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INTERNATIONAL APPLICATION PUBLISHED UNDER THE PATENT COOPERATION TREATY (PCT)

(54) Title: SPREAD SPECTRUM COMMUNICATION SYSTEM PARTICULARLY SUITED FOR RF NETWORK COMMUNICATION
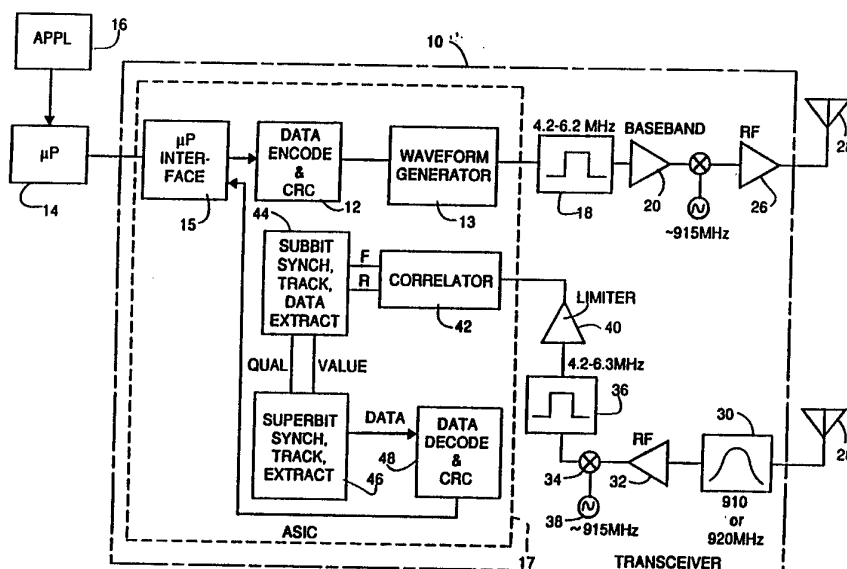
(57) Abstract

   The transceiver (10) receives data for transmission via a microprocessor (14) by means of an interface (15) under the control of application software (16). The data are transformed into chip sequences by the data encoder (12) and the waveform generator (13). The data then pass through a bandpass filter (18), a baseband amplifier (20), a mixer (22) connected to a local oscillator (24), an RF amplifier (26), and antenna (28). Similarly, the received signal goes through an RF bandpass filter (30), an RF amplifier (32), a mixer (34) connected to a local oscillator (38), a bandpass filter (36), a limiter (40), and a correlator (40). The decoded data packets from the receiver are fed back to the microprocessor (15).

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SPREAD SPECTRUM COMMUNICATION SYSTEM
PARTICULARLY SUITED FOR RF NETWORK COMMUNICATION
Background of the Invention

5   This application relates to spread-spectrum
communication systems and in particular to spread-
spectrum communication systems for use on noisy network
media, such as RF channels or AC power lines.

Spread-spectrum communication is a method whereby
10 information is communicated using a bandwidth that
greatly exceeds that required by information theory.
These methods provide signals over a wide bandwidth, and
with proper signal processing the communication is immune
to large amounts of noise within that bandwidth. In
15 chirp spread-spectrum methods, a signal burst known as a
chirp is transmitted. Each chirp has energy spread
across a frequency range. The frequency spread may be
achieved by frequency sweeping or by such techniques as
direct sequence coding. Chirps may be sent
20 asynchronously, or at synchronous intervals, including as
concatenated chirps. Data modulation of the chirp stream
can be accomplished by means such as phase reversal
modulation of the chirps or reversal of the frequency
sequence of the chirp. A transversal filter in the
25 receiver is matched to the chirp(s) expected, enabling
individual chirps to be detected even on noisy network
media such as power lines.

In a communication network, several transmitters
and receivers may communicate with each other over a
30 network medium. In certain networks, contention
resolution and/or collision detection strategies are
implemented to resolve situations in which two or more
transmitters simultaneously require use of the network
medium.

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In U.S. Patent No. 5,090,024, entitled Spread-
Spectrum Communications System for Networks, and
copending application, U.S. Serial No. 07/863,213,
entitled Timing for Spread-Spectrum Communication Across
5 Noisy Media, systems are proposed for applying spread-
spectrum communication to carrier-sense, multiple-access
networks. Various embodiments are disclosed, including
ones in which frequency-swept chirps are transmitted
using ASK modulation for contention resolution and PRK
10 modulation for data transmission.

Spread spectrum communication has been used for
wireless, RF communication in local area networks. But
these prior art RF systems have used pseudorandom direct
sequence (typically maximal linear sequence) techniques,
15 in which continuous synchronization of transmitter and
receiver is required and maintained. These systems have
generally required careful phase tracking mechanisms, and
have been intolerant of any frequency offsets between
transmitter and receiver carrier signals.

20   Summary of the Invention
In a first aspect, the invention features spread
spectrum communication using direct sequences that
approximate a swept frequency waveform in which
successive square waves are formed by the chips making up
25 the sequence, and in which the durations of the square
waves extend over a plurality of chips, and in which the
frequency of the square waves varies approximately
progressively across at least a portion of the sequence.
In preferred embodiments, information is encoded by
30 varying the direction of the frequency sweep in the
direct sequences, so that the sequences have either a
forward or a reverse frequency sweep; the spectrum of the
direct sequences is in a passband spaced from baseband;
the direct sequence is used to modulate a carrier (e.g.,

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biphase modulation of an RF carrier) to produce a spread-
spectrum signal also having a passband characteristic
(preferably two passbands centered about the frequency of
the modulated carrier). In other preferred embodiments,
5 the direct sequences are transmitted directly without any
modulation of a carrier.

In a second aspect, the invention features spread
spectrum communication using direct sequences that have
been chosen to produce a spread spectrum signal with a
10 passband spectral characteristic. In preferred
embodiments, this is achieved by the above-described
technique of using as the direct sequence an approximate
frequency swept signal.

In a third aspect, the invention features spread
15 spectrum communication in which frequency-swept spread
spectrum signals with different sweep directions convey
information across the channel. In preferred
embodiments, two spread spectrum signals one being the
same as the other except for the direction of sweep are
20 used; the frequency sweep is achieved using the above-
described technique of using as the direct sequences
forward and reverse swept approximations of a frequency
swept signal. Alternatively, frequency swept chirps
could be directly transmitted.

25   In a fourth aspect, the invention features spread
spectrum communication in which the transmitted signals
are forward and reverse swept signals, the received
signals are fed to a correlator having forward and
reverse sequence outputs, the outputs of the correlator
30 are independently tracked, and generally different
sampling times are used for taking samples of the forward
and reverse correlator outputs for purposes of making
decoding decisions. In preferred embodiments, the
forward and reverse swept signals are generated using a
35 direct sequence transmitted in forward or reverse

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directions; and the receiver uses a local oscillator for demodulation that is not synchronized in frequency or phase with the local oscillator used for modulation in the transmitter.

5   In a fifth aspect, the invention features direct sequence spread spectrum communication in which the overall phase of successive direct sequences is pseudorandomly varied to produce a more uniform spreading of energy across the frequency band of the transmitted

10 signal, and further reduce potential phase cancellation by interfering continuous wave signals. In preferred embodiments, each successive direct sequence has its overall phase pseudorandomly varied with respect to prior direct sequences.

15   In a sixth aspect, the invention features spread spectrum communication with dual level coding. Information is transmitted as "superbits", which are encoded as unique sequences of "subbits", each represented by a unique spread spectrum signal.

20 Correlator outputs in the receiver are used to generate both value and quality indicators at the subbit level, with the quality indicator providing a measure of the likelihood that the value indicator is correct. The value and quality indicators are combined in decoding the

25 superbits. In preferred embodiments, superbits are decoded using, in effect a superbit correlator which compares the value bits to unique subbit sequences associated with superbits, and combines the results of that comparison with the quality bits to generate for

30 each successive subbit a superbit correlator output. Superbit tracking is initiated when the superbit correlator is above threshold, and subsequent sampling for superbit decoding is done by sampling at intervals spaced by the length of a superbit.

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In a seventh aspect, the invention features spread
spectrum communication using a direct sequence and a
reverse ordered version of the same direct sequence. In
preferred embodiments, the direct sequence is either
5 transmitted directly without modulation of a carrier
(e.g., for transmission across power lines), or the
sequence is used to modulate a carrier signal (e.g., in
the RF frequency range); the direct sequence approximates
a swept frequency waveform, in which successive square
10 waves are formed by the chips, and in which the durations
of the square waves extend over a plurality of the chips,
and in which the frequency of the square waves varies
approximately progressively across at least a portion of
the sequence, from a low to a higher frequency in a
15 forward swept sequence, and from a high to a lower
frequency in a reverse swept sequence; the direct
sequence has a spectrum with energy primarily in a
passband (i.e., spaced from baseband).

In an eighth aspect, the invention features
20 receiving and decoding a spread spectrum signal of the
type in which forward and reverse order direct sequences
are used to convey information. In preferred
embodiments, the correlator uses a shared shift register
and shared summing networks to model four matched
25 filters--forward 0 and 90 degree filters, and reverse 0
and 90 degree filters. The correlator design takes
advantage of the identity of the forward and reverse
ordered direct sequences. Because of this identity, the
matched filter for the forward ordered sequence is the
30 mirror image of the matched filter for the reverse
ordered sequence. To conserve on the size of the shift
register and summing networks, approximately half of the
taps in the shift register are assigned to represent the
0 degree filter sections and the remainder to represent
35 the 90 degree filter sections. Each half of the taps is

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further divided into a common group for which values are
common to both the forward and reverse ordered sequences
and a difference group for which values are opposite for
the forward and reverse ordered sequences, to produce a
5 total of four groups of taps (0-degree common, 0-degree
difference, 90-degree common, 90-degree difference).
Four summing subnetwork are used to produce a sum of each
of the four groups of taps. Each of the four filter
section outputs is readily formed by combining two of the
10 four subnetwork outputs. For example, the 0-degree
forward output might be formed by adding the 0-degree
difference sum to the 0-degree common sum, whereas the 0-
degree reverse output might be formed by subtracting the
0-degree difference sum from the 0-degree common sum.
15   The invention has a variety of advantages. It
provides a much improved communication technique for RF
communications in a network environment. It allows a
carrier-sense, multiple-access transceiver to be
constructed with inexpensive frequency generating
20 components. Decoding is insensitive to phase shifts and
frequency shifts resulting from variations in the
frequency of the receiver's local oscillator with respect
to the transmitter's local oscillator. There is no need
to synchronize the local oscillator used for demodulation
25 to the local oscillator used for modulation. Improved
spreading of energy across the spectral band is achieved
by pseudorandom phase modulation of the direct sequences,
with resulting decrease in transmitted peak signal power
at any one frequency. The use of dual-level coding
30 reduces the size of the matched filters while retaining
much of the processing gain of a longer direct sequence.
Multipath nulls, which can be a problem in RF
communications within buildings and elsewhere, can be
largely eliminated by transmitting two redundant
35 sidebands, so that decoding can proceed even if a

receiver is positioned at a null for one of the
sidebands.

        Other features and advantages of the invention
will be apparent from the following description of
5 preferred embodiments and from the claims.

        Brief Description of the Drawings

        FIG. 1 is a block diagram of an RF transceiver of
the invention.

        FIG. 2 is a block diagram of the waveform
10 generator logic of the transceiver.

        FIG. 3 is a block diagram of the superbit
synchronizing, tracking, and data extraction logic for
the transceiver.

        FIGS. 4A and 4B are plots of the forward and
15 reverse chip sequences used by the transceiver.

        FIG. 5 is a table showing the entire 360-chip
forward sequence.

        FIGS. 6A and 6B are plots of the forward and
reverse subbit sequences used by the transceiver.

20   FIG. 7 is a diagrammatic illustration of the
organization of a data packet.

        FIG. 8 is a diagrammatic illustration of the
modified superbit used in the preamble of the data
packet.

25   FIGS. 9A-9D are diagrammatic illustrations of the
effect of frequency shift on timing in the receiver.

        FIG. 10 is a block diagram of the correlator of
FIG. 1.

