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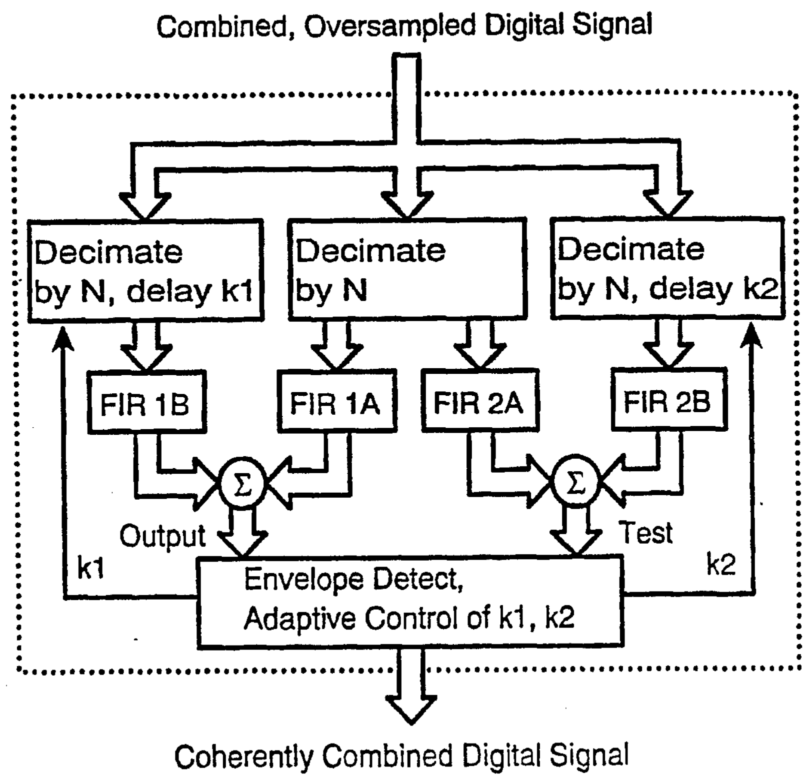
(54) Title: A SIGNAL COMBINER SYSTEM

(57) Abstract

A method of signal combining, comprises the addition of a number of received bandpass signals occupying different spectral positions to give a signal sum, the spectral replication of the resulting signals using bandpass sampling, and the interpolation over at least one frequency bandwidth of the combined and spectrally relocated signals, producing a combined digital signal. The individual received signals may be first spectrally relocated to individual integer positioned intermediate frequencies where the individual received signals are not appropriately positioned for bandpass sampling and interpolation, prior to adding the signals and bandpass sampling. The bandpass sampling is preferably second-order or higher order. The method may be used in frequency or time diversity systems or for demodulating a frequency-hopping spread spectrum signal without a priori knowledge of the frequency-hop pattern.



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A SIGNAL COMBINER SYSTEM

Field of the Invention

The invention comprises a method and system for frequency-shifting and combining received radio communications signals.

Background

A basic operation in signal processing is the combination of two or several bandlimited signals. For L signals, $s_1(t)$ to $s_L(t)$, a linear combination may be expressed as $c(t)$ such that

$$c(t) = \sum_{i=1}^{L} w_i(t)\, s_i(t)\,, \tag{1}$$

where $w_i(t)$ are complex weights. Of particular importance is the combining of multiple copies of the same signal, such as in a receiving array of antennas. This type of combining can be used to improve the signal-to-noise ratio, reduce envelope fading, and decrease interference in communications systems.

The present invention provides and improved or at least alternative method and system for signal combining.

Summary of Invention

In broad terms the invention comprises a method of signal combining, comprising:

the addition of a number of received bandpass signals occupying different spectral positions to give a signal sum, the spectral replication of the resulting signals using bandpass sampling, and

the interpolation over at least one frequency bandwidth of the combined and spectrally relocated signals, producing a combined digital signal.

Where the individual received signals are not appropriately positioned for bandpass sampling and interpolation, the method of invention includes first spectrally relocating the individual received signals to individual integer positioned intermediate frequencies prior to adding the signals and bandpass sampling. This may not be necessary where the signals are already suitably located across the frequency spectrum, such as may be the case for frequency diversity systems and frequency-hopping spread spectrum systems.

Preferably the bandpass sampling is second-order or higher order.

Preferably the bandpass sampling rate is an integer multiple of the nominal bandwidth of the individual bandpass signals to be combined and most preferably is four times this bandwidth.

Preferably the interpolation of the spectrally relocated bandpass signals resulting from the bandpass sampling consists of bandlimiting and phase-shifting the bandpass signals. The interpolation is carried out using either analogue bandpass filters, digital finite impulse response (FIR) filters, or digital infinite impulse response (IIR) filters.

The invention also comprises a system for combining signals received by a diversity system using space, frequency or time diversity, comprising:

means to demodulate the received signals to individual integer positioned intermediate frequencies,

means to add the intermediate frequency signals to give a signal sum,

means to bandpass sample preferably by second-order bandpass sampling the signal sum,

means to digitally or otherwise interpolate the replicated signal sum over at least one frequency bandwidth containing information from each received signal resulting from the bandpass sampling to provide a means to phase-control the component signals of the signal sum, and

means to detect the performance of the system, deduce the component signal phases and thereby adaptively control the phases of the component signals of the combined signal sum using the second-order sampling parameter k to control the phases.

The invention also comprises a system for demodulating a frequency-hopping spread spectrum signal without *a priori* knowledge of the frequency-hop pattern, comprising

means to bandpass sample the received bandpass signal preferably by quadrature or other second-order sampling, means to digitally or otherwise interpolate the replicated signal over at least one frequency bandwidth containing the information from the received signal to produce the in-phase and quadrature components of the received signal, and

means to combine the in-phase and quadrature components of the received signal so as to detect the envelope of the signal.

The invention also comprises a system for producing uncorrelated signals using an array of receiving antennas and bandpass sampling as a switchable pattern antenna, comprising

a means to demodulate the signals from each element of the receiving antenna array to individual integer positioned intermediate frequency bandpositions,

a means to add the intermediate frequency signals to produce a signal sum,

a means to bandpass sample the signal sum preferably using second-order bandpass sampling to produce a spectrally replicated signal sum,

a means to digitally or otherwise interpolate the relocated signal over at least one frequency bandwidth to produce a combined signal with a unique radiation pattern which is selectable using the second-order sampling parameter k.

Brief Description of Figures

The invention will be further described with reference to the accompanying figures wherein:

Fig 1 shows the minimum rate sampling of bandpass signals,

Fig 2 shows the separated spectra of a second-order sampled bandpass signal,

Fig 3a shows a typical conventional coherent signal combiner,

Fig 3b shows a bandpass sampling signal combiner of the invention for three branch space diversity,

Fig 4 shows digital signal processing detail of the final stage of the second-order bandpass sampling signal combiner of Fig 3b,

Fig 5a is a contour plot of the performance surface of the combiner of Figs 3b and 4,

Fig 5b shows the performance surface of Fig 5a in three dimensions,

Fig 6 is a typical Rayleigh diagram for the combiner of Figs 3b and 4,

Fig 7 show the signal envelopes of some of the systems which data are shown in the Rayleigh diagram in Fig 6, Fig 8a shows a typical conventional frequency-hopping spread spectrum demodulator,

Fig 8b shows a bandpass sampling frequency-hopping spread spectrum demodulator of the invention,

Fig 9a shows typical system output for a quadrature bandpass sampling frequency-hopping spread spectrum demodulator of the invention and shown in figure 8b,

Fig 9b shows the relative signal phase of the data shown in Fig 9b

Fig 10 shows a switchable pattern antenna using the bandpass sampling signal combiner of the invention,

Fig 11 shows typical radiation patterns for the system of Fig 11.

Bandpass Sampling

Bandpass sampling is the process of sampling a bandpass signal at a rate dependent upon the signal's bandwidth rather than its bandposition as described by A.Kohlenburg "Exact interpolation of bandlimited functions" *J. Appl. Phys.* vol.24 no.12 Dec. 1953 pp. 1432-1436, further described later. Interpolants may then be used to reconstruct the signal at any of a number of bandpositions, these being determined by the original bandposition and the sampling rate.

Spectral Replication through Sampling

A complex low-pass signal of two-sided bandwidth B, $m(t)$, is modulated to a bandpass position such that

$$r_1(t) = |m(t)| \cos(2\pi f_{C1} t + arg[m(t)]), \tag{2}$$

for a carrier f_{C1}. The spectrum of $r_1(t)$, shown in figure 1a, is integer positioned if $f_{C1}=(n_1+½)B$ for integer n_1, see R.G.Vaughan, N.L.Scott and D.R.White "The Theory of bandpass sampling" *IEEE Trans. Sig. Proc.* vol.39, no.9 Sept.1991. $r_1(t)$ may be minimum-rate sampled, where $f_S=2B$, such that the sampled signal is

$$r_1^{\delta 1}(t) = r_1(t) \sum_{p=-\infty}^{\infty} \delta\left(t - \frac{p}{2B}\right), \tag{3}$$

where p is an integer index. The notation $r_1^{\delta 1}(t)$ is used to denote a first-order sampled signal. The spectrum of $r_1^{\delta 1}(t)$, $R_1^{\delta 1}(f)$, is given by the convolution

$$R_1^{\delta 1}(f) = R_1(f) \bigstar \sum_{p=-\infty}^{\infty} \delta(f - p2B), \tag{4}$$

and shown in figure 1c for n_1 even. The notation $R_1^{\delta 1}(f)$ is used to denote the spectrum of a first-order sampled signal. Here, p corresponds the bandposition of the pth replica of $m(t)$.

For the purpose of demonstrating signal combination, a second real bandpass signal, $r_2(t)$, is produced by modulating $m(t)$ with a second carrier, $f_{C2}=(n_2+½)B$ for integer $n_2 \neq n_1$, such that

$$r_2(t) = |m(t)| \cos(2\pi f_{C2}t + arg[m(t)]) . \tag{5}$$

The spectrum of $r_2(t)$ is shown in figure 1b. Because $r_1(t)$ and $r_2(t)$ have identical complex envelopes, Minimum-rate sampling $r_2(t)$ using the sampling function in equation (3) produces a signal, $r_2^{\delta 1}(t)$, whose spectrum, $R_2^{\delta 1}(f)$, is identical to that of $R_1^{\delta 1}(f)$, for n_2 even. Further, the sampling of an arithmetic sum of the two bandpass signals, $r_1(t)+r_2(t)$, will produce the same spectrum, with double the energy density. Provided the complex envelopes of $r_1(t)$ and $r_2(t)$ have the same phase at the sample times, each sample of $r_1^{\delta 1}(t)+r_2^{\delta 1}(t)$ will be double the amplitude of the corresponding sample of $r_1^{\delta 1}(t)$.

