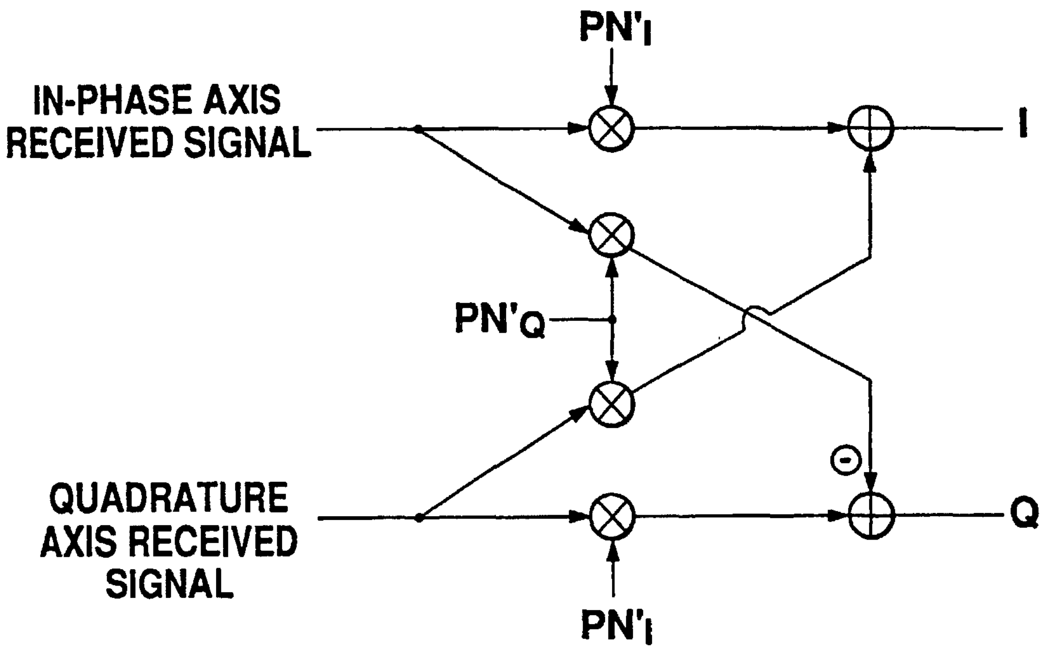

PN'I
IN-PHASE AXIS RECEIVED SIGNAL
I
PN'Q
QUADRATURE AXIS RECEIVED SIGNAL
Q
PN'I

Fig. 13