        FIG. 11 is an illustration showing operation of
30 the subbit tracking logic.

        FIGS. 12A-12D are graphs showing four subnetworks
used to construct the matched filters of the correlator.

        FIGS. 13A-13D are graphs showing the four matched
filters constructed from the subnetworks.

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Description of the Preferred Embodiments

An RF transceiver 10 is shown in Fig. 1. The
transmitter portion is shown at the top of the figure.
Data to be transmitted across the RF channel are

5 forwarded to the transmitter via a microprocessor 14
under control of application software 16. The data
(which are grouped in packets) are received by
microprocessor interface 15, and then encoded (with a
CRC) at block 12. The data are transformed into chip

10 sequences, with a 25.2 Mhz pulse frequency, by waveform
generator block 13. Blocks 12 and 13 are implemented,
along with other logic associated with the receiver, as
an ASIC. The chip sequence is passed through a 4.2-6.3
Mhz bandpass filter 18 to remove out of band energy, and

15 amplified by baseband amplifier 20. The output of the
amplifier is supplied to mixer 22, where the chip
sequence biphase modulates a 915 Mhz carrier generated by
local oscillator 24. The modulated carrier is then
supplied to RF amplifier 26 and antenna 28. The

20 transmitted signal consists of two 2.1 Mhz-wide bands
centered at approximately 910 Mhz and 920 MHz.

The receiver portion of the RF transceiver is
shown at the bottom of FIG. 1. The signal received from
RF antenna 28 is passed through RF bandpass filter 30,

25 which can pass either of the two transmitted bands. The
bandpass filtered signal is amplified at RF amplifier 32,
and demodulated by mixer 34 (supplied with the output of
a 915 MHz local oscillator 38) and bandpass filter 36
(approximately the same as 4.2-6.3 MHz filter 18 in the

30 transmitter). The demodulated signal is passed through a
limiting amplifier 40 to transform it into a two-level
signal, which is sampled and clocked at 4.2 MHz into
correlator 42. The correlator has forward direction
sequence and reverse direction sequence outputs F and R.

35 These are supplied to subbit synchronizing, tracking and

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data extraction logic 44, which outputs a quality and a
value bit for each decoded subbit. The quality and value
bits are supplied to superbit synchronizing, tracking and
data extraction logic 46, the outputs of which are the
5 superbits and a threshold-crossing indicator bit. Data
decoder 48 decodes the sequence of superbits, checks the
CRC for validity, and supplies decoded data packets to
the microprocessor 14 via interface 15.

       The transceiver uses direct sequence spread
10 spectrum modulation, in which a code sequence of 1's and
0's specifies biphase modulation of a carrier. The
sequences are not, however, conventional pseudorandom
direct sequences, or maximal linear sequences. Instead,
they approximate a swept frequency chirp, as shown in
15 FIGS. 4A and 4B. Each sequence contains 360 chips
occurring at a frequency of 25.2 MHz. A single code
sequence is used, and transmitted in either the forward
or reverse direction. In a forward direction sequence,
the frequency of the square waves begins at 4.2 MHz (six
20 chips per wave) and ends at 6.3 MHz (four chips per
wave). A reverse direction sequence is exactly the same
pattern transmitted in reverse order. The sequence is a
dither pattern as shown in the table in FIG. 5 is used to
slowly vary the frequency from 4.2 to 6.3 MHz. The
25 frequency spectrums of the forward and reverse direction
sequences are identical and cover a band extending from
approximately 4.2 to 6.3 MHz.

       In the receiver, two correlators, one for the
forward direction sequence, and one for the reverse
30 direction, are designed to correlate with these
approximate swept frequency sequences. Digital subbits
are thereby transmitted across the channel, with a "1"
being represented by a sequence in one direction, and a
"0" by the reverse direction sequence. In some
35 implementations data can be transmitted using just these

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subbits, with each bit in the data packet corresponded to
a subbit, but a further enhancement in processing gain is
achieved by grouping the subbits into maximal linear
sequence patterns, each seven (7) subbits in length. The
5 patterns are the exact inverse of one another, beginning
with three subbits of one state (111), followed by two
subbits of the other state (00), and then one subbit of
each state (10). The 1110010 pattern of subbits, shown
in FIG. 6A, corresponds to what is termed a 1 "superbit,"
10 and the inverse pattern 0001101 shown in FIG. 6B what is
termed a 0 superbit. The 1110010 pattern is, of course,
a sequence of forward and reverse subbit sequences,
FFFRRFR, where F is a forward direction 360-chip
sequence, and R is a reverse direction 360-chip sequence.
15 Similarly, the 0001101 pattern corresponds to RRRFFRF.
But the particular starting point of the pattern is
picked arbitrarily.

        The superbits correspond to the actual bits making
up the data packet and the preamble and packet framing
20 sequences. In some embodiments, e.g., the EIA CEBus
protocol, information is transmitted using pulse-width
modulation (using what are known as USTs, Unit Symbol
Times). Two superbits of the same state, preceded and
followed by a superbit of the opposite state, represent
25 one logical state (e.g., a "0" UST), whereas a single
superbit of the same state, preceded and followed by a
superbit of the opposite state, represents the other
logical state (e.g., a "1" UST).

        To more evenly spread the energy of the
30 transmitted sequences across the 2.1 MHz-wide band, a
pseudorandom modulation of phase with a maximal length 15
code sequence is applied to the final subbit sequences.
This is done by exclusive-OR combining the subbit
sequence with the repeating 15-bit sequence. The phase
35 modulation operates on the subbit level, either reversing

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or leaving unchanged the phase of each of the 360 chips
in the chip sequence for the particular subbit. The
relative primeness of the 15 bit sequence with the seven
subbit long sequences (1110010 and 0001101) further
5   enhances the randomness of the modulation of phase.

Using phase modulation in this manner is effective
in reducing continuous wave interference, by continually
changing the phase of the transmitted signal, and
improving the power spreading in band. Phase modulation
10 of the chip sequences does not impact the ability of the
matched filters to detect them, as the matched filters
are able to detect a forward or a reverse sequence
independently of the overall, absolute phase of the
sequence.

15   The combined effect of the forward/reverse chip
sequences, subbit groupings, and pseudorandom phase
modulation is a more uniform spreading of energy across
the 2.1 MHz band. All three components of the modulation
have an effect on the spectrum. Before modulation, all
20 of the energy is, of course, concentrated in a single
spectral line at the carrier frequency, 915 MHz. The
subbits, which occur at 70 KHz, divide the energy across
the 2.1 MHz-wide band into approximately 30 spectral
lines. The forward/reverse pattern of the superbits,
25 which has the effect of impressing a 10 KHz pattern on
the subbit-modulated signal, divides the energy in the
band into approximately 210 spectral lines. But the
energy level of the various lines is uneven due to the
approximation dithering of a swept frequency signal in
30 the basic 360 chip sequences and the combination of
forward and reverse direction sequences in the superbits.
The phase modulation, which impresses a 15-subbit phase
pattern that repeats at 4.7 KHz, increases the number of
effective lines to greater than 400, thereby further
35 spreading out the power levels of the components, so that

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the peak power density across the 2.1 MHz band is
reduced.

      The circuitry for generating the waveform is shown
in FIG. 2. Superbits enter the generator logic at 100,
5 and are exclusive-OR combined at gate 102 with the seven-
bit subbit sequence generated by polynomial generator
104. The output of gate 102 is supplied to 720 bit ROM
106, in which is stored for implementation convenience
separate copies of the chip sequence in forward and
10 reverse order (each requiring 360 bits). Eight bits of
the sequences are read out of the ROM at one time into
shift register 108. The chip sequences are exclusive-OR
combined at gate 110 with the fifteen-bit pseudorandom
phase modulation sequence generated by polynomial
15 generator 112. A 25.2 MHz clock and associated divide-
by-eight circuit 114 and counter 116 control readout of
the chip sequence from the ROM 106.

      FIG. 10 shows the organization of correlator 42.
The correlator functions as two matched filters, one for
20 detecting the forward direction sequence, and one for the
reverse direction sequence. The two matched filters
could be constructed independently, each with its own
shift registers and summing networks, but lower
manufacturing costs can be achieved by sharing the shift
25 register and portions of the summing networks between the
two matched filters.

      Because absolute phase is not preserved in
demodulation (as the local oscillator in the receiver is
not synchronized with the oscillator in the transmitter),
30 it is actually necessary to have two matched filters, a
0° and 90° filter, for each of the forward and reverse
sequences. If only one filter were provided for each
sequence, the filter would have no sensitivity to a
sequence received with precisely 90° phase shift. By
35 having both 0° and 90° filters, and using the largest of

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the outputs of the two filters, the correlator can be
made less sensitive to phase shift.

As each sequence is 360 bits long, and received at
a rate of 25.2 MHz, each of the four matched filters
5 (forward 0° and 90°, reverse 0° and 90°), if constructed
in a conventional manner, would need to be 360 bits long
and be capable of sampling the received signal at 25.2
MHz. To reduce the size of the matching filters, the
received waveform is undersampled at 4.2 MHz (at 238 in
10 FIG. 10A). This produces an aliasing equivalent to
mixing the 4.2-6.3 MHz signal with a 4.2 MHz waveform and
thereby transformed to a 0-2.1 MHz spectrum. To achieve
the necessary sampling accuracy for the 0-2.1 MHz signal,
the size requirement of the matching filters is reduced
15 to 60 bits, a 6:1 reduction. Just as importantly, the
ASIC can function at an overall lower frequency of 4.2
MHz, except for the waveform generator, where the 25.2
MHz signal is assembled.

To further reduce the complexity of the
20 correlator, the four matched filters are implemented with
a single 60 bit shift register 240, and with substantial
sharing of resources in the summing networks 242. The
symmetry between the forward direction and reverse
direction sequences makes much of the sharing possible.
25 The summing network is divided into four subnetworks,
each handling approximately 1/4th of the 60 bits in the
shift register, and the outputs of the subnetworks are
summed in different ways to produce the desired four
matching filter outputs. The shared filter is created in
30 the following way. Two forward direction sequences are
generated representing frequency swept waveforms that are
shifted 90 degrees from each other. The two waveforms
are then used to create two separate matched filters of
60 points each. A combined correlator filter is then
35 constructed by assigning to the 90 degree filter those

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points where its values are larger than the 0 degree
filter, and vice versa. Since the waveforms are 90
degrees offset from each other, the minimum and maximum
value points for the correlator filter of the 90 and 0
5 degree waveforms are also offset. Therefore, the more
significant points of each filter are saved. Those
points are represented by only one bit in this
implementation, representing a positive or negative
waveform value. The result is a correlation filter that
10 holds both the forward direction 0 and 90 degree points
and which can be used in two approximately equal
correlation summing networks.

      Since the reverse sequence correlator filter will
be the mirror image of the forward sequence, the reverse
15 filter is created by inverting in sequence the bits of
the respective 0 and 90 degree points. It will be
discovered that the mirror image reverse and the forward
filter points will have approximately half of their
points in common (the same value) and half different (the
20 opposite value). Thus, to represent the two directional
filters, one can divide the existing filter into two
parts--a common sub part and a difference subpart. To
now create a summing network for the forward correlator
filter, the common subpart and the difference subpart are
25 added. To create a summing network for the reverse
filter, the difference subpart is subtracted from the
common subpart. The above procedure is used for both the
0 and 90 degree portions of the correlator filter.

      The result is 4 subnetworks, Com0, Com90, Dif0,
30 and Dif90, that appropriately summed give the 0 and 90
degree correlator values for both the forward and reverse
sequences.

            For0 = Com0 + Dif0
            Rev0 = Com0 - Dif0
35   For90 = Com90 + Dif90

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Rev90 = Com90 - Dif90

The subnetworks, which are implemented using 59 taps of
the shift register, are described in FIGS. 12A-12D. Each
of the 59 taps (from 0 to 58) are shown along the

5   horizontal axis of each figure. Each tap has either a
+1, 0 or -1 value, wherein each +1 and -1 represents a
valid filter tap and each 0 the absence of that tap, for
that subnetwork. Each subnetwork uses approximately one-
fourth of the taps.