The signal sum, $r_1(t)+r_2(t)$, contains two distinct copies of the original signal, $m(t)$, but both copies are individually retrievable by bandpass filtering. Once sampled, however, the signals are no longer individually retrievable, as they have been combined by the sampling process. If the complex envelopes of $r_1(t)$ and $r_2(t)$ are not in-phase, the sampled sum of signals will be of less than double the amplitude of $r_1^{\delta 1}(t)$. In the case of $r_1(t)$ and $r_2(t)$ being in exact anti-phase at the sample times, the sampled sum of signals will have zero amplitude. The relative phase of $r_1(t)$ and $r_2(t)$ is, therefore, critical to the sampling process. The phases of the sampled signals may be controlled by using second-order sampling.

Second-Order Sampling and the Bandposition-Dependent Phase-Shift

A second-order sampled signal [2] is defined as

$$\begin{aligned} r_1^{\delta 2}(t) &= r_1(t) \sum_{p=-\infty}^{\infty} \left[\delta\left(t - \frac{2p}{f_S}\right) + \delta\left(t - \frac{2p}{f_S} - k\right) \right] \\ &= r_1^{\delta 2A}(t) + r_1^{\delta 2B}(t) , \end{aligned} \tag{6}$$

where k is the time-separation between the two interleaved sampling functions (k is often referred to as the *sample stream delay*), and f_S is the average sampling rate of $r_1^{\delta 2}(t)$. Here the notation $r_1^{\delta 2}(t)$ is used to denote a second-order sampled signal. This signal is the sum of two first-order sampled signals, $r_1^{\delta 2A}(t)$ and $r_1^{\delta 2B}(t)$, as defined in equation (6). The second-order sampled signal, $r_1^{\delta 2}(t)$, has the spectrum

$$\begin{aligned} R_1^{\delta 2}(f) &= R_1(f) \bigstar B \sum_{p=-\infty}^{\infty} \left[\delta\left(f - \frac{pf_S}{2}\right) + \gamma^p \delta\left(f - \frac{pf_S}{2}\right)\right] \\ &= BR_1^{\delta 2A}(f) + BR_1^{\delta 2B}(f), \end{aligned} \tag{7}$$

where $\gamma = exp(-j\pi k f_S)$. The notation $R_1^{\delta 2}(f)$ is used to denote the spectrum of a second-order sampled signal. The separated spectra of $R_1^{\delta 2}(f)$ are shown in figure 2 for minimum-rate sampling, where $f_S = 2B$ and $\gamma = exp(-j2\pi kB)$. These spectra are the separated spectra of $R_1^{\delta 2A}(f)$ and $R_1^{\delta 2B}(f)$, as defined by equation (7). The superscript "+" is used in figure 2 to indicate replicas of the positive frequency components of $r_1(t)$. Similarly, the superscript "-" is used to indicated replicas of the negative frequency components of $r_1(t)$. Note the phase-shift of the replicas of $R_1^{\delta 2B}(f)$ relative to the replicas of $R_1^{\delta 2A}(f)$, as shown in figure 2. For each replica of $m(t)$, this phase-shift is a function of k and the replica's bandposition, p, relative to the pre-sampled bandpass signal's bandposition, n_1, as defined in figure 1.

The original signal, $m(t)$, can be reconstructed from the second-order samples at any integer bandposition by interpolation, see D.A.Linden "A discussion of sampling theorems" *Proc. IRE.* vol.47 1959 pp.1219-1226. Thus, second-order sampling and interpolation can be used both to frequency-shift, and to phase-shift a bandpass signal. For example, by applying the second-order interpolant pair,

$$s_A(t) = -\frac{\sin(\pi Bt)}{\pi Bt}\frac{\sin(\pi Bt - (n_1+1)\pi kB)}{\sin((2n_1+1)\pi kB)}, \tag{8}$$

and

$$s_B(t) = \frac{\sin(\pi Bt)}{\pi Bt}\frac{\sin(\pi Bt + (n_1+1)\pi kB)}{\sin((2n_1+1)\pi kB)}, \tag{9}$$

to $r_1^{\delta 2}(t)$, $m(t)$ is reconstructed at integer bandposition 0 ($0<f<B$), and is phase-shifted by $exp(jn_1\pi kB)$, see A.J.Coulson, R.G.Vaughan and M.A.Poletti "Interpolation in Bandpass Sampling" *Proc. ISSPA92*, vol.1 1992 pp.23-26. Thus, $m(t)$ has been effectively frequency-shifted from bandposition n_1 to bandposition 0.

The second-order interpolants, $s_A(t)$ and $s_B(t)$, phase-shift the two sample streams, $r_1^{\delta 2A}(t)$ and $r_1^{\delta 2B}(t)$, respectively, such that the replicas of the negative frequency components of $r_1(t)$ in $r_1^{\delta 2A}(t)$ will exactly cancel with their counterparts in $r_1^{\delta 2B}(t)$ upon addition. In other words, interpolation ensures that $R_1^{\delta 2A-}(f)$ exactly cancels with $R_1^{\delta 2B-}(f)$, in figure 2, over the region $0<f<B$. These interpolants do not correctly reconstruct $r_2^{\delta 2}(t)$ at bandposition 0, as the replicas of the negative frequency components of $r_2(t)$ will not cancel upon addition. A special case exists where $r_1(t)$ and $r_2(t)$ are sampled such that the interpolants will correctly reconstruct both $r_1^{\delta 2}(t)$ and $r_2^{\delta 2}(t)$ at bandposition 0. This case is when the complex envelopes of $r_1(t)$ and $r_2(t)$ are in-phase at the sampling times, such that

$$k = \frac{a}{(n_2-n_1)B}, a \in \mathrm{I}. \tag{10}$$

The interpolated signal,

$$g(t) = r_1^{\delta 2A}(t) \star s_A(t) + r_1^{\delta 2B}(t) \star s_B(t)\,, \tag{11}$$

may be complex frequency-shifted to the low-pass position ($-\frac{1}{2}B<f<\frac{1}{2}B$), by multiplying by $e^{-j\pi Bt}$. Where the interpolants are implemented using digital filters, the complex frequency-shift can be implemented digitally. This special case is only useful when the contributing signals are aligned. This alignment constraint can be removed by doubling the sampling rate at the expense of halving the available spectral positions for the contributing signals.

Signal Combination using Bandpass Sampling

By doubling the sample rate to $f_S=4B$, the requirement for the second-order interpolants, $s_A(t)$ and $s_B(t)$, is eased and implementation becomes more straightforward. In this case, the interpolants need only to act as windowing functions. For $r_1^{\delta 2}(t)$ from equation (6) but sampled at $f_S=4B$, a second-order sampling pair can be derived which ensures that the replicas of $M(f)$ in $R_1^{\delta 2A}(f)$ and $R_1^{\delta 2B}(f)$ add constructively over the frequencies of interest. Following the method of Linden, the constraint for constructive combining over $0<f<B$ is given by

$$BM\left(f-\frac{B}{2}\right)S_A(f) = \gamma^X M\left(f-\frac{B}{2}\right) \tag{12}$$

and

$$BM\left(f-\frac{B}{2}\right)\gamma^{-n}S_B(f) = \gamma^X M\left(f-\frac{B}{2}\right) \tag{13}$$

where γ^X is the phase of the interpolated signal. Solving for $S_A(f)$ and $S_B(f)$,

$$S_A(f) = \frac{\gamma^X}{B} \tag{14}$$

and

$$S_B(f) = \frac{\gamma^{X+n}}{B}. \tag{15}$$

Solving for X such that $S_A(f)=S_B^*(f)$, which ensures that $s_A(t)=s_B(-t)$ thus reducing hardware requirements for digital implementation, and performing Fourier transforms

$$\begin{aligned} s_A(t) &= \int_0^B \frac{e^{jn_1\pi kB}}{B} e^{j2\pi ft} df + \int_{-B}^0 \frac{e^{-jn_1\pi kB}}{B} e^{j2\pi ft} df \\ &= \frac{2\sin(\pi Bt)}{\pi Bt}\cos(\pi Bt + n_1\pi kB), \end{aligned} \tag{16}$$

and

$$\begin{aligned} s_B(t) &= \int_0^B \frac{e^{-jn_1\pi kB}}{B} e^{j2\pi ft} df + \int_{-B}^0 \frac{e^{jn_1\pi kB}}{B} e^{j2\pi ft} df \\ &= \frac{2\sin(\pi Bt)}{\pi Bt}\cos(\pi Bt - n_1\pi kB). \end{aligned} \tag{17}$$

These interpolants will reconstruct $m(t)$ by phase-aligning the positive frequency components of $r_1^{\delta 2}(t)$ at bandposition 0. After complex frequency-shifting, the low-pass signal is

$$m_{1D}(t) = m(t)\, e^{jn_1\pi kB}, \tag{18}$$

where $m_{1D}(t)$ denotes that $m(t)$ has been demodulated from $r_1(t)$. The phase-shift in equation (18) is proportional to n_1, which is the integer bandposition of $r_1(t)$.

The same interpolants, and complex frequency-shifting, will reconstruct $r_2^{\delta 2}(t)$ at the low-pass position as

$$m_{2D}(t) = m(t)\, e^{jn_2\pi kB} \cos((n_2 - n_1)\pi kB)\,, \tag{19}$$

It can be seen from equation (19) that, in addition to the phase-shift term, $exp(jn_2\pi kB)$, there is an amplitude-scaling term $\cos((n_2-n_1)\pi kB$. This results from the phase-shifting by $s_A(t)$ and $s_B(t)$ not, in general, phase-aligning the replicas of the positive-frequency components from $r_2(t)$. The phase-shift in equation (19) is proportional to n_2, the integer bandposition of $r_2(t)$. The second-order sampling and interpolating process imposes different phase-shifts on the two bandpass signals, according to the bandpositions of the pre-sampled bandpass signals and the bandposition(s) of the interpolated signals.

The sum of the two frequency-shifted and phase-shifted signals, $m_{1D}(t)$ and $m_{2D}(t)$, is

$$m_{1D}(t) + m_{2D}(t) = m(t)\left(e^{jn_1\pi kB} + e^{jn_2\pi kB}\cos((n_2 - n_1)\pi kB)\right). \tag{20}$$

An identical result is produced if the two bandpass signals, $r_1(t)$ and $r_2(t)$, are added before second-order sampling, interpolation, and complex frequency-shifting. Thus, the two bandpass signals are combined and independently phase-shifted using analogue addition, second-order bandpass sampling and interpolation. The amplitude-scaling term $\cos((n_2-n_1)\pi kB)$ in equation (20) constrains the phase-shifts that can be applied to $m_{2D}(t)$, in that some values of k, n_1 and n_2 will result in unacceptable attenuation of $m_{2D}(t)$.

For L bandpass signals which are copies of $m(t)$, defined by

$$r_i(t) = |m(t)|\cos(2\pi f_{Ci}t + arg[m(t)]),\ 1 \le i \le L, \tag{21}$$

at bandpositions n_1 to n_L, the summed signals can be frequency-shifted to bandposition 0 using second-order sampling and the interpolants, $s_A(t)$ and $s_B(t)$, from equations (12) and (13) respectively. Upon complex frequency-shifting to the low-pass position, the resulting signal is

$$m_D(t) = m(t)\sum_{i=1}^{L} e^{jn_i\pi kB}\cos((n_i - n_1)\pi kB)\,. \tag{22}$$

Comparison with equation (1) defines the complex combining weights. By varying k, the sample streams' separation, the phases of the bandpass copies of $m(t)$ can be varied quasi-independently during combination. Judicious choice of bandpositions can increase the degree of independence of the phase-shifts.

Equation (22) shows that this combining technique provides a means by which the signals may be phase-aligned during combination.

Detailed Description of Applications Using Combining of the Invention

A Combiner for Diversity Systems

Signal fading can be a significant problem in a wireless communications system. A major cause of signal fading is multiple path propagation. This is where a signal travels down a number of paths from transmitter to receiver. The effect of multiple path propagation on the signal envelope is multiplicative. The received signal, $Rx(t)$, may be expressed as

$$\begin{aligned} Rx(t) &= r(t)\, c(t,f) \\ &= r(t)\, A(t,f)\, e^{j\phi(t,f)}, \end{aligned} \tag{23}$$

where $c(t,f)$ is the frequency-dependent, time-varying multiple path channel which may be broken down into amplitude and phase components. For a stationary receiver, the multiple path channel sometimes is approximated as being time-invariant. For a mobile receiver, the multiple path channel is time-varying. This application is concerned with flat fading, in which the desired signal is narrow-band with respect to the fading channel.

Reduction of Signal Fading using Antenna or Frequency Diversity

A common remedy for signal fading is the use of antenna or frequency diversity. Antenna diversity is where a signal is received from a number of spatially diverse sensors. The sensor spacing is set such that the received signals have travelled through uncorrelated channels. By coherently combining the signals from the spatially diverse sensors, the effects of fading are reduced. Frequency diversity is where a signal is transmitted over a number of frequency diverse channels. As with space diversity, the channel (frequency) spacing is set to produce uncorrelated channels. Again, coherent combination is used to reduce the signal fading.

In many ways, the kernel of a diversity system is the signal combiner. A conventional phase-only signal combining system is shown in figure 3a. The system in figure 3a requires phase-detectors, variable phase-shifters and some form of adaptive controller to cohere the contributory branches. These items can be complicated and expensive to implement.

A Single Control Parameter Diversity System

A bandpass sampling coherent signal combiner of the invention for diversity systems is shown in figure 3b. The signals received by the antennas are amplified, filtered and demodulated, using conventional receiver technology, to individual integer-positioned intermediate frequencies (IFs) by Rx1 Rx2 and Rx3. In figure 3b, signals from three branches are combined, but the system can be implemented for any number of branches from two upwards. The separate branches are added, second-order bandpass sampled, and digitally interpolated and complex frequency-shifted. By adding the signals before sampling, the hardware requirements are reduced to one ADC and four digital interpolators per system, as shown in figures 3b and 4. The control mechanism of the combiner, labelled "Envelope Detect, Adaptive Control of k1, k2" in figure 4 is explained in the ensuing paragraphs.

Having fixed the integer bandpositions for the IFs, the signal combination control parameter is the second-order sample streams' separation, k, as can be seen from equation (22). In practice, k is quantized to a finite set of N levels. The combiner has a performance surface which allows adaptive control of k. This performance surface is the instantaneous envelope of the combined signal as a function of the relative phases of the pre-combined signals.

Figure 5a is a contour plot of the performance surface of a three branch system. The performance surface is also shown in three dimensions in figure 5b. Superimposed on the contour plot are the boundaries of the areas for which a particular value of k produces the maximum system output. The quantity NkB, where N is the number of quantization levels of k (16, in this case), indicates which value of k is used to produce each marked area. At any instant in time, the relative phases of the branches and the operating value of k specify a point on the performance surface. As the relative phases of the branches change, this point on the performance surface will migrate. As the point passes a boundary, the operating value of *k* will cease to be optimum, and the new optimum value must be found. A simple method of finding the new optimum value of *k* is to continually monitor the system outputs for those values of *k* adjacent to the operating value (for example *1*, *12*, *15*, *4* for the operating value $NkB=0$ in figure 5a). It is possible to reduce the processing so that only the system output, and a test output for an adjacent value of *k* are calculated at any one time. Where the test output is found to be larger than the system output, it is assumed that the value of *k* used to produce the test output is now optimum, and the test output becomes the system output. At this point, a new table of adjacent values of *k* is used to produce future test outputs. This table is produced using knowledge of the boundary pattern, of which figure 5a is an example. Where the test output is less than the system output, another adjacent value of *k* is chosen from the table to produce the next test output. In this way, the test output continuously cycles through the adjacent values of *k*, as indicated in figure 4.

The boundary pattern shown in figure 5a is a for 3-branch system where the number of quantization levels of *k*, *N*, was such that

$$N = \left(\frac{n_3-n_1}{n_2-n_1}\right)^2 . \tag{24}$$

For a given change in *k*, the resulting change in phase difference between $r_3(t)$ and $r_1(t)$ will be $(n_3-n_1)/(n_2-n_1)$ times greater than the change in phase difference between $r_2(t)$ and $r_1(t)$. By setting *N* according to equation (24), a lattice-type boundary pattern is obtained, such as that shown in figure 5a. This results in every area of best system performance for a given value of *k* having exactly four nearest neighbours, thereby minimising the search algorithm outlined above. For example, a system using bandpositions *16*, *18*, *24* produces the boundary pattern shown in figure 5a for $N=16$.

Simulation and Results

The bandpass sampling signal combination system was simulated for three branch diversity. The multiple path impaired channel was modelled as being Rayleigh distributed, and was uncorrelated between branches. The nominal signal bandwidth was 30 kHz and the average sampling rate, as defined in section 2.2, was 120 kHz. The radio frequency chosen was 851 MHz, which is a common mobile frequency. The three uncorrelated branches were demodulated to 495 kHz, 555 kHz and 735 kHz corresponding to integer bandpositions 16, 18 and 24, respectively. The number of quantization levels of k, N, was chosen to be 16, according to equation (24). The receiver was simulated as travelling at 0.14 m/s. For Rayleigh fading, the mean distance between deep fades is one half wavelength of the radio frequency carrier. This simulation, therefore, had approximately 120,000 samples per fade. The ADC was assumed to be perfect, with zero aperture-time, zero timing jitter, and IEEE double precision floating point dynamic range. The interpolators were simulated as being implemented by 51-tap finite impulse response (FIR) filters using a single, floating point DSP. The envelope detector, used to adapt k, was simulated using a 51-tap sliding rectangular window.

A second system was simulated to provide some indication of the performance of the bandpass sampling signal combination algorithm independently of the implementation. This system calculated the signal envelope in zero time, and could choose the optimum value of k, which most reduced the fading, in zero time. This system will be referred to as "the ideal bandpass sampling signal combiner".

Figure 6 is a Rayleigh diagram showing the results. All data was plotted with respect to the mean-square power of branch 1. The three receiver branches are plotted independently, and can be seen to follow Rayleigh distributed fading. These represent the worst-case performance for the system. A second reference is an ideal equal-gain combination of the simulation data. This represents the best performance of a combination system using phase control only in the absence of interference. All results are given for the 99% probability level.

It can be seen that the performance of the ideal bandpass sampling signal combiner is about 3 dB down on the performance of the ideal equal-gain combiner. The more realistic bandpass sampling signal combiner performance is 2 dB down on the performance of the ideal bandpass sampling signal combiner, giving a combiner loss" of about 5 dB. This is a significant improvement over any of the individual Rayleigh branches. Figure 7 shows the signal envelopes produced by these systems.

These results show that signal combination using bandpass sampling is an effective means of combatting slowly varying multiple path fading.

A Demodulator for Frequency-Hopping Spread Spectrum

Spread spectrum systems are used for a number of purposes. Frequency-hopping spread spectrum is generated by periodically altering the carrier frequency of the transmitter. In order to receive frequency-hopping spread spectrum, the receiver carrier is typically frequency-hopped at the same times as the transmitter carrier. The process of time-aligning the receiver carrier hops to the transmitter carrier hops is known as synchronization. To perform synchronization, the receiver must have *a priori* knowledge of the pattern of frequency hops. In addition, there will be phase discontinuities at the time of hopping, unless both the transmitter and receiver are able to perform phase-continuous frequency-hopping. Frequency-dependent fading also causes phase-discontinuities, as the propagation paths are different for different frequencies.

If the carrier frequencies used by the transmitter are limited such that the frequency-hopped signal is always integer positioned, it is possible to receive frequency-hopping spread spectrum using the bandpass sampling signal combiner of the invention. This system demodulates frequency-hopping spread spectrum without any need for synchronization, or phase-continuity from either the transmitter or the receiver.

Frequency-Hopping Minimum Shift Keying (MSK)

Signals for transmission by spread spectrum systems are normally transmitted as binary digits using a signal coding process such as minimum shift keying (MSK). This converts binary digits into a symbol set. In the case of MSK, the symbol set consists of one period of sinusoid at frequency f_1 (binary *0*), and one and a half periods of sinusoid at frequency $f_2=1.5f_1$ (binary *1*). The symbols are either non-inverted, or inverted, so as to produce a phase-continuous waveform. This minimizes the bandwidth of the transmitted signal. The baseband signal is modulated by transmitter carrier, which is periodically altered to produce a frequency-hopping spread spectrum signal.

The received signal is de-spread, using *a priori* knowledge of the pattern of frequency-hops. The de-spread signal is correlated with matched filters of the two symbols, and the outputs of the matched filters combined to reproduce binary digits. A conventional frequency-hopping spread spectrum receiver is shown in 8a.