10   The resulting matched filters, formed by the
addition or subtraction of the subnetworks, according to
the expressions given above, are shown in FIGS 13A-13D.
The forward and reverse filters for the same phase are
exactly alike except inversely ordered, and the 0 and 90

15 degree filters for the same direction sequence use
different ones of the 59 taps.

     Organization of the subbit synchronizing,
tracking, and data extraction logic is shown in FIG. 3.
Separate forward and reverse sequence tracking logic is

20 necessary to accommodate differences in the times at
which the matched filters will detect forward and reverse
sequences. Before describing that tracking logic, it is
worthwhile going over briefly the reasons why it is
needed. Small variations in the frequency of the local

25 oscillators in the transmitters and receivers have the
effect of shifting the frequency content of the received
chip sequences to a slightly higher or slightly lower
frequency band. And these shifts result in shifts in the
time at which the matched filters detect the forward and

30 reverse sequences. These time shifts would not be a
problem, but for the fact that the direction of the time
shift for forward and reverse sequences is different.

     FIGS. 9A-9D illustrate the effect of oscillator
frequency shift on a received sequence. The chip

35 sequence in FIG. 9A is the original, forward swept

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sequence, in which the chips approximate a sweep from 4.2
MHz to 6.3 MHz. The alignment of the correlator at which
a peak correlator output occurs is shown diagrammatically
at C in the figure.

5   The chip sequence in FIG. 9B has been frequency
shifted by 200 KHz so that it begins at 4.0 MHz and ends
at 6.1 MHz. The effect of the frequency shift is to
change slightly the alignment of the correlator C that
produces a maximum output. Maximum correlator output

10 occurs when the left edge of the correlator C is aligned
with that portion of the sequence having a frequency
content of about 4.2 MHz. In the original, unshifted
sequence (FIG. 9A), that occurs when the correlator is
aligned exactly with the leading edge of the sequence.

15 But in the frequency-shifted sequence, it occurs slightly
into the swept pattern. This shift in alignment means
that the time at which the correlator output reaches a
maximum is shifted slightly to an earlier time. The
difference in time is suggested by arrows 208 and 210.

20 In the case of the frequency-shifted sequence, the time
at which the correlator output reaches a maximum
(suggested by arrow 210) is slightly earlier than the
corresponding time for the unshifted sequence (arrow
208).

25   FIGS. 9C and 9D show the same comparison between
an unshifted reverse sequence and a frequency-shifted
reverse sequence. The same phenomenon occurs, except
that the time at which the maximum correlator output
occurs is shifted later (instead of earlier) in time (as

30 suggested by arrows 212 and 214).
         If the direct sequences used were not frequency
swept, but were instead conventional pseudorandom
sequences (maximal linear sequences), the effect of
frequency shift would not be a simple time shift in

35 correlator output, but a severe (or fatal) degradation in

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correlation. The reason for this is that conventional
pseudorandom sequences have frequency components
distributed widely throughout the duration of the
sequence, which means that a frequency shift will not
5   simply shift the alignment between the correlator and the
received sequence, but so corrupt the received signal as
to largely destroy all correlation.

The subbit synchronizing, tracking, and data
extraction logic 44 functions as follows. Each
10 correlator output F, R is fed to threshold detection
logic. The correlator outputs range in value from 0 to
15. A threshold, e.g., 10 is used for making an initial
detection of "carrier", i.e., initial detection of a
forward or reverse sequence. If the threshold is met or
15 exceeded, the threshold logic goes through a procedure to
assure that it has located the local peak in the
correlator output. It looks ahead by one 4.2 MHz clock
interval, and if the correlator value one clock ahead is
larger, it adjusts the clock accordingly, and looks ahead
20 again. The procedure is followed until the correlator
value ahead is smaller. At that point, both a forward
and a reverse tracker are initiated by tracking and
timing logic.

FIG. 11 illustrates operation of the tracking
25 logic following the initial detection of an above
threshold output in the forward correlator output. Both
the F and R trackers are started at that location (more
precisely at the location of the peak above-threshold
output), as suggested by window brackets 300, 302. At
30 this point, the forward tracker is properly synchronized,
and waits to examine the contents of a 3-bit wide window
304 centered 60 4.2 MHz clock times (60 bits of the
correlator, approximately 14 μsec) from the peak that
initiated the tracker. The reverse tracker has not,
35 however, synchronized itself with the reverse correlator

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outputs, which are generally skewed relative to the
forward correlator outputs because of the effect of
frequency shifts introduced by demodulation. Therefore,
beginning 30 clock times after the trackers are initiated
5 (as suggested at location 306) the reverse tracker looks
for an above threshold reverse correlator output. If one
is found prior to 60 clock times, as suggested at window
bracket 308, then the same rules for selecting the
largest correlator output (as given above for the forward
10 correlator) are used, and the reverse tracker is reset to
start its 60 clock counter at the location of that
largest correlator output. If an above threshold reverse
correlator output is not found by expiration of the 60
clock period, then the forward correlator output is
15 sampled (largest value within 3-clock interval window
centered at 60 clock intervals), and the reverse tracker
continues to look for an above threshold output over the
next 30 clock times, until location 310. If one is
found, then the reverse tracker is reset to start at that
20 location. From this point on, the two trackers function
independently, each looking in a 3-clock wide window at
60 clock intervals and choosing the largest value in the
window. The inverse of this procedure occurs if it is an
above threshold reverse correlator output that starts
25 both trackers.
        Each tracker selects the largest output within the
3-clock wide window centered at 60 clocks from the last
sample, and provides the output to data extraction logic.
The data extraction logic compares the forward and
30 reverse outputs, and decides whether a subbit is a
forward or reverse sequence on the basis of which output
is larger (ties are decided using an arbitrary but
consistent rule, e.g., by choosing forward). The data
extraction logic then outputs a 1 or a 0 for the "value"

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bit (with a 1 designating a forward sequence, and a 0 a
reverse sequence).

　　　The trackers also adjust their tracking at each 60
clock interval.　Each tracker determines, within the 3-
5　clock wide window, whether it is the lefthand, center, or
righthand correlator value that is the largest.　That
determination is used to nudge the clock timing of the
winning tracker (i.e., the one corresponding to the
decision made by the data extraction logic) by +1, 0, or
10　-1 clock time.　E.g., if the data extraction logic has
decided the subbit is a forward sequence, an adjustment
is made to the timing of the forward tracker but not the
reverse tracker.　These small clock timing adjustments
take care of the timing differences that occur because of
15　variations between transmitter and receiver clocking
circuitry.

　　　The data extraction circuit also outputs a
"quality" value, which is simply a 1 if the decision was
made on the basis of an above threshold output, and a 0
20　if the decision was made on a below-threshold basis.
Thus, for each decoded subbit there is a value bit
(indicating whether the subbit is decoded as a forward or
reverse sequence) and a quality bit (indicating whether
or not the decision was made on the basis of an above
25　threshold correlator output).

　　　FIG. 3 shows the organization of the superbit
synchronizing, tracking, and data extraction logic 46.
Two seven-bit shift registers 270, 272 hold the value and
quality bits, respectively, for the most recent seven
30　subbits.　At each subbit interval, each subbit in the
value shift register 270 is compared (at logic element
276) to the corresponding bit of the correct subbit
pattern for a 1 superbit (i.e., 1110010), generated by
polynomial number generator 274, to generate a "match"
35　bit.　If the value matches the correct pattern, the match

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bit is set to 1, if not it is set to 0. A string of
match bits all equal to 1 (i.e., 1111111) would mean
every subbit corresponded to the pattern for a 1
superbit, whereas a string of match bits all equal to 0
5   (i.e., 0000000) would mean every subbit corresponded to
the pattern for a 0 superbit. Summer 278 adds each match
bit to each quality bit, according to the rules set forth
in the table in the figure. If the match bit is 0, and
the quality is 1 (indicating an above threshold based
10   subbit decision), the summer uses a 0, indicating a high
likelihood that the subbit in question was properly
decoded as part of a 0 superbit. However, if the match
bit is 0, but the quality bit is 0 (indicating no
threshold crossing in the subbit decoding), the summer
15   adds a 1, indicating a lower likelihood that the subbit
was properly decoded as a part of a 0 superbit.
Similarly, the summer produces an output of 3 for a 1
match bit and a 1 quality bit, and an output of 2 for a 1
match bit and a 0 quality bit. All seven summer outputs,
20   for each subbit in the shift register, are accumulated,
resulting in an output from the summer that amounts to a
superbit correlator, and that varies across the range
from 0 to 21. This number is supplied to comparator 280,
which decides whether the superbit is a 1 or a 0
25   according to the rule given in the figure. Quite simply,
if the summer output S is 11 or greater the superbit is
set to 1, and if it is 10 or less the superbit is set to
0.
        The comparator also decides whether the superbit
30   is above or below a superbit threshold, using thresholds
of 5 and 16. A superbit of 1 is above threshold if the
output of the summer is 16 or greater; a superbit of 0 is
above threshold if the output of the summer is 5 or less.
The above-threshold determination is important for
35   synchronizing on superbits, i.e., deciding where the

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superbit boundaries are in the stream of subbits. If the
comparator produces an above threshold output at a
particular subbit, the synchronizing logic 282 assumes
that the seven subbits stored in registers 270, 272
5 correspond to a superbit, and starts a 7-subbit clock.
From that point forward in a data packet, superbit
decoding is done by looking at every seventh output of
the comparator, and decoding based on whether the summer
output S is above or below nominal, i.e., assigning a
10 superbit to 1 if S is 11 or above, and to 0 is S is 10 or
below. During acquisition of the start-of-packet symbol
(the initial packet framing sequence), an above threshold
superbit output occurring at other than every seventh
output will, however, cause an overall resynchronizing at
15 the superbit level (see discussion below).

        The organization of the data packets, as shown in
FIG. 7, is quite similar to that disclosed in copending
application, U.S. Serial No. 07/863,213, filed April 3,
1992, entitled Timing for Spread-Spectrum Communication
20 Across Noisy Media, hereby incorporated by reference.
Each transmitted packet begins with a contention-
resolution preamble 120, followed by a start-of-packet
symbol 122 (a predetermined initial sequence of
superbits, also referred to as an initial EOF), and then
25 by the packet body data 124. Although not shown, each
packet terminates with an end-of-packet symbol (EOP), and
is followed by an error-detection cyclical redundancy
code (CRC).

        Any transceiver may initiate a transmission across
30 the RF medium. Resolution of overlapping contentions for
the medium are resolved using the contention-resolution
preamble 120. Before a transmitter contends for access
to the communication channel, it must observe a quiet
period of a prescribed length. The length is a function
35 of the priority of the message, with high priority

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messages requiring shorter quiet periods than low
priority messages. The length of the required quiet
period is also randomized by having each transmitter add
a further time interval of pseudorandom length to the
5 interval prescribed by the priority of the message. For
CEBus the quiet periods are measured in integral numbers
of 100-μsec USTs ("Unit Symbol Times"), and range in
length from about four to twenty USTs in length. A quiet
period is defined as a matched filter in the receiver not
10 recognizing an above-threshold output (either at the
superbit or subbit level).

After a transmitter has detected the required
quiet period, it begins sending preamble 120, an ASK
(amplitude shift keying) sequence of modified superbits
15 130 having a pitch of 114 μsec. The modified superbits
are shown in FIG. 8. The same seven subbits 132 are
present, each approximately 14.3 μsec in length, but a
2.0 μsec quiet period 134 is inserted between each
subbit. This lengthens the modified superbit to 114
20 μsec.