Incoherent Demodulation Using Bandpass-Sampling

Many frequency-hopping spread spectrum systems operate using regularly spaced carrier frequencies. For such systems, it is possible to ensure that the received signal "bins" are integer positioned; by suitable choice of sampling rate, and nominal bandwidth at the receiver. This will allow the receiver to act as a signal combiner, combining the signals from the various frequency "bins" into a single bandpass, or low-pass signal.

Figure 8b shows a bandpass sampling signal combiner of the invention which may be used to demodulate an MSK signal which has been transmitted using frequency-hopping spread spectrum. This system quadrature samples the wideband received signal at a rate which is twice the bandwidth of the individual frequency bins. The in-phase (I) and quadrature (Q) sampling streams are each then digitally filtered using both matched filters and the Hilbert transform of the matched filters. The outputs of each Hilbert pair of matched filters are then square-law combined to produce a matched signal envelope. The magnitudes of the two sets of envelopes are subtracted to reproduce the data symbols. Conventional symbol timing recovery is then used.

The system was simulated for symbol frequencies 1200 kHz (binary 0), and 1800 kHz (binary 1). The transmitter hopped carrier frequencies every 20 symbols. The carrier frequencies used were 459 kHz, 465 kHz and 471 kHz. The carriers were not phase-continuous at the point of frequency-hopping, as shown by the phase-discontinuities in figure 9b. The received signal was simulated as having a signal-to-noise ratio of 10 dB in each bin, with out-of-bin noise assumed to be zero. Symbol timing recovery was not part of the simulated system.

Figure 9a shows the system output for a quadrature bandpass sampling frequency-hopping spread-spectrum demodulator. The dashed line represents the original symbols, and the solid line represents the detector output. It can be seen that the magnitude of the detector output is independent of the relative phases of the carrier and the sampling stream. This system was effective using frequency-hopping rates of up to one hop per symbol.

The simulation shows that bandpass sampling provides a simple, and effective, method of demodulating frequency-hopping spread spectrum.

A Switchable Pattern Antenna

Antenna arrays are commonly used to produce increased-performance antennas using simple elements. By demodulating the radio-frequency signals received by the elements to different, integer positioned intermediate frequencies, as described earlier, it is possible to provide a single control parameter switchable beam antenna array using second-order bandpass sampling. Such a system is shown in figure 10. The integer positioned signals were combined using the signal combiner of the invention such that the second-order sampling control parameter k was used to select the radiation pattern shape. It was found that the antenna radiation pattern shape could be substantially changed using this technique. For a small number of quantization levels of k, N, the shapes of the radiation patterns were quite different for each value of k. These radiation patterns correspond to a different interpretation of the diversity system technique, as described. The switching of antenna pattern shapes may have application in reconfigurable antenna systems, or for "null steering".

Simulation and Results

Radiation patterns for the antenna array shown in figure 10, were produced using computer simulation. The elements were assumed to be ideal monopoles, and the ADCs were also assumed to be ideal. Interpolation was done assuming infinite FIR filters, using an sampling rate of twice the nominal signal bandwidth. The intermediate frequencies were chosen such that the signal bandpositions were *20* to *23* for the four element array. The number of quantization levels of k, N, was *8*.

The radiation patterns produced are shown in figure 11. Results are shown for kB = 0, 1/8, 2/8, and 3/8 only. Succeeding values of kB produce patterns which are identical to those shown, but reflected about the horizontal axis. The results show that a set of radiation patterns having different can be obtained for linear antenna arrays using the bandpass sampling signal combiner. A given pattern can be selected by using the appropriate value of k.

Summary

The bandpass sampling combination technique uses the spectral replication property of the sampling process to perform the signal combination, and the phase-shifting property of second-order bandpass sampling to provide a bandposition dependent phase-shift. This new signal combination technique may be implemented using analogue-to-digital converters (ADCs), and digital signal processors (DSPs).

For diversity systems, the bandpass sampling signal combiner facilitates cohering of the branch signals. The system performs well for slow Rayleigh fading. For frequency-hopping spread spectrum, bandpass signal combining eliminates the need to synchronize the receiver to the frequency-hopping of the transmitter. For antenna arrays, antenna array pattern switching with a single parameter control is possible.

In each case, the bandpass sampling signal combiner may be implemented using simple and inexpensive digital systems.

The foregoing describes the combining system and method of invention, including examples thereof. Alternatives and modifications as will be obvious for those skilled in the art are intended to be incorporated in the scope hereof.

CLAIMS

1. A method of signal combining, comprising:

the addition of a number of received bandpass signals occupying different spectral positions to give a signal sum,

the spectral replication of the resulting signals using bandpass sampling, and

the interpolation over at least one frequency bandwidth of the combined and spectrally relocated signals, producing a combined digital signal.

2. A method according to claim 1, including first spectrally relocating the individual received signals to individual integer positioned intermediate frequencies where the individual received signals are not appropriately positioned for bandpass sampling and interpolation, prior to adding the signals and bandpass sampling.

3. A method according to either one of claims 1 and 2, wherein the bandpass sampling is second-order or higher order.

4. A method according to any one of claims 1 to 3, wherein the bandpass sampling rate is an integer multiple of the nominal bandwidth of the individual bandpass signals to be combined.

5. A method according to claim 4, wherein the bandpass sampling rate is four times the nominal bandwidth of the individual bandpass signals to be combined.

6. A method according to any one of the preceding claims, wherein the interpolation of the spectrally relocated bandpass signals resulting from the bandpass sampling consists of bandlimiting and phase-shifting the bandpass signals.

7. A system for combining signals received by a diversity system using space, frequency or time diversity, comprising:

means to demodulate the received signals to individual integer positioned intermediate frequencies,

means to add the intermediate frequency signals to give a signal sum,

means to bandpass sample preferably by second-order bandpass sampling the signal sum,

means to digitally or otherwise interpolate the replicated signal sum over at least one frequency bandwidth containing information from each received signal resulting from the bandpass sampling to provide a means to phase-control the component signals of the signal sum, and means to detect the performance of the system, deduce the component signal phases and thereby adaptively control the phases of the component signals of the combined signal sum using the second-order sampling parameter k to control the phases.

8. A system for demodulating a frequency-hopping spread spectrum signal without *a priori* knowledge of the frequency-hop pattern, comprising

means to bandpass sample the received bandpass signal preferably by quadrature or other second-order sampling,

means to digitally or otherwise interpolate the replicated signal over at least one frequency bandwidth containing the information from the received signal to produce the in-phase and quadrature components of the received signal, and

means to combine the in-phase and quadrature components of the received signal so as to detect the envelope of the signal.

9. A system for producing uncorrelated signals using an array of receiving antennas and bandpass sampling as a switchable pattern antenna, comprising

means to demodulate the signals from each element of the receiving antenna array to individual integer positioned intermediate frequency bandpositions, means to add the intermediate frequency signals to produce a signal sum,

means to bandpass sample the signal sum preferably using second-order bandpass sampling to produce a spectrally replicated signal sum,

means to digitally or otherwise interpolate the relocated signal over at least one frequency bandwidth to produce a combined signal with a unique radiation pattern which is selectable using the second-order sampling parameter k.

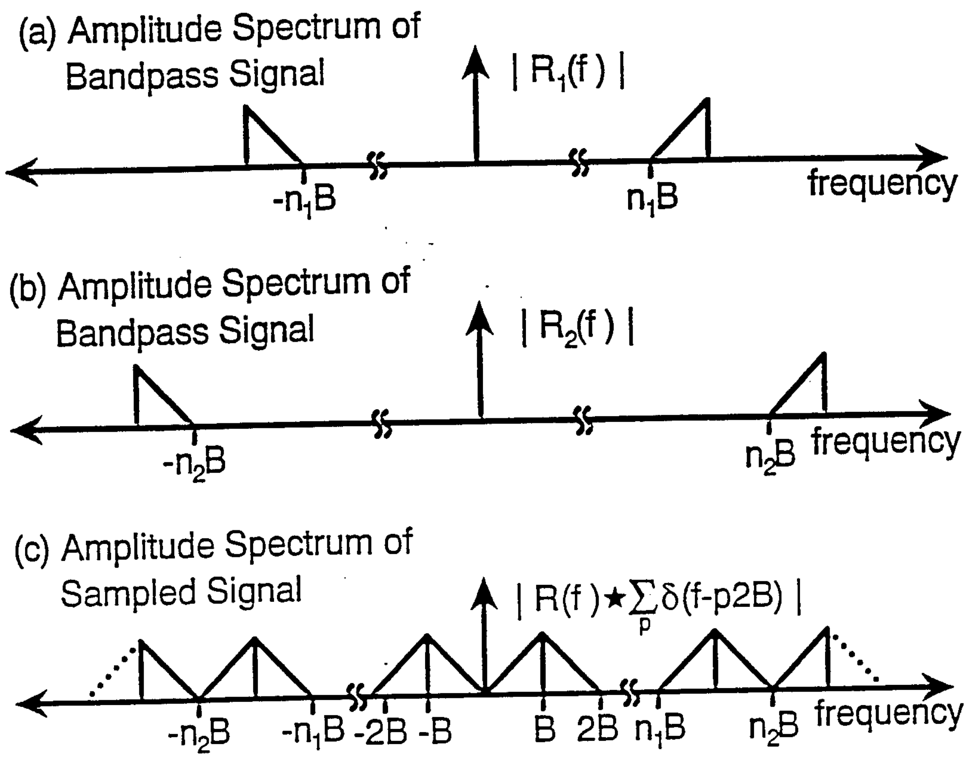

(a) Amplitude Spectrum of Bandpass Signal
| R1(f) |
-n1B
n1B
frequency
(b) Amplitude Spectrum of Bandpass Signal
| R2(f) |
-n2B
n2B
frequency
(c) Amplitude Spectrum of Sampled Signal
| R(f)★∑δ(f-p2B) |
p
-n2B
-n1B
-2B
-B
B
2B
n1B
n2B
frequency