The preamble consists of from eight to sixteen
114-μsec periods, in which a modified superbit is either
present (superior state) or absent (inferior state). The
length of the preamble is variable because it includes
25 eight pulse-width-modulated symbols. A logical one is
represented by a single 114-μsec interval followed by a
change of state (e.g., a superior state followed by an
inferior state or vice versa). A logical zero is
represented by two 114-μsec intervals in which the state
30 remains the same followed by a change of state (e.g., two
superior states followed by an inferior state or vice
versa). In the example shown in FIG. 7, the eight
symbols are 01110111, and thus the preamble is ten 114-
μsec periods long, with the two zero symbols using up two
35 114-μsec periods, and the six one symbols using only a

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single 114-μsec period apiece. As there are normally
eight symbols in the CEBus preamble, there are 256
different possible preambles. The actual pattern used by
any transmitter varies, and is assigned pseudorandomly
5   (either by the transceiver or the host device). During
the inferior states, when the transceiver is not sending
a chirp, the transceiver listens for activity on the
medium (i.e., a an above-threshold subbit correlator
output). If there is activity, the transmitter quits
10 immediately, without sending the next modified superbit
in the sequence. If there is not activity, the next
modified superbit is sent. By the time the preamble
sequence has been completed, a transmitter still sending
is able to conclude, with some probability of error, that
15 any other contending transmitters have withdrawn from
contention.
        If the transmitter survives the contention-
resolution process, it immediately sends a start-of-
packet symbol 122, which consist of a series of superbits
20 of the same phase (e.g., 1) followed by a superbit of the
opposite phase (e.g., 0). These are the superbits
described earlier. Each has a duration of 100 μsec, and
consists of seven subbits, each of which has a duration
of 100/7 μsec, or about 14.3 μsec. They are transmitted
25 without any quiet space between them, and thus at a pitch
of 100 μsec. Immediately following the start-of-packet
symbol is the data 24 of the packet. Data is transmitted
using the same superbits as in the start-of-packet
symbol.
30   As described in more detail in the above
referenced copending application, the receiver does not
participate in the contention resolution process. It
simply looks for the start-of-packet symbol on the
communication medium, and resynchs when it detects a chip
35 sequence inconsistent with the start-of-packet symbol.

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The receiver has no awareness of whether a contention-
resolution phase is ongoing or not, and reacts
identically to any sequence it detects, whether it be
from the preamble or the start-of-packet symbol.

5   At the subbit level, the receiver expects to
receive a new subbit, either a reverse or a forward
sequence, inside of a 3-clock window every 14.3 μsec, as
described above. Prior to acquisition of the start-of-
packet symbol, the synchronizing, tracking, and data
10 extraction logic continuously examines the correlator
outputs F, R for an above-threshold output. If such an
output is detected outside of a 7-clock wide blackout
window centered at the expected correlator peak (i.e.,
spaced 60 4.2 MHz clock intervals apart), the subbit and
15 superbit tracking are resynchronized. Thus, when a
receiver detects modified superbits, which contain
subbits at a 16.3 μsec pitch (because of the 2 μsec quiet
periods interposed between the subbits) resynchronizing
occurs repeatedly. Thus, ordinarily all of the receivers
20 listening to the channel will be constantly
resynchronizing during the contention resolution phase,
but the resynchronizing will ordinarily stop once the
surviving transmitter has switched to the start-of-packet
symbol. But the synchronizing logic will call for
25 resynchronizing if, for some reason, prior to acquisition
of the start-of-packet symbol, an above threshold
correlator output occurs outside of the 7-clock wide
blackout window spaced at 14.3 μsec pitch.

          Resynchronizing also occurs at the superbit level,
30 prior to acquisition of the start-of-packet symbol, if an
above threshold superbit output is detected by comparator
280 at other than the expected every-seven-subbit time
interval. This superbit resynchronization occurs at the
superbit level only, however, and thus all that happens
35 is that the superbit data is cleared and superbit

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alignment is moved to the point of the resynch. The
value and quality bits generated by the subbit logic 48
are not disturbed.

The rules used for recognition of the start-of-
5   packet symbol are as follows: (1) four successive
superbits must have been detected without
resynchronization at either the subbit or superbit level;
(2) the last four superbits must have the pattern 1110 or
0001; (3) at least one confirming, above-threshold
10 superbit must have been detected at the expected 100-μsec
pitch following the correlation that caused the last
resynchronization.

Thus, although the start-of-packet symbol is eight
100-μsec superbits in length, the receiver need not
15 detect the entire pattern of superbits to conclude that
the symbol has been received. What must happen is that
in the period following a resynchronization there be a
confirming correlation at the expected 100-μsec spacing
as well as a detected pattern matching the end portion of
20 the start-of-packet symbol (i.e., 1110 or 0001, meaning
three superbits of one polarity followed by a single
superbit of the other polarity). Thus, there must be,
altogether, two above threshold correlations, the one
that initiated the resynchronization and the confirming
25 correlation.   These rules for recognizing a start-of-
packet symbol are designed to increase the processing
gain of that portion of the transmission so that it
roughly matches the processing gain of the body of the
packet.
30   After the start-of-packet symbol has been
acquired, the resynchronization functions are shut down,
and data is acquired irrespective of whether an above
threshold output occurs at the wrong place. Data
decisions are made solely on whether the output of the
35 superbit comparator 280 is above or below nominal, i.e.,

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the superbit is assigned to a 1 if the output is 11 or
greater, and to a 0 if the output is 10 or less. If
errors do occur, they are typically detected by
examination of the CRC following acquisition of the data
5 packet, and the application would ordinarily instruct
that the packet be resent.

The subbit tracking and data extraction logic 44
does, however, continue to make sure during acquisition
of the body of the packet that "carrier" is present, i.e,
10 that subbits are being received. This could be done in a
variety of ways, but the preferred embodiment uses the
simple rule that an above threshold subbit must be
acquired every 32 subbits. Thus, if the logic goes more
than 32 subbits without an above threshold output of the
15 correlator, the receiver assumes loss of carrier, so
informs the application, and goes back to a mode of
looking for a start-of-packet symbol.

Other embodiments are within the scope of the
following claims. For example, in some applications,
20 forward and reverse swept chirps (such as disclosed in
copending U.S. Application No. 07/863,213, filed April 3,
1992, entitled Timing for Spread-Spectrum Communication
Across Noisy Media) could be used instead of the direct
sequence technique of the preferred embodiment. The
25 spread spectrum communication techniques are useful over
other media than RF channels; e.g., they could be used
over power lines, where the direct sequences (which,
themselves, have a passband spectrum) could be
transmitted directly without modulation of a carrier.

30   What is claimed is:

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Claims

1. A method of transmitting data on a
communication channel, comprising the steps of:
    receiving a stream of digital signals for
5 transmission on said communication channel, said digital
signals having at least two possible values;
    generating at least two different direct sequences
each containing a plurality of chips, said chips being
selected to approximate a swept frequency waveform in
10 which successive square waves are formed by said chips,
in which the durations of the square waves extend over a
plurality of said chips, and in which the frequency of
said square waves varies across at least a portion of
said sequence;
15   encoding said digital signals by using different
ones of said direct sequences to represent different
values of said digital signals;
    using said direct sequences to generate a direct
sequence spread spectrum signal; and
20   transmitting said spread spectrum signal across
said channel.

    2. Apparatus for transmitting data on a
communication channel, comprising:
    means for receiving a stream of digital signals
25 for transmission on said communication channel, said
digital signals having at least two possible values;
    means for generating at least two different direct
sequences each containing a plurality of chips, said
chips being selected to approximate a swept frequency
30 waveform in which successive square waves are formed by
said chips, in which the duration of each square wave
extends over a plurality of said chips, and in which the
frequency of said square waves varies progressively
across at least a portion of said sequence;

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means for encoding said digital signals by using different ones of said direct sequences to represent different values of said digital signals;

means for using said direct sequences to generate

5   a direct sequence spread spectrum signal; and

means for transmitting said spread spectrum signal across said channel.

3.   The subject matter of claim 1 or 2 wherein said direct sequences are used to modulate a carrier

10 signal to generate said direct sequence spread spectrum signal.

4.   The subject matter of claim 1 or 2 wherein said direct sequences are transmitted directly without modulation of a carrier signal.

15   5.   The subject matter of claim 1 or 2 wherein said direct sequences differ in the direction of sweep of said approximate swept waveform.

6.   The subject matter of claim 5 wherein said digital signal is a sequence of two-level bits, and

20 wherein there are two different direct sequences, one with a forward sweep and another with a reverse sweep, and wherein said forward sweep represents one said level of said bits, and said reverse sweep represents the other said level.

25   7.   The subject matter of claim 1 or 2 wherein said direct sequence has a spectrum, and wherein the energy in said spectrum is primarily in a passband.

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8. The subject matter of claim 3 wherein said
modulated carrier signal has a spectrum, and wherein the
energy in said spectrum is primarily in one or more
passbands.

5   9. The subject matter of claim 8 wherein said
modulated carrier signal transmitted on said channel has
two passbands approximately equidistant above and below
said carrier signal, with either passband containing
sufficient information from which to decode said signal.

10   10. The subject matter of claim 3 wherein the
carrier signal is biphase modulated with said direct
sequences.

             11. The subject matter of claim 3 wherein said
modulated carrier signal is in the RF frequency range.

15   12. A method of transmitting data on a
communication channel, comprising the steps of:
             receiving a stream of digital signals for
transmission on said communication channel, said digital
signals having at least two possible values;
20   generating at least two different direct sequences
each containing a plurality of chips;
             encoding said digital signals by using different
ones of said direct sequences to represent different
values of said digital signals;
25   modulating a carrier signal with said direct
sequences thereby generating a direct sequence spread
spectrum signal, said signal having a spectrum; and
             transmitting said spread spectrum signal across
said channel,

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wherein said direct sequences have a non-
pseudorandom pattern of chips selected so that the energy
in said spectrum is primarily in a passband.

13.   Apparatus for transmitting data on a
5 communication channel, comprising:
          means for receiving a stream of digital signals
for transmission on said communication channel, said
digital signals having at least two possible values;
          means for generating at least two different direct
10 sequences each containing a plurality of chips;
          means for encoding said digital signals by using
different ones of said direct sequences to represent
different values of said digital signals;
          means for modulating a carrier signal with said
15 direct sequences thereby generating a direct sequence
spread spectrum signal, said signal having a spectrum;
and
          means for transmitting said spread spectrum signal
across said channel,
20   wherein said direct sequences have a non-
pseudorandom pattern of chips selected so that the energy
in said spectrum is primarily in a passband.

14.   The subject matter of claim 12 or 13 wherein
the direct sequences each contain a plurality of chips,
25 said chips being selected to approximate a swept
frequency waveform in which successive square waves are
formed by said chips, in which the duration of each
square wave extends over a plurality of said chips, and
in which the frequency of said square waves varies
30 progressively across at least a portion of said sequence.

15.   A method of transmitting data on a
communication channel, comprising the steps of:

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            receiving a stream of digital signals for
transmission on said communication channel, said digital
signals having at least two different values;
            generating first and second spread spectrum
5 signals, said first spread spectrum signal having the
form of a frequency-swept spread spectrum signal in which
frequency is swept in a first direction over at least a
portion of said signal, and said second spread spectrum
signal having the form of a frequency-swept spread
10 spectrum signal in which frequency is swept in a second
direction opposite the first direction over at least a
portion of said signal;
            encoding said digital signals by representing one
said value of said digital signal using said first spread
15 spectrum signal, and representing another said value of
said digital signal using said second spread spectrum
signal; and
            transmitting said spread spectrum signals across
said channel.

20   16. Apparatus for transmitting data on a
communication channel, comprising:
            means for receiving a stream of digital signals
for transmission on said communication channel, said
digital signals having at least two different values;
25   means for generating first and second spread
spectrum signals, said first spread spectrum signal
having the form of a frequency-swept spread spectrum
signal in which frequency is swept in a first direction
over at least a portion of said signal, and said second
30 spread spectrum signal having the form of a frequency-
swept spread spectrum signal in which frequency is swept
in a second direction opposite the first direction over
at least a portion of said signal;

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means for encoding said digital signals by
representing one said value of said digital signal using
said first spread spectrum signal, and representing
another said value of said digital signal using said
5   second spread spectrum signal; and
        means for transmitting said spread spectrum
signals across said channel.

17.   The subject matter of claim 15 or 16 wherein
said first and second spread spectrum signals are used to
10   modulate a carrier signal.

18.   The subject matter of claim 15 or 16 wherein
said first and second spread spectrum signals are
transmitted directly without modulating a carrier signal.

19.   The subject matter of claim 15 or 16 wherein
15   said frequency swept spread spectrum signals are
frequency swept chirps.