Fig. 1

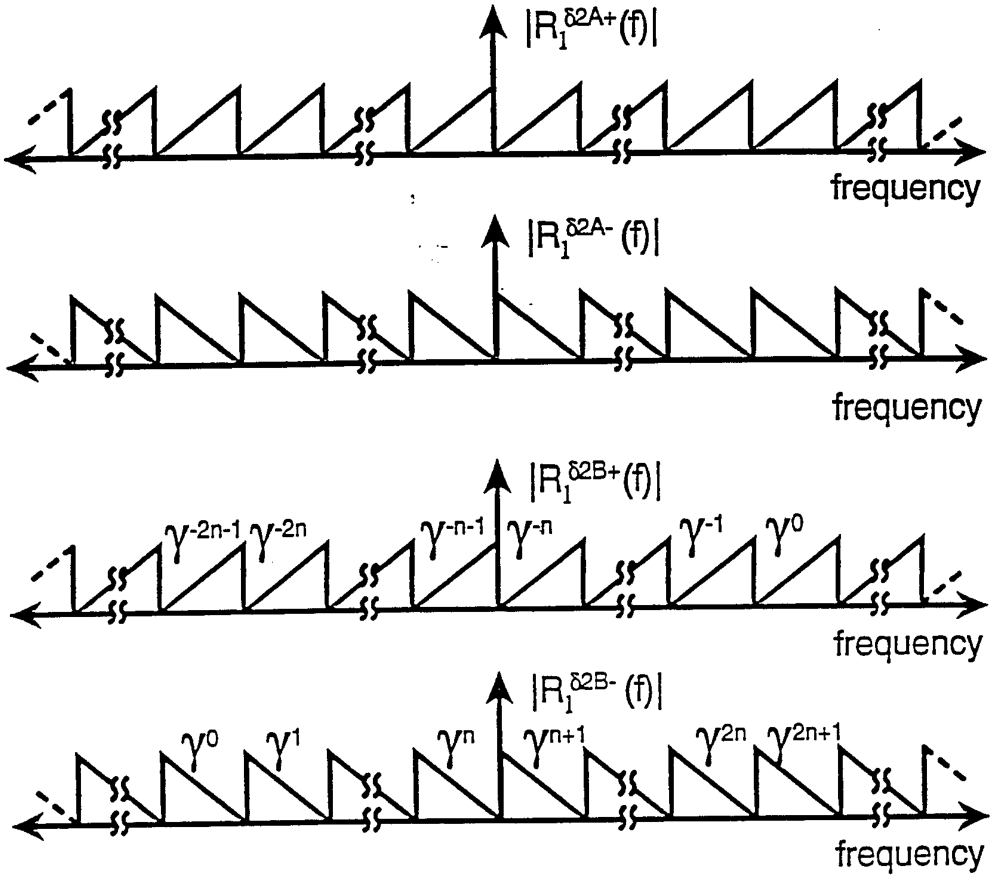

$|R_1^{\delta 2A+}(f)|$
frequency
$|R_1^{\delta 2A-}(f)|$
frequency
$|R_1^{\delta 2B+}(f)|$
γ^{-2n-1} γ^{-2n} γ^{-n-1} γ^{-n} γ^{-1} γ^{0}
frequency
$|R_1^{\delta 2B-}(f)|$
γ^{0} γ^{1} γ^{n} γ^{n+1} γ^{2n} γ^{2n+1}
frequency

Fig. 2

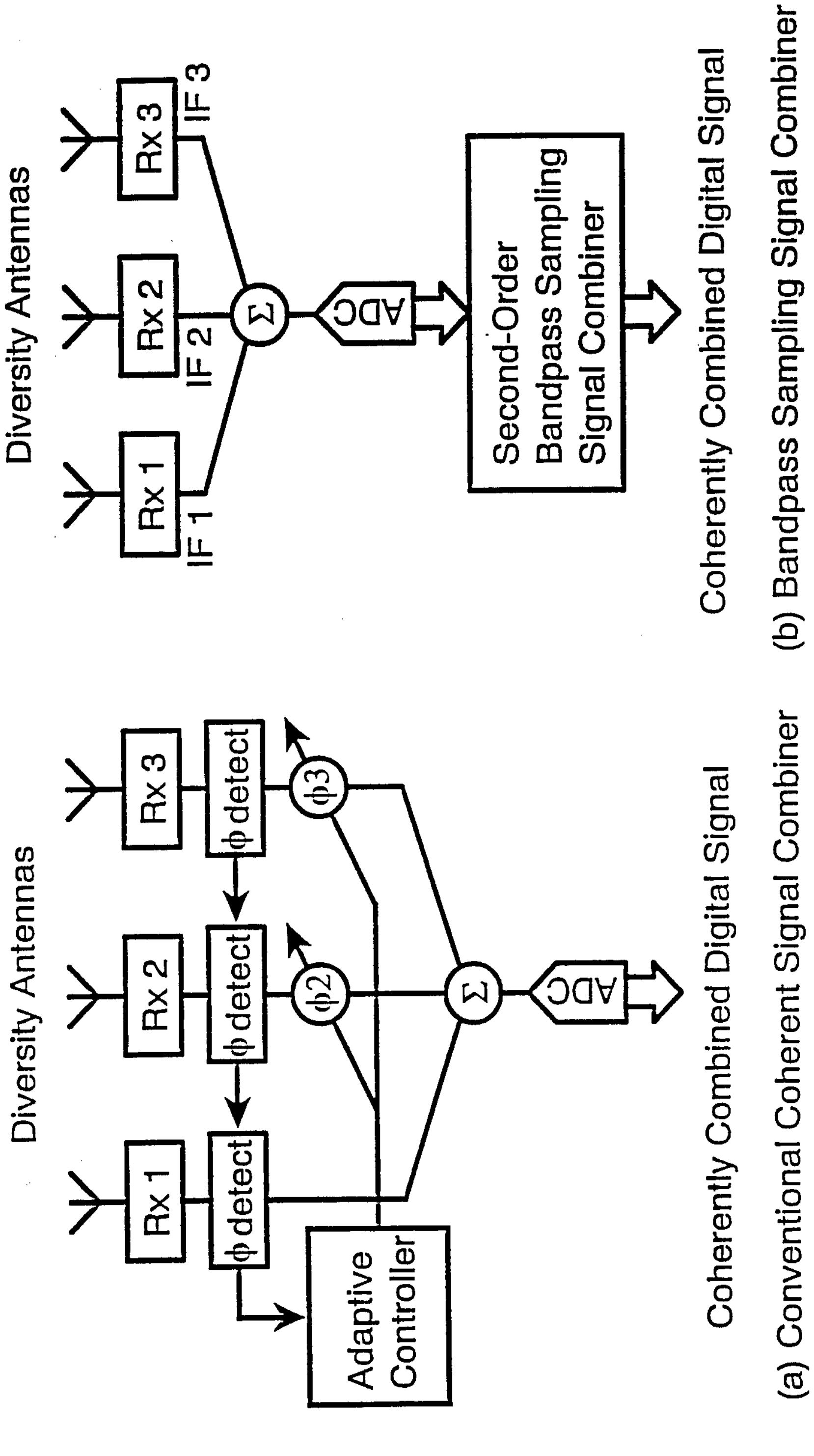
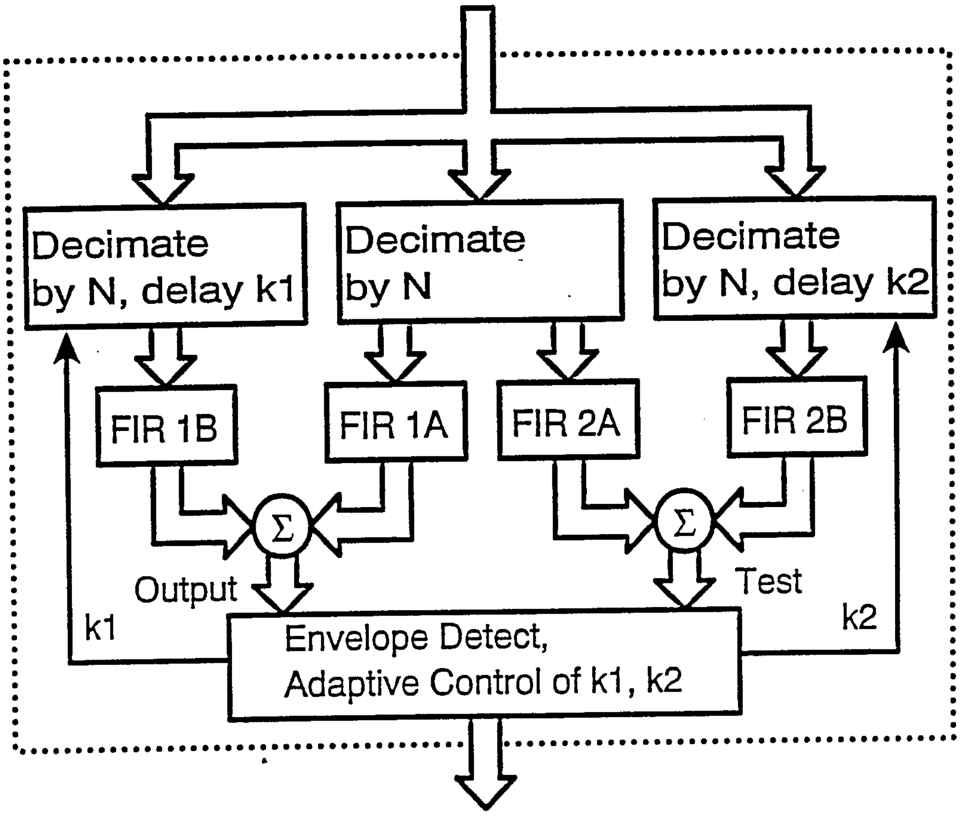

Diversity Antennas
Rx 1
Rx 2
Rx 3
φ detect
φ detect
φ detect
φ2
φ3
Σ
ADC
Adaptive Controller
Coherently Combined Digital Signal
(a) Conventional Coherent Signal Combiner
Diversity Antennas
Rx 1
Rx 2
Rx 3
IF 1
IF 2
IF 3
Σ
ADC
Second-Order Bandpass Sampling Signal Combiner
Coherently Combined Digital Signal
(b) Bandpass Sampling Signal Combiner

Fig. 3

Combined, Oversampled Digital Signal



Coherently Combined Digital Signal

Fig. 4

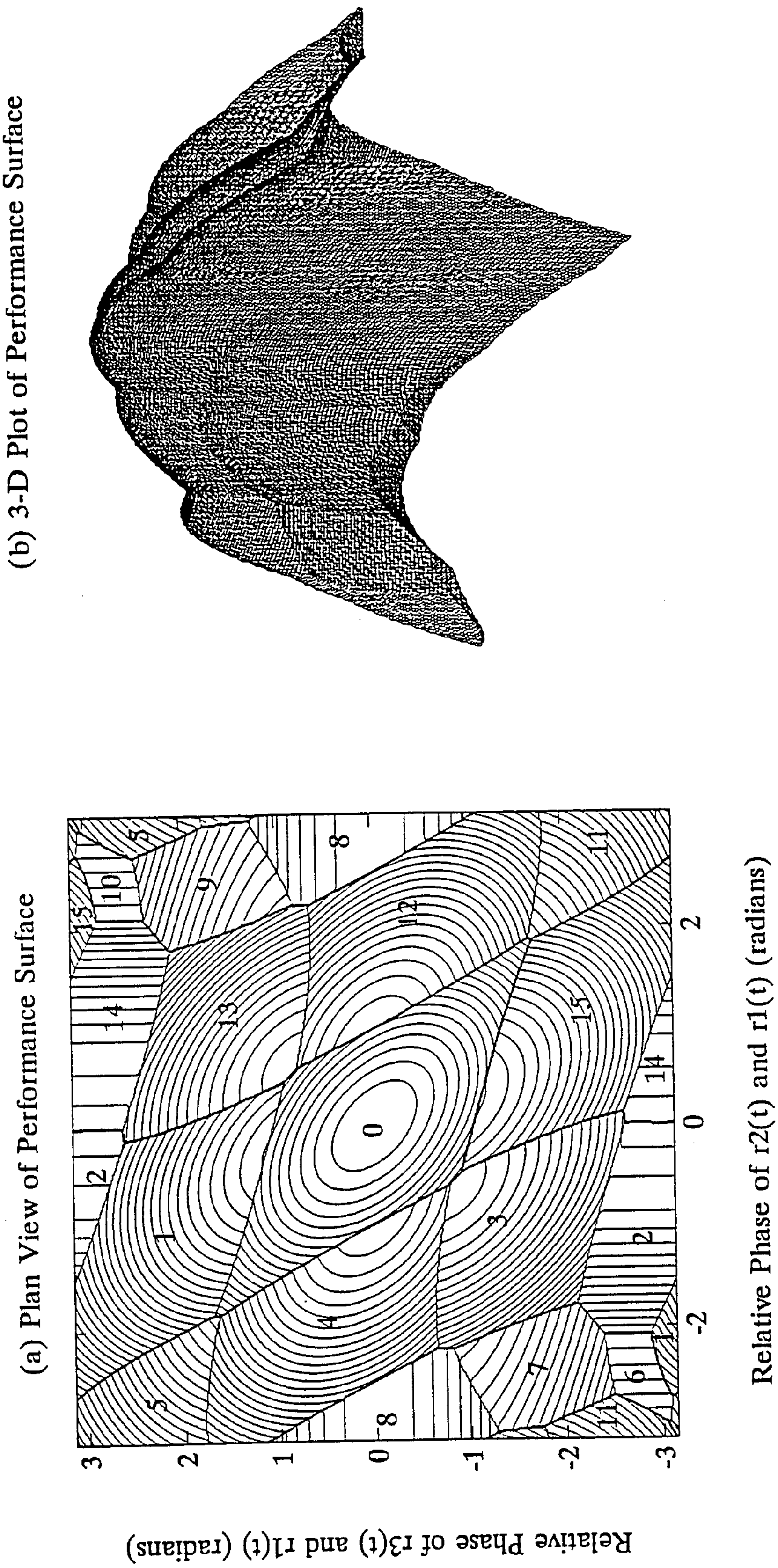

(a) Plan View of Performance Surface
Relative Phase of r2(t) and r1(t) (radians)
Relative Phase of r3(t) and r1(t) (radians)
-3
-2
-1
0
1
2
3
-2
0
2
(b) 3-D Plot of Performance Surface

Fig. 5

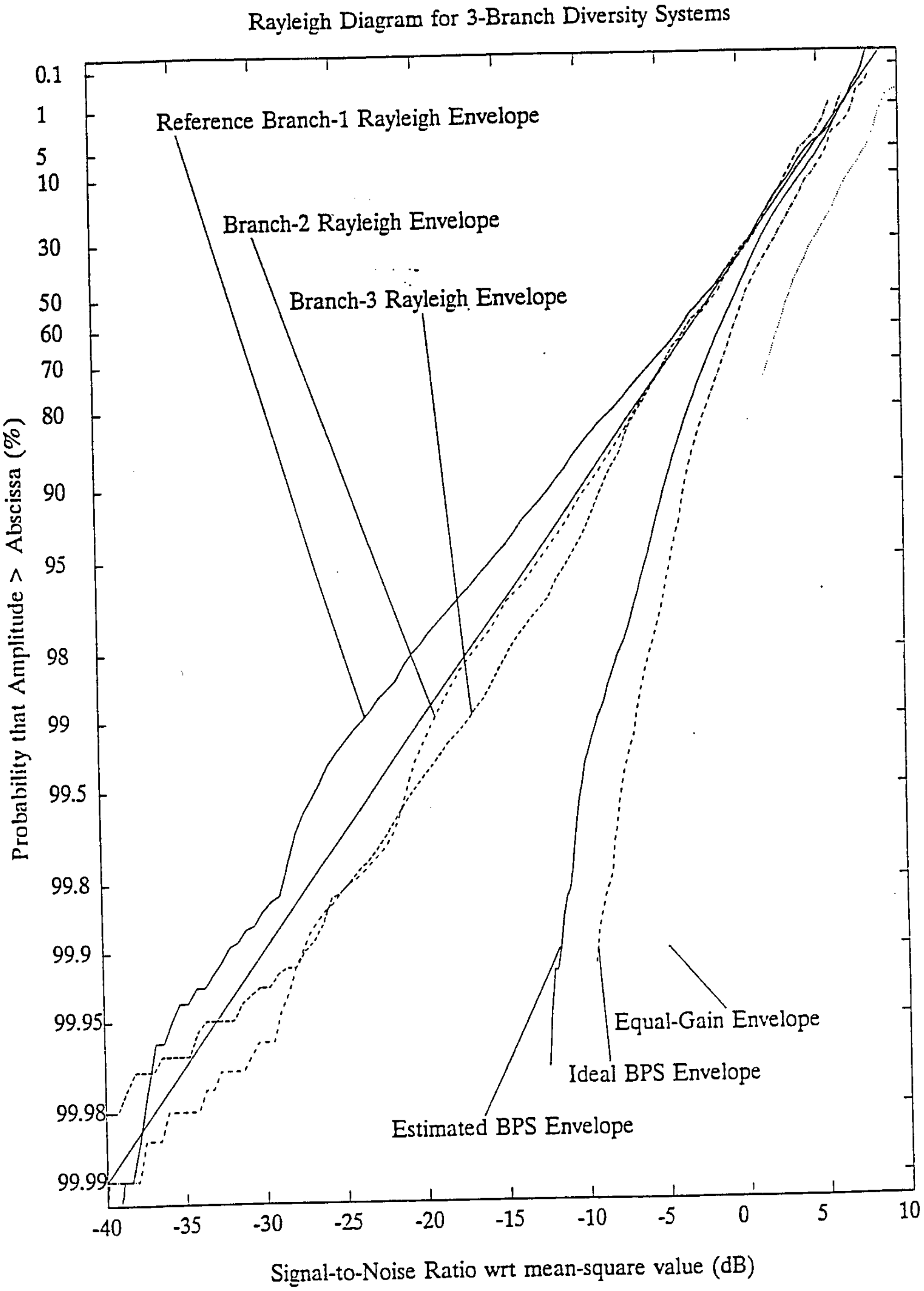

Rayleigh Diagram for 3-Branch Diversity Systems
Reference Branch-1 Rayleigh Envelope
Branch-2 Rayleigh Envelope
Branch-3 Rayleigh Envelope
Equal-Gain Envelope
Ideal BPS Envelope
Estimated BPS Envelope
Probability that Amplitude > Abscissa (%)
0.1
1
5
10
30
50
60
70
80
90
95
98
99
99.5
99.8
99.9
99.95
99.98
99.99
-40
-35
-30
-25
-20
-15
-10
-5
0
5
10
Signal-to-Noise Ratio wrt mean-square value (dB)

Fig. 6

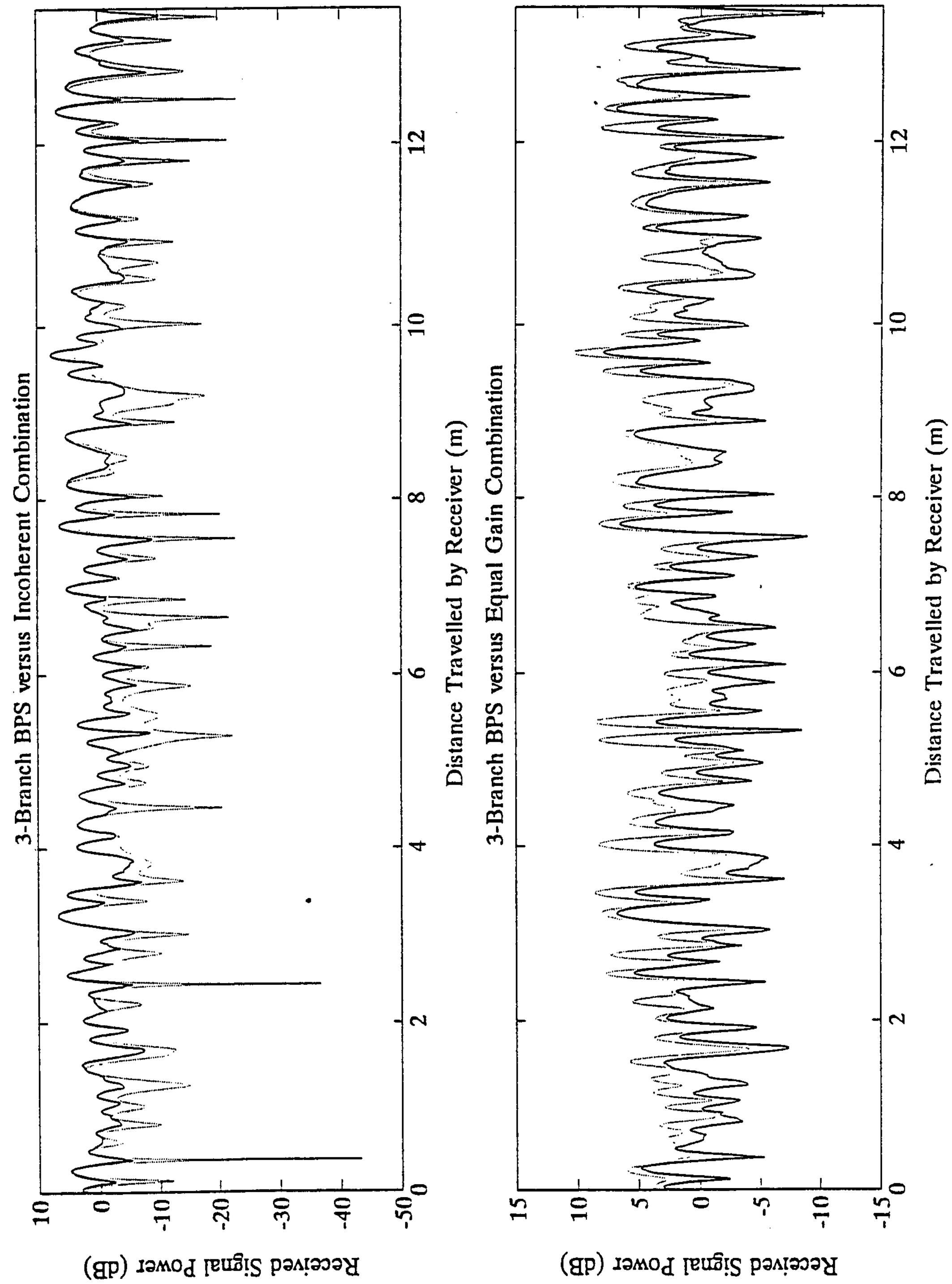

3-Branch BPS versus Incoherent Combination
Received Signal Power (dB)
10
0
-10
-20
-30
-40
-50
0
2
4
6
8
10
12
Distance Travelled by Receiver (m)
3-Branch BPS versus Equal Gain Combination
Received Signal Power (dB)
15
10
5
0
-5
-10
-15
0
2
4
6
8
10
12
Distance Travelled by Receiver (m)