20.   The subject matter of claim 17 wherein
        said first and second spread spectrum signals are
generated by generating first and second direct sequences
20   each containing a plurality of chips, said chips being
selected to approximate a swept frequency waveform in
which successive square waves are formed by said chips,
in which the duration of each square wave extends over a
plurality of said chips, and in which the frequency of
25   said square waves varies progressively across at least a
portion of said sequence;
        said digital signals are encoded by using said
first and second direct sequences to represent different
values of said digital signals; and
30   said carrier signal is modulated with said first
and second direct sequences to generate a direct sequence

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spread spectrum signal in which information is
represented by the direction of frequency sweep in the
spread spectrum signal.

21. The subject matter of claim 20 wherein
5 modulation of said carrier signal with said direct
sequence comprises biphase modulation of said carrier
signal.

22. The subject matter of claim 17 wherein the
spectral energy of said spread spectrum signals is
10 primarily in a passband.

23. The subject matter of claim 22 wherein said
carrier signal is further modulated with a phase reversal
sequence selected to increase the spreading of energy
across said passband.

15   24. The subject matter of claim 15 or 16, further
comprising a receiver with matched filters for detecting
said forward and reverse swept signals, and with
independent tracking for accommodating differences in the
times at which said matched filters indicate receipt of
20 the signals.

25. A method of receiving and decoding data that
has been encoded and transmitted using forward and
reverse swept frequency spread spectrum signals, so that
the received signal consists of a succession of spread
25 spectrum signals in which information is encoded as the
difference in direction of sweep, so that different
digital values are represented by different patterns of
forward and reverse swept frequency signals, said method
comprising the steps of:

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        receiving said stream of frequency swept spread
spectrum signals;

        passing said signals through a correlator having
forward and reverse outputs, the forward output
5 representing the correlation of a received signal to the
forward swept spread spectrum signal, and the reverse
output representing the correlation of a received signal
to the reverse swept spread spectrum signal;

        tracking the forward correlator output so that the
10 forward correlator output is sampled at times when a
received signal would be received if it was a forward
swept signal;

        tracking the reverse correlator output
independently of the tracking of the forward correlator
15 output so that the reverse correlator output is sampled
at times when a received signal would be received if it
was a reverse swept signal; and

        making decisions as to whether a received signal
is a forward or reverse swept signal based on samples of
20 the forward and reverse correlator outputs that are taken
generally at different times.

        26. Apparatus for receiving and decoding data
that has been encoded and transmitted using forward and
reverse swept frequency spread spectrum signals, so that
25 the received signal consists a succession of spread
spectrum signals in which information is encoded as the
difference in direction of sweep, so that different
digital values are represented by different patterns of
forward and reverse swept frequency signals, said
30 apparatus comprising:

        means for receiving said stream of frequency swept
spread spectrum signals;

        means for passing said signals through a
correlator having forward and reverse outputs, the

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forward output representing the correlation of a received
signal to the forward swept spread spectrum signal, and
the reverse output representing the correlation of a
received signal to the reverse swept spread spectrum
5   signal;
         means for tracking the forward correlator output
so that the forward correlator output is sampled at times
when a received signal would be received if it was a
forward swept signal;
10   means for tracking the reverse correlator output
independently of the tracking of the forward correlator
output so that the reverse correlator output is sampled
at times when a received signal would be received if it
was a reverse swept signal; and
15   means for making decisions as to whether a
received signal is a forward or reverse swept signal
based on samples of the forward and reverse correlator
outputs that are taken generally at different times.

         27. The subject matter of claim 25 or 26 wherein
20   the receiver includes a local oscillator for generating a
signal that is mixed with the received signal to
demodulate the received signal, and wherein said local
oscillator is not synchronized to the local oscillator
used for modulating the signal at the transmitter, and
25   wherein said local oscillator is therefore generally
operating at a different frequency and phase from that of
the local oscillator used for modulating.

         28. A method of transmitting data on a
communication channel, comprising the steps of:
30   receiving a stream of digital signals for
transmission on said communication channel, said digital
signals having at least two possible values;

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generating at least two different direct sequences each containing a plurality of chips, the difference between said direct sequences being recognizable independently of the relative phase between the two
5   sequences;

encoding said digital signals by using different ones of said direct sequences to represent different values of said digital signals;

pseudorandomly varying the overall phase of said
10 direct sequences;

using said direct sequences to generate a direct sequence spread spectrum signal, said signal having a spectrum; and

transmitting said spread spectrum signal across
15 said channel.

29.   Apparatus for transmitting data on a communication channel, comprising:

means for receiving a stream of digital signals for transmission on said communication channel, said
20 digital signals having at least two possible values;

means for generating at least two different direct sequences each containing a plurality of chips, the difference between said direct sequences being recognizable independently of the relative phase between
25 the two sequences;

means for encoding said digital signals by using different ones of said direct sequences to represent different values of said digital signals;

means for pseudorandomly varying the overall phase
30 of said direct sequences;

means for using said direct sequences to generate a direct sequence spread spectrum signal, said signal having a spectrum; and

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means for transmitting said spread spectrum signal across said channel.

30. The subject matter of claim 28 or 29 wherein said direct sequences are used to modulate a carrier to
5 generate said direct sequence spread spectrum signal.

31. The subject matter of claim 28 or 29 wherein said pseudorandom variation of phase is done so that each successive direct sequence has its phase pseudorandomly varied with respect to prior sequences.

10   32. The subject matter of claim 31 in which the double level coding is used, in which subbits are represented by one or more of said direct sequences, and said subbits are arranged in at least two different subbit patterns, and in which the pseudorandom variation
15 of phase is at the subbit level so that the phase of the direct sequences varies pseudorandomly from subbit to subbit.

33. The subject matter of claim 32 wherein the pseudorandom variation repeats after a time period that
20 is greater than the number of subbits in said subbit patterns, and that is not an integral multiple of the length of said subbit patterns.

34. The subject matter of claim 28 or 29 wherein the pseudorandom variation of phase reduces the peak
25 power in the spectrum of the transmitted signal by comparison to the peak power in the spectrum of the transmitted signal without the pseudorandom variation of phase.

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35.   A method of receiving and decoding data that
has been encoded and transmitted using dual level spread
spectrum modulation in which information is conveyed as
superbits, and in which superbits are encoded as unique
5   sequences of a plurality of subbits, each represented by
a unique spread spectrum signal, said method comprising
the steps of:
            receiving a stream of spread spectrum signals;
            passing said signals through a correlator
10 configured to distinguish between said unique spread
spectrum signals representing said subbits;
            sampling the output of the correlator, and
determining therefrom the values of a succession of
subbits;
15   determining a quality indicator from the strength
of the sampled correlator output used in determining the
value of a subbit;
            combining the value and quality indicators for a
succession of subbits in decoding the superbits.

20   36.   Apparatus for receiving and decoding data
that has been encoded and transmitted using dual level
spread spectrum modulation in which information is
conveyed as superbits, and in which superbits are encoded
as unique sequences of a plurality of subbits, each
25 represented by a unique spread spectrum signal, said
apparatus comprising:
            means for receiving a stream of spread spectrum
signals;
            means for passing said signals through a
30 correlator configured to distinguish between said unique
spread spectrum signals representing said subbits;
            means for sampling the output of the correlator,
and determining therefrom the values of a succession of
subbits;

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means for determining a quality indicator from the
strength of the sampled correlator output used in
determining the value of a subbit;
means for combining the value and quality
5 indicators for a succession of subbits in decoding the
superbits.

37.   The subject matter of claim 35 or 36 wherein
each superbit is represented by a fixed number of
subbits, and wherein a superbit correlation output is
10 determined at each subbit from the value and quality
indicators for a segment of subbits of the same length as
said fixed number of subbits making up a superbit.

38.   The subject matter of claim 37 wherein the
superbit correlation is determined by accumulating a
15 total of a match indicator and the quality indicator for
each subbit in said segment of subbits, wherein the match
indicator is determined by comparing said unique
sequences of subbits to the actual subbit values in said
segment.

20   39.   The subject matter of claim 38 wherein the
superbit correlation output has a range of values
representative of the likelihood that the received
pattern of spread spectrum signals is one of the unique
sequences of subbits representing a superbit, wherein
25 superbit decoding is performed by initiating a superbit
tracker when the superbit correlation output exceeds a
threshold, and by sampling the superbit correlation
output at intervals spaced by the length of a superbit
following detection of the above-threshold correlation
30 output.

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40. The subject matter of claim 12, 13, 15, 16, 25, 26, 28, 29, 35, or 36 wherein said spread-spectrum signals are transmitted at RF frequencies.

41. The subject matter of claim 25 or 26 wherein
5 the forward and reverse swept signals are generated using forward and reverse swept direct sequences.

42. A method of transmitting data on a communication channel, comprising the steps of:
        receiving a stream of digital signals for
10 transmission on said communication channel, said digital signals having at least two possible values;
        generating a direct sequence containing a plurality of chips;
        encoding said digital signals by using said direct
15 sequence and a reverse ordered sequence, identical to said direct sequence except having the reverse order, to represent different values of said digital signals; and
        transmitting said direct and reverse ordered sequences across said channel.

20   43. Apparatus for transmitting data on a communication channel, comprising:
        means for receiving a stream of digital signals for transmission on said communication channel, said digital signals having at least two possible values;
25   means for generating a direct sequence containing a plurality of chips;
        means for encoding said digital signals by using said direct sequence and a reverse ordered sequence, identical to said direct sequence except having the
30 reverse order, to represent different values of said digital signals; and

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means for transmitting said direct and reverse
ordered sequences across said channel.

44.   The subject matter of claim 42 or 43 wherein
said direct and reverse ordered sequences are used to
5   modulate a carrier signal, and the modulated carrier
signal is transmitted across said channel.

45.   The subject matter of claim 42 or 43 wherein
said direct and reverse ordered sequences approximate
swept frequency waveforms, in which successive square
10   waves are formed by said chips, and in which the
durations of the square waves extend over a plurality of
said chips, and in which the frequency of the square
waves varies approximately progressively across at least
a portion of said sequence, from a low to a higher
15   frequency in a forward swept sequence, and from a high to
a lower frequency in a reverse swept sequence.

46.   The subject matter of claim 42 or 43 wherein
said direct sequence and reverse ordered sequence each
have a spectrum, and wherein the energy in said spectrum
20   is primarily in a passband.

47.   The subject matter of claim 42 or 43 wherein
said direct sequence and said reverse-ordered sequence
are transmitted directly without modulation of a carrier.

48.   The subject matter of claim 47 wherein said
25   direct and said reverse-ordered sequence are transmitted
over power lines.

49.   The subject matter of claim 44 wherein said
modulated carrier signal is in the RF frequency range.

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50. A method of receiving and decoding data that has been encoded and transmitted using forward and reverse ordered versions of a direct sequence, so that the transmitted signal comprises a sequence of spread

5 spectrum signals corresponding to one of said forward and reverse ordered versions of said direct sequence, said method comprising the steps of:

receiving said sequence of spread spectrum signals;

10 passing said received signals through a correlator configured to provide forward and reverse correlator outputs, the forward output representing the correlation of a said received signal to the forward directed sequence, and the reverse output representing the

15 correlation of a said received signal to the reverse directed sequence; and

using said forward and reverse correlator outputs to decode said received signals to reconstruct the digital data encoded in the transmitter.

20   51. Apparatus for receiving and decoding data that has been encoded and transmitted using forward and reverse ordered versions of a direct sequence, so that the transmitted signal comprises a sequence of spread spectrum signals corresponding to one of said forward and

25 reverse ordered versions of said direct sequence, said apparatus comprising:

means for receiving said sequence of spread spectrum signals;

a correlator for receiving said sequence of

30 signals and configured to provide forward and reverse correlator outputs, the forward output representing the correlation of a said received signal to the forward directed sequence, and the reverse output representing

- 43 -

the correlation of a said received signal to the reverse
directed sequence; and
          a decoder for using said forward and reverse
correlator outputs to decode said received signals to
5 reconstruct the digital data encoded in the transmitter.

          52. The subject matter of claim 50 or 51 wherein
said received signals are demodulated using a locally
generated carrier signal prior to being applied to the
correlator.

10   53. The subject matter of claim 50 or 51 wherein
          said correlator comprises a forward matched filter
configured to recognize said forward ordered direct
sequence and a reverse matched filter configured to
recognize said reverse ordered direct sequence,
15   wherein said forward matched filter is the mirror
image of said reverse matched filter, and
          wherein said forward and reverse matched filters
share the same shift register.

          54. The subject matter of claim 53 wherein said
20 correlator comprises four matched filter sections, a 0
and 90 degree phase section for each of said forward and
reverse ordered sequences.

          55. The subject matter of claim 54 wherein
approximately half of the taps in said shift register are
25 assigned to 0-degree sections and the remainder of the
taps are assigned to 90-degree sections;
          wherein for each of the taps assigned to 0-degree
sections and for each of the taps assigned to 90-degree
there are formed two additional groups of taps, a common
30 group that have the same value for the forward and
reverse ordered sequences and a difference group that

- 44 -

have opposite values for the forward and reverse ordered
sequences;

wherein each of the four groups of taps is summed
to form four sums, a 0-degree common sum, a 0-degree
5 difference sum, a 90-degree common sum, and a 90-degree
difference sum; and

wherein the outputs of each of said four filter
sections are formed by combining corresponding common and
difference sums.

10   56.   The subject matter of claim 55 wherein said
combining of corresponding common and difference sums
comprises forming the sum of said common and difference
sums for one of said forward and reverse ordered
sequences and forming the difference of said common and
15 difference sums for the other of said sequences.

          57.   The subject matter of claim 15 or 16, further
comprising a receiver with matched filters for detecting
said forward and reverse swept signals, and with
independent tracking for accommodating differences in the
20 times at which said matched filters indicate receipt of
the signals.

1/11

FIG 1

FIG 2

FIG 3

4/11

" 1 " SUBBIT

← 360 CHIPS →
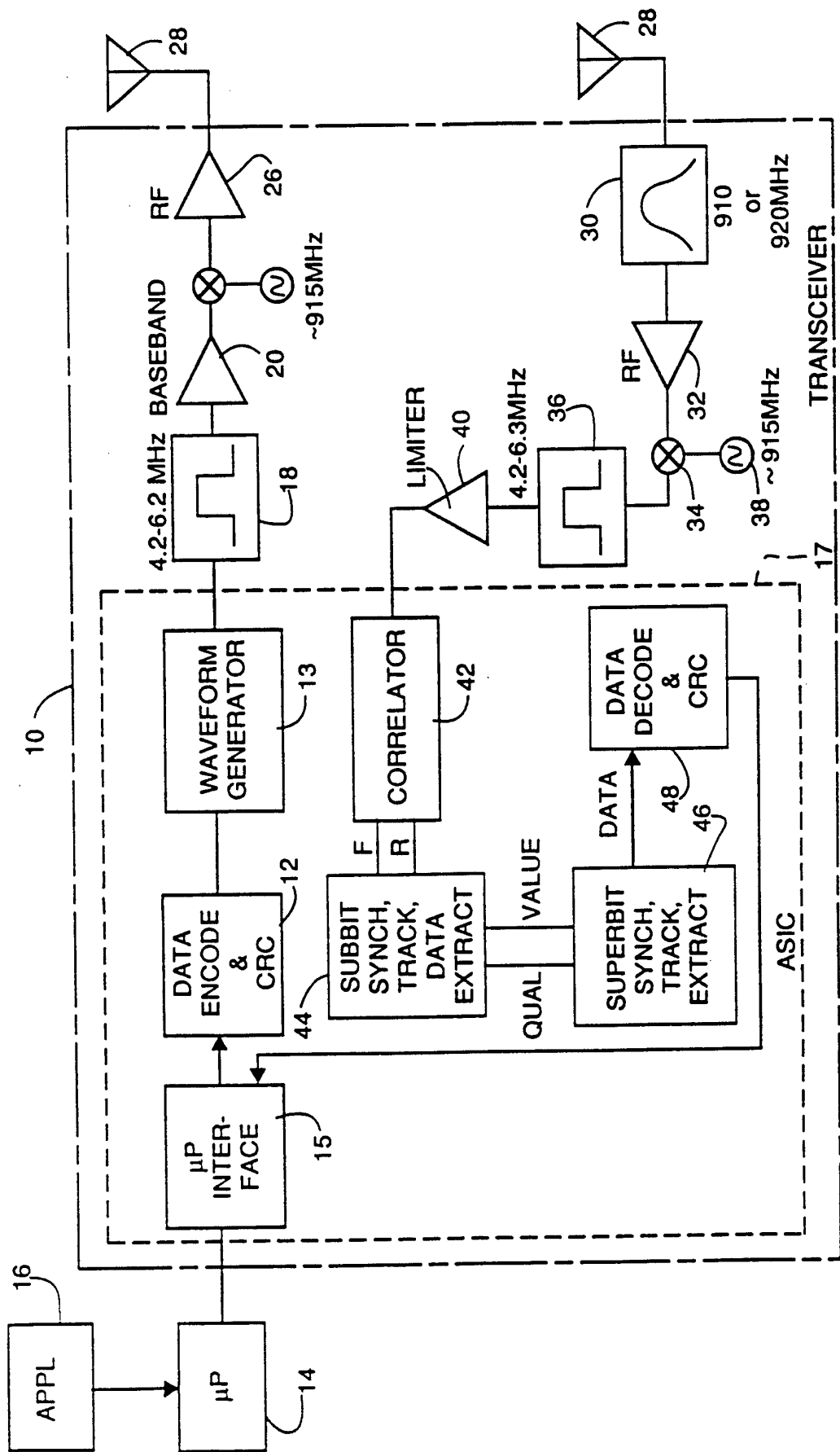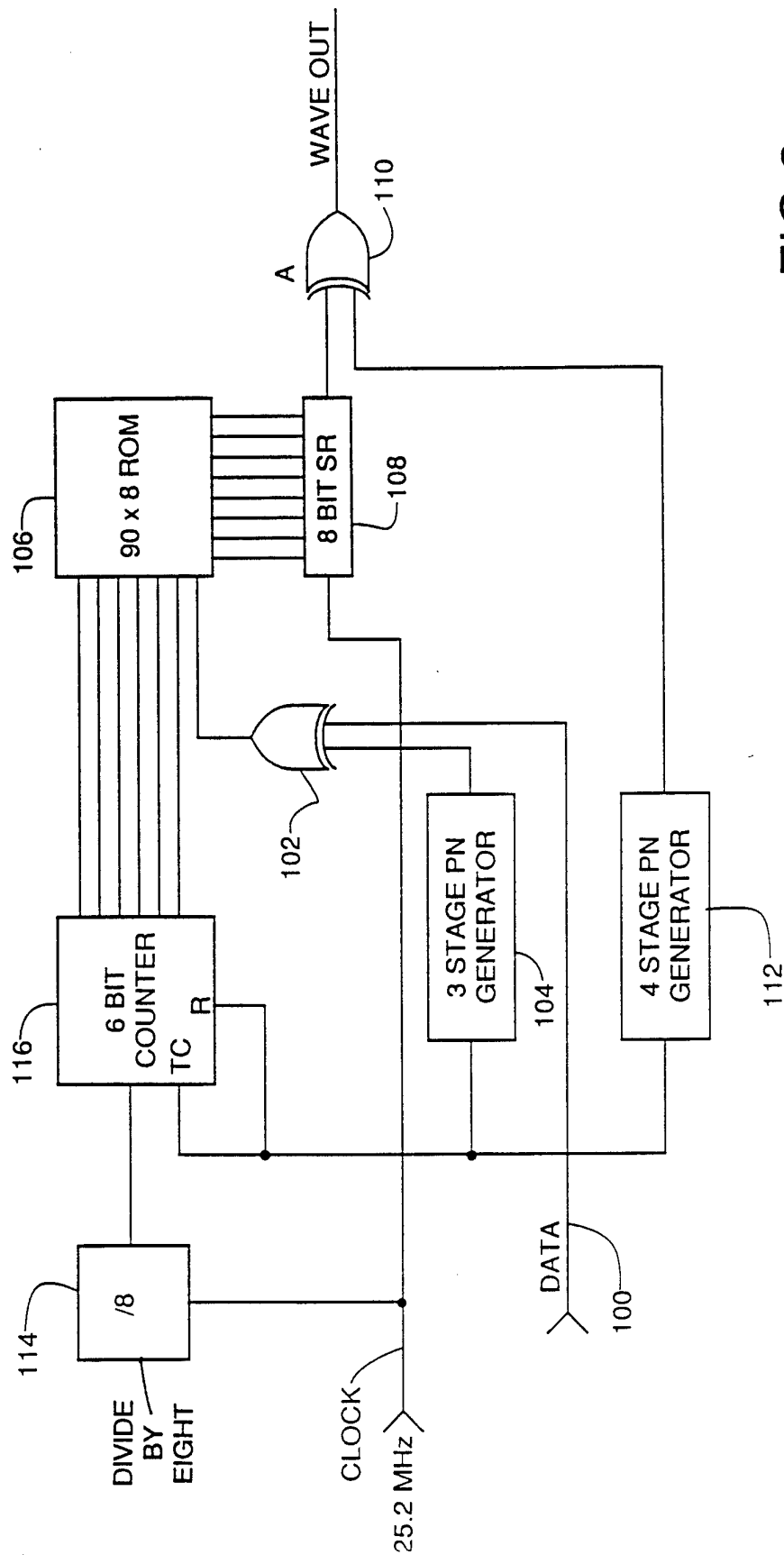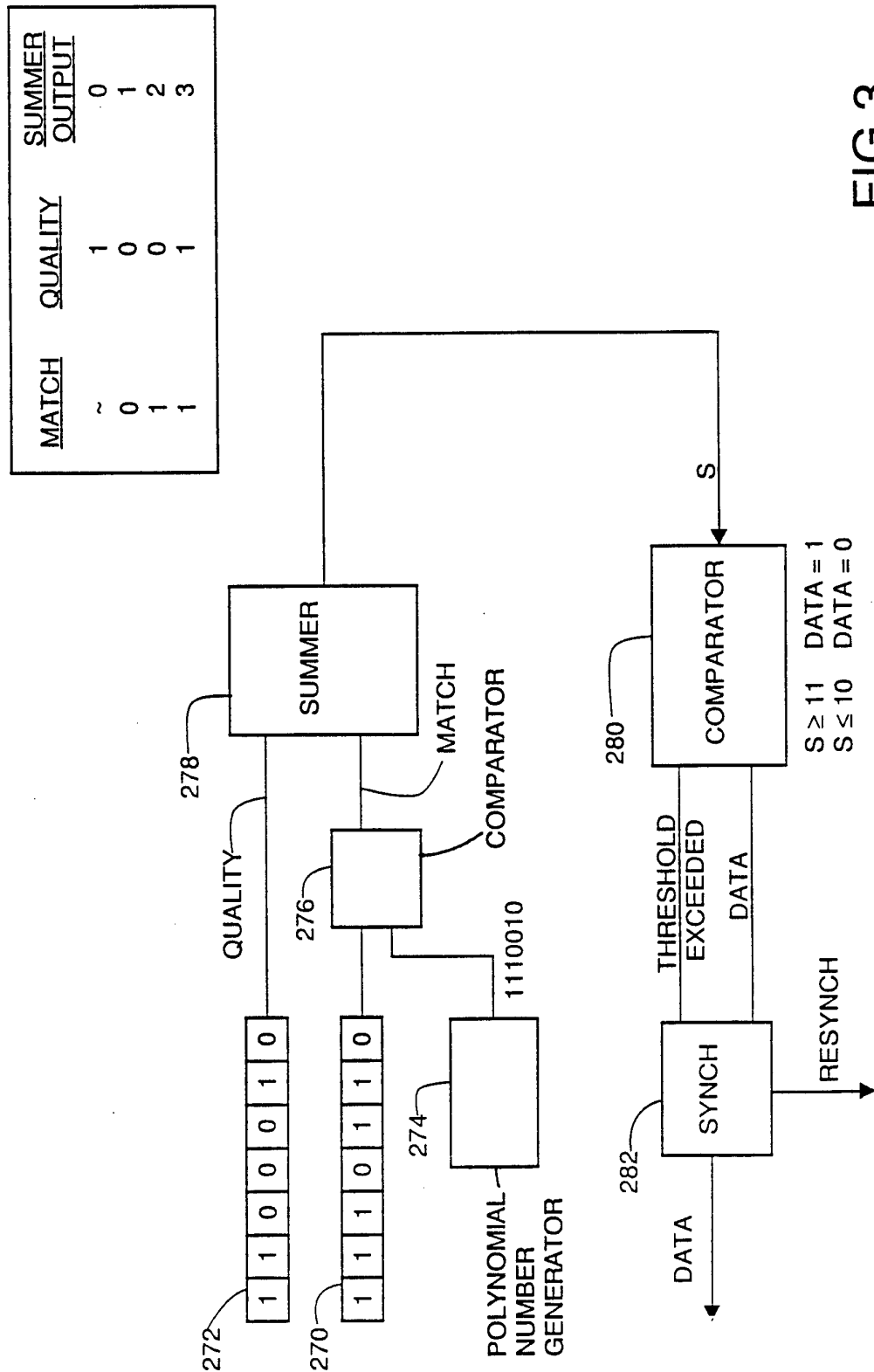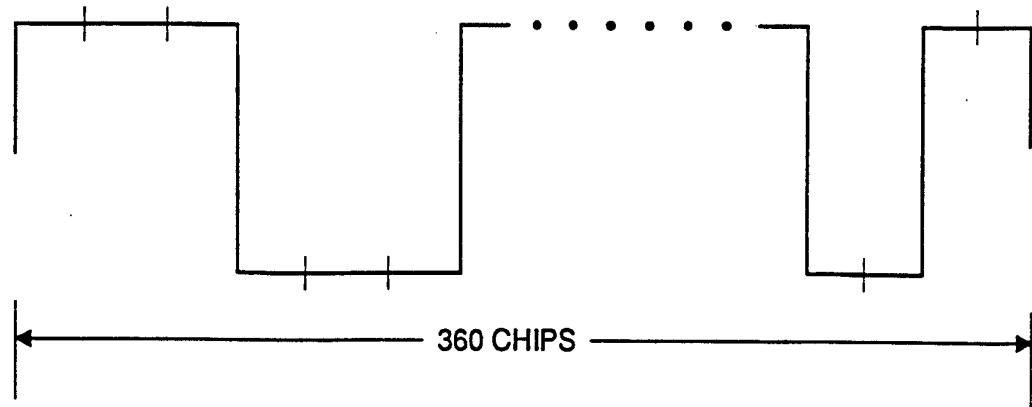
FIG 4A