Fig. 7

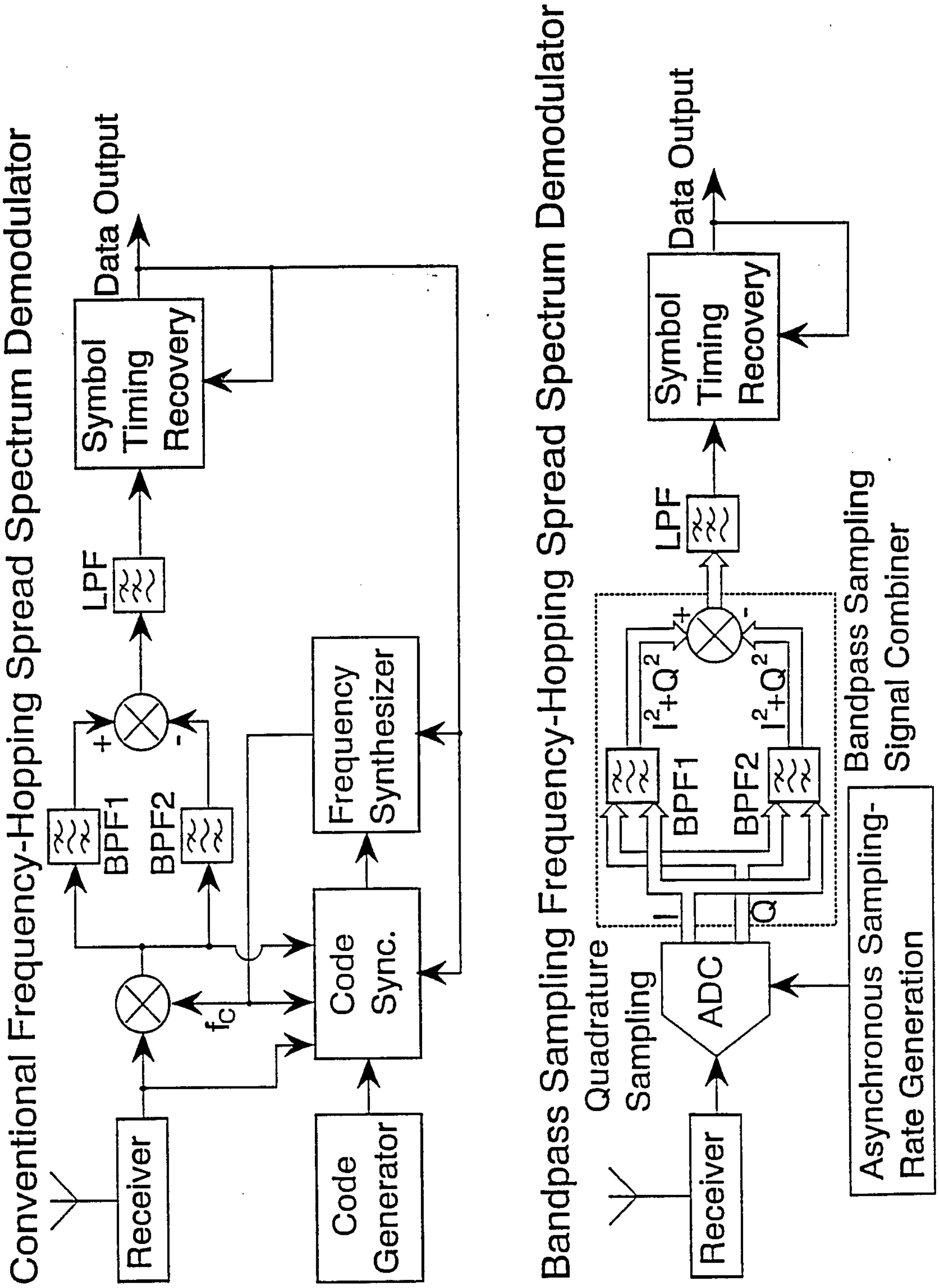

Conventional Frequency-Hopping Spread Spectrum Demodulator
Receiver
Code Generator
Code Sync.
Frequency Synthesizer
f_c
BPF1
BPF2
+
-
LPF
Symbol Timing Recovery
Data Output
Bandpass Sampling Frequency-Hopping Spread Spectrum Demodulator
Receiver
Quadrature Sampling
ADC
I
Q
BPF1
BPF2
I^2+Q^2
I^2+Q^2
+
-
LPF
Symbol Timing Recovery
Data Output
Bandpass Sampling Signal Combiner
Asynchronous Sampling-Rate Generation

Fig. 8

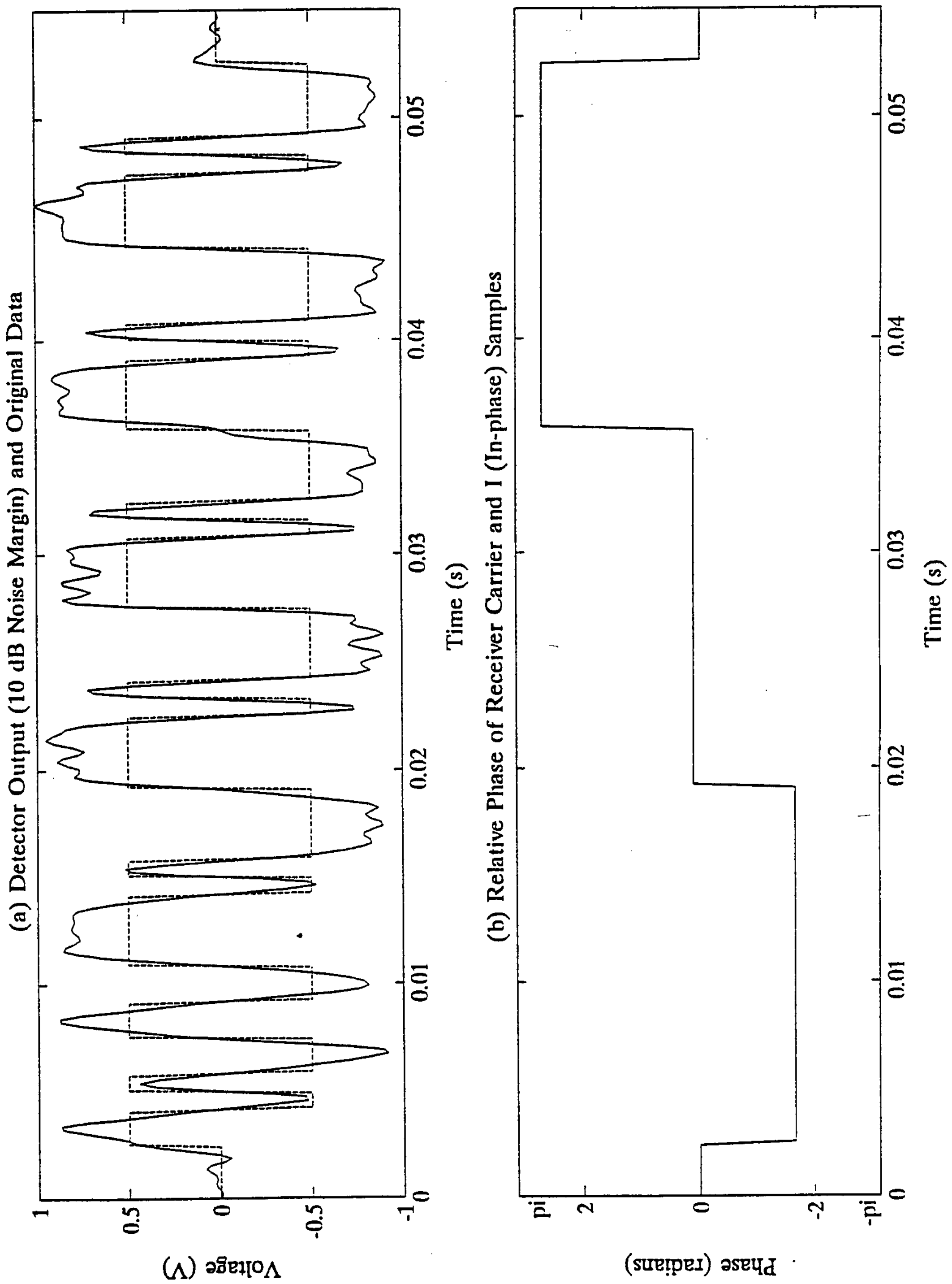

(a) Detector Output (10 dB Noise Margin) and Original Data
Voltage (V)
1
0.5
0
-0.5
-1
0
0.01
0.02
0.03
0.04
0.05
Time (s)
(b) Relative Phase of Receiver Carrier and I (In-phase) Samples
Phase (radians)
pi
2
0
-2
-pi
0
0.01
0.02
0.03
0.04
0.05
Time (s)