" 0 " SUBBIT
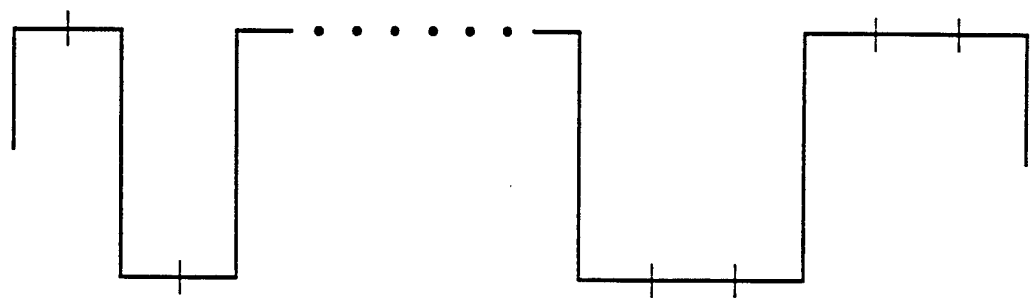
FIG 4B

```
111000111000111000110001110001110001110001100011100011100111
000111000110001110011100011000111001110001100011100111001110
001100011000110001100011000110001100011000110001100111001110
011000110001100111001100011001110011000110011100110011100110
011100110011100110011000110011001110011001100111001100110011
001110011001100110011001100111001100110011001100110011001100
```

FIG 5

" 1 " SUPERBIT

SUBBIT

| 1 | 1 | 1 | 0 | 0 | 1 | 0 |

SWEEP
DIRECTION   F   F   F   R   R   F   R

" 1 " SUBBIT   " 0 " SUBBIT

FIG 6A

" 0 " SUPERBIT

SUBBIT

| 0 | 0 | 0 | 1 | 1 | 0 | 1 |

SWEEP   R   R   R   F   F   R   F
DIRECTION
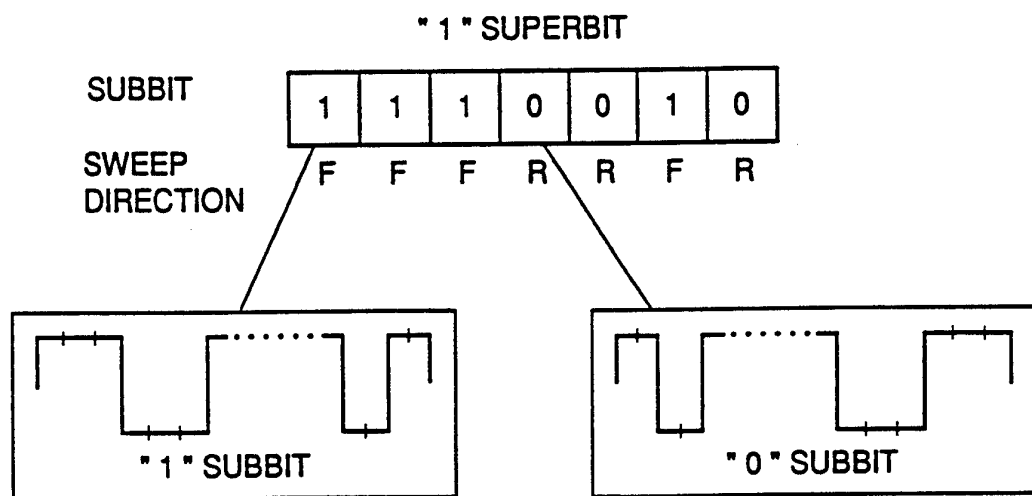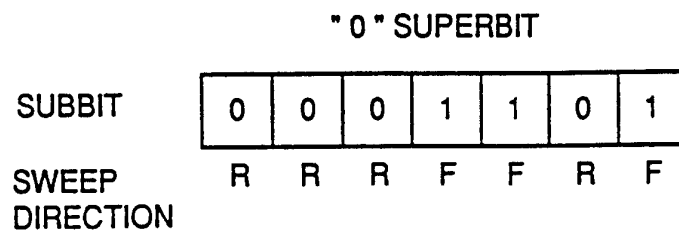
FIG 6B

FIG 7

FIG 8

FIG 9A

4.2 MHz   6.3 MHz

C

208

FIG 9B

4.0 MHz   6.1 MHz

C

210

FIG 9C

6.3 MHz   4.2 MHz

C

212

FIG 9D

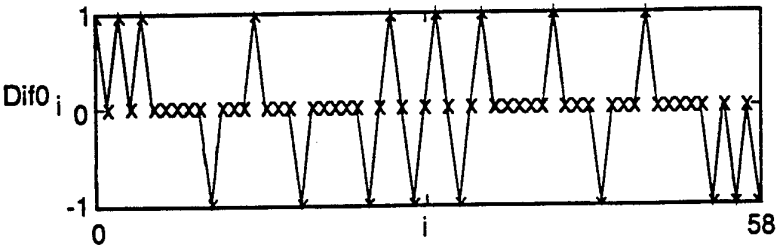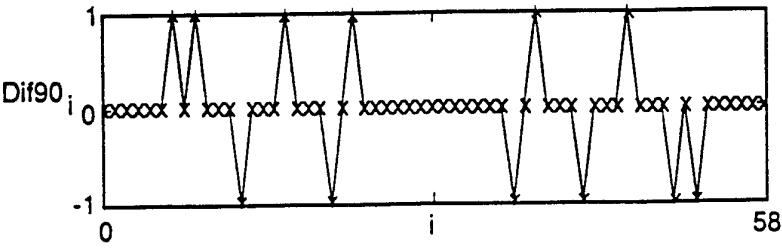6.1 MHz   4.0 MHz