Fig. 9

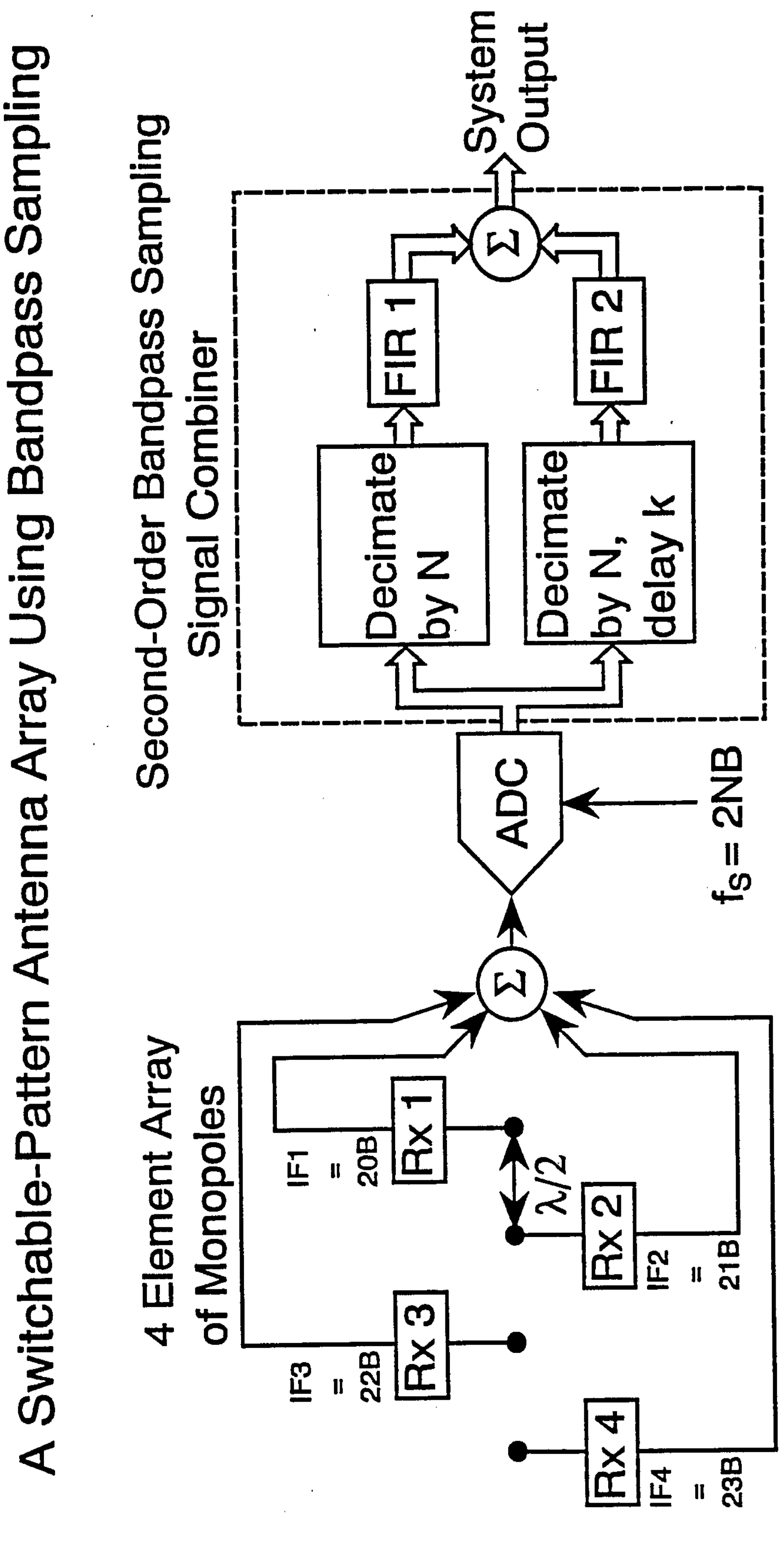

A Switchable-Pattern Antenna Array Using Bandpass Sampling
4 Element Array of Monopoles
Rx 1
Rx 2
Rx 3
Rx 4
IF1 = 20B
IF2 = 21B
IF3 = 22B
IF4 = 23B
λ/2
Σ
ADC
$f_S = 2NB$
Second-Order Bandpass Sampling Signal Combiner
Decimate by N
Decimate by N, delay k
FIR 1
FIR 2
Σ
System Output

Fig. 10

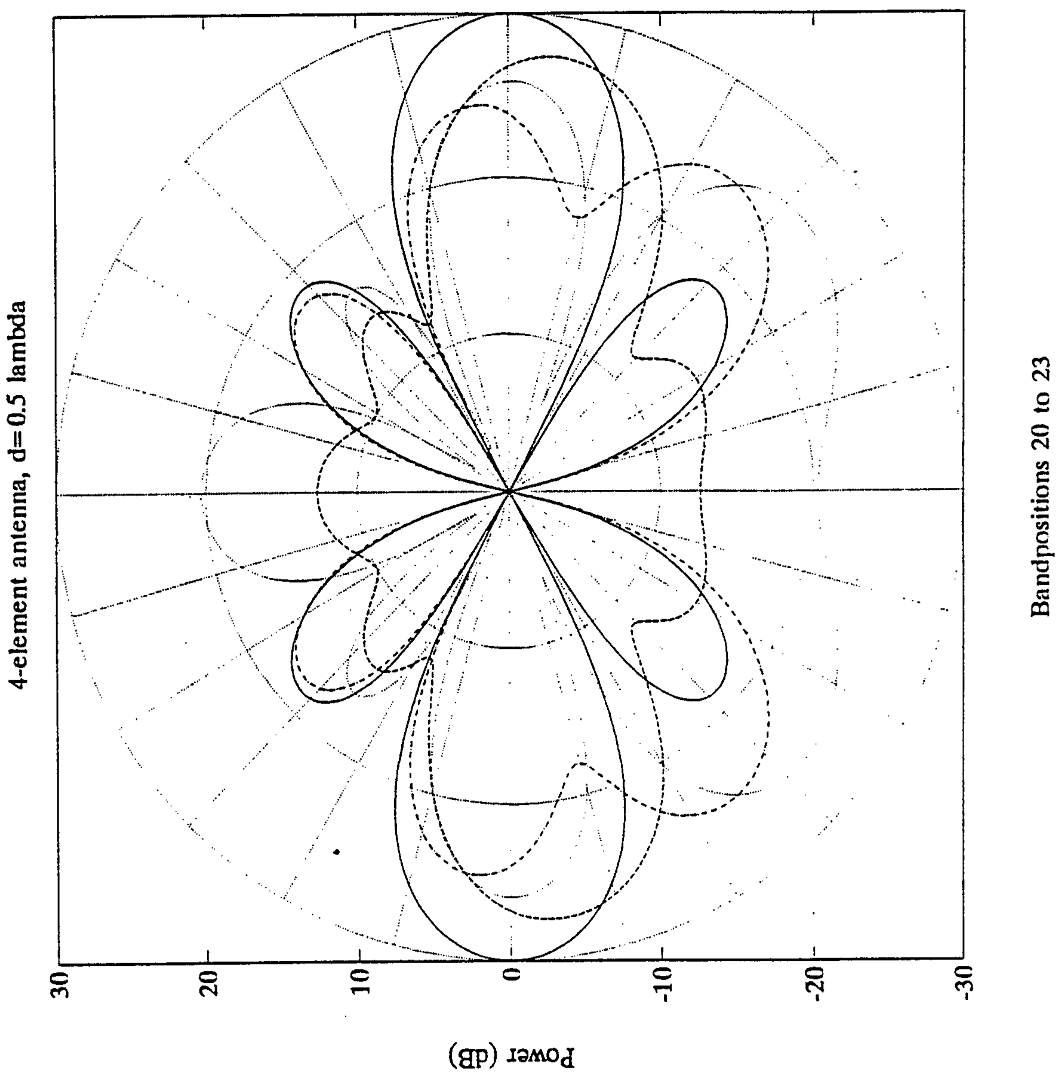


Fig. 11

SUBSTITUTE SHEET

International application No.
PCT/NZ 94/00083

A. CLASSIFICATION OF SUBJECT MATTER

Int. Cl.[6] H04L 27/00 H04B 7/08, 7/12, H03D 9/00 H01Q 21/28

According to International Patent Classification (IPC) or to both national classification and IPC

B. FIELDS SEARCHED

Minimum documentation searched (classification system followed by classification symbols)
IPC H04L 27/00, H04B 7/08, 7/12, H03D 9/00, H01Q 21/28

Documentation searched other than minimum documentation to the extent that such documents are included in the fields searched
AU: IPC as above

Electronic data base consulted during the international search (name of data base, and where practicable, search terms used)
DERWENT: Bandpass sampling

C. DOCUMENTS CONSIDERED TO BE RELEVANT

Category*	Citation of document, with indication, where appropriate, of the relevant passages	Relevant to Claim No.
A	IEEE Transaction on Signal Processing, Vol. 39, No. 9, issued September 1991. R.G. Vaughan et al., "The Theory of Bandpass Sampling", pages 1973-1983, Whole document	1-9
A	EP,A, 327268 (American Telephone and Telegraph Company) 9 August 1989 (09.08.89) Whole document	1-9

☐ Further documents are listed in the continuation of Box C.

☒ See patent family annex.

* Special categories of cited documents :

"A" document defining the general state of the art which is not considered to be of particular relevance
"E" earlier document but published on or after the international filing date
"L" document which may throw doubts on priority claim(s) or which is cited to establish the publication date of another citation or other special reason (as specified)
"O" document referring to an oral disclosure, use, exhibition or other means
"P" document published prior to the international filing date but later than the priority date claimed

"T" later document published after the international filing date or priority date and not in conflict with the application but cited to understand the principle or theory underlying the invention
"X" document of particular relevance; the claimed invention cannot be considered novel or cannot be considered to involve an inventive step when the document is taken alone
"Y" document of particular relevance; the claimed invention cannot be considered to involve an inventive step when the document is combined with one or more other such documents, such combination being obvious to a person skilled in the art
"&" document member of the same patent family

Date of the actual completion of the international search	Date of mailing of the international search report
7 December 1994 (07.12.94)	21 Dec 1994 (21.12.94)

Name and mailing address of the ISA/AU	Authorized officer
AUSTRALIAN INDUSTRIAL PROPERTY ORGANISATION PO BOX 200 WODEN ACT 2606 AUSTRALIA Facsimile No. 06 2853929	R. FINZI Telephone No. (06) 2832213

This Annex lists the known "A" publication level patent family members relating to the patent documents cited in the above-mentioned international search report. The Australian Patent Office is in no way liable for these particulars which are merely given for the purpose of information.

Patent Document Cited in Search Report		Patent Family Member					
EP	327268	CA	1298354	JP	2007710	US	4866647
							END OF ANNEX