C

214
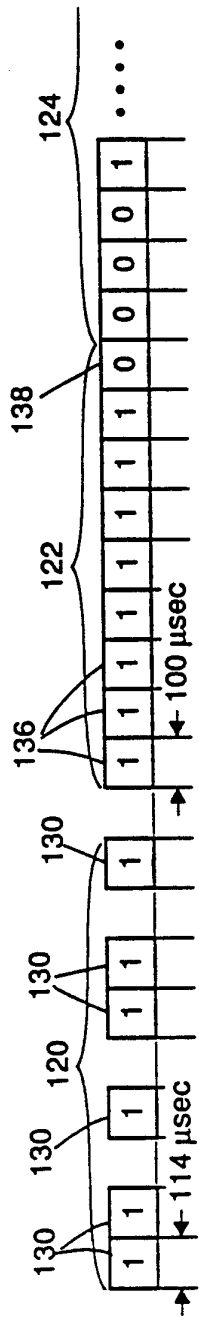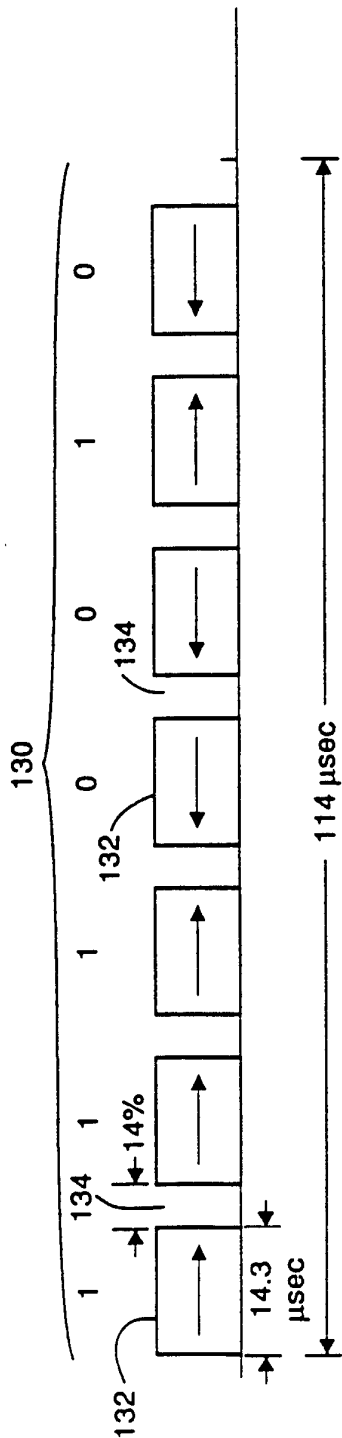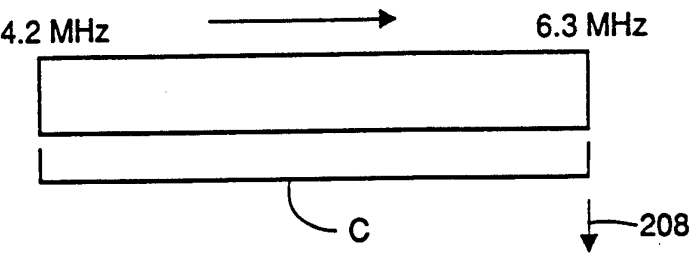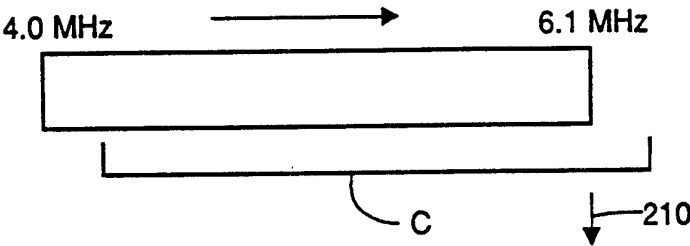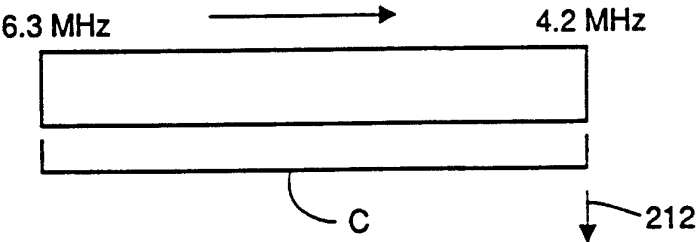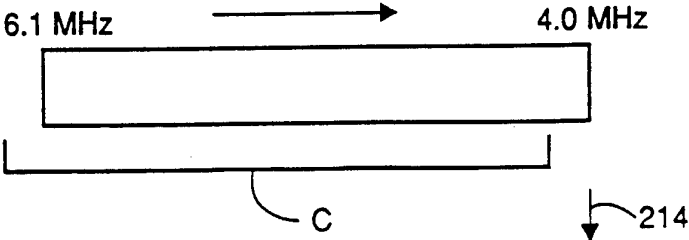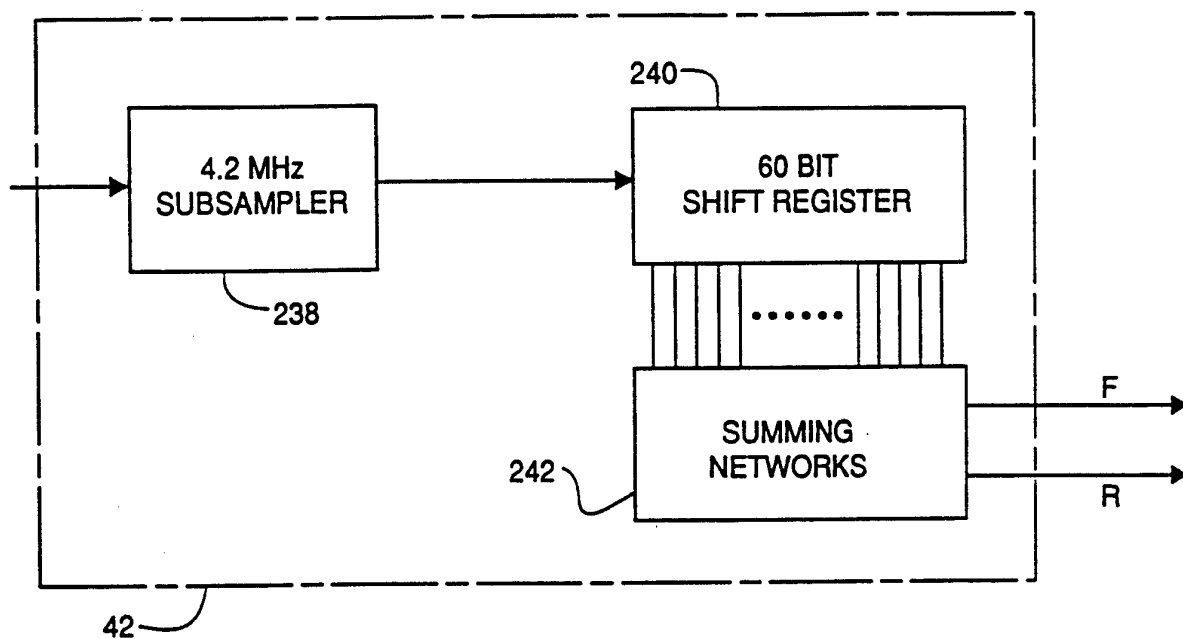
FIG 10
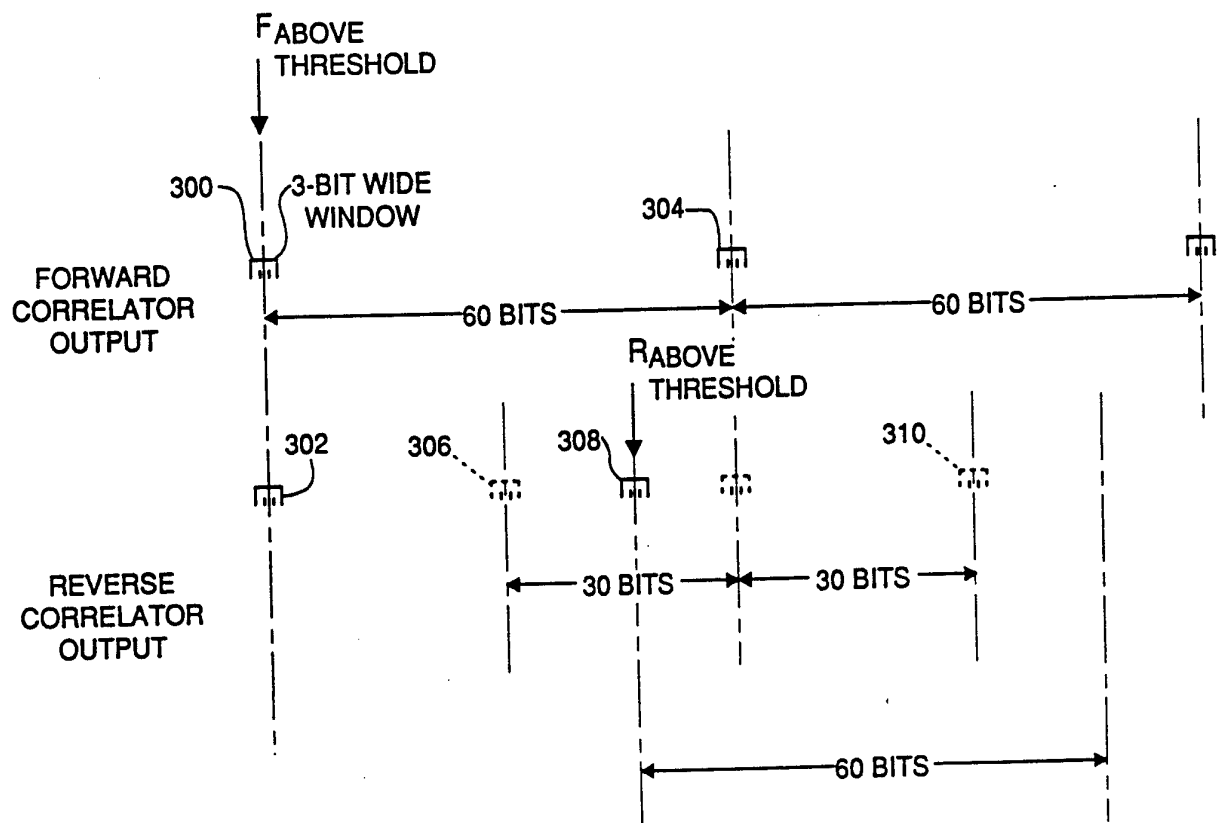
FIG 11
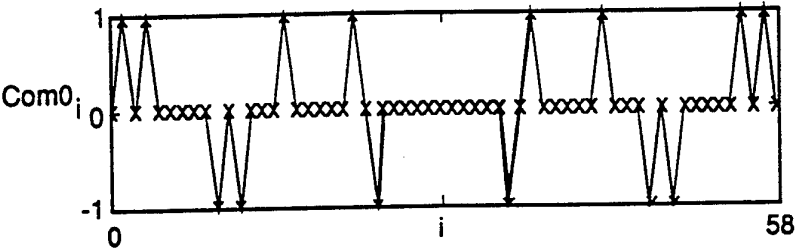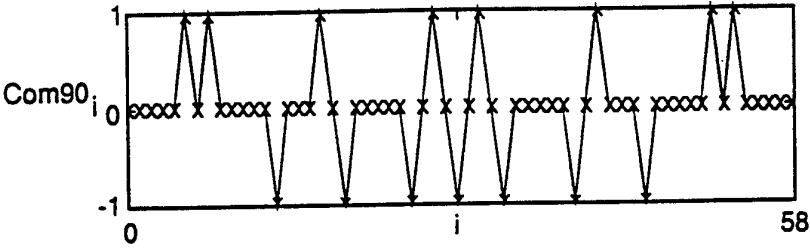
FIG 12A

FIG 12B

FIG 12C

FIG 12D
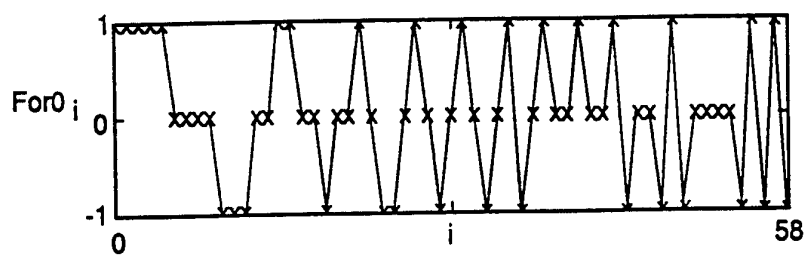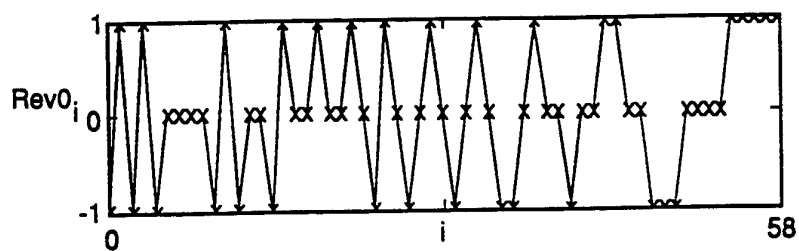
FIG 13A

FIG 13B

FIG 13C
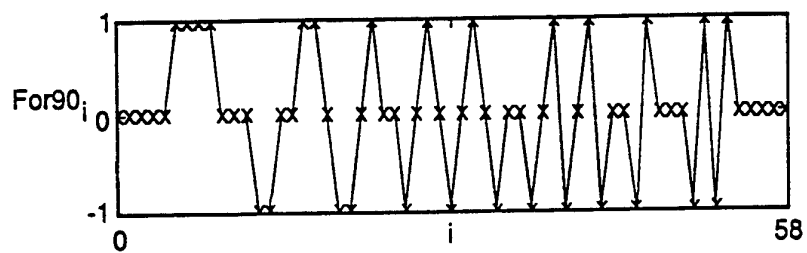
FIG 13D
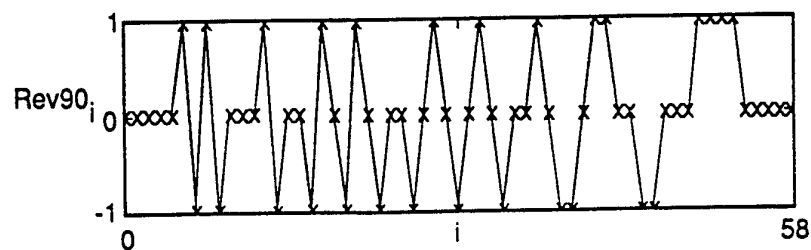
INTERNATIONAL SEARCH REPORT

A.	CLASSIFICATION OF SUBJECT MATTER

IPC(5) : H04L 27/30
US CL   : 375/1

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)

U.S. :   380/34; 370/18.

Documentation searched other than minimum documentation to the extent that such documents are included in the fields searched

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

C. DOCUMENTS CONSIDERED TO BE RELEVANT

Category*	Citation of document, with indication, where appropriate, of the relevant passages	Relevant to claim No.
A	US, A, 3,665,474 (THAYER) 23 MAY 1972	1-57
A	US, A, 4,300,235 (OUTRAM) 10 NOVEMBER 1981	1-57
A	US, A, 4,468,792 (BAKER ET AL) 28 AUGUST 1984	1-57
A	US, A, 4,653,076 (JERRIM ET AL) 24 MARCH 1987	1-57
A	US, A, 4,969,159 (BELCHER ET AL) 06 NOVEMBER 1990	1-57
A	US, A, 5,132,.986 (ENDO ET AL) 21 JULY 1992	1-57

☐ Further documents are listed in the continuation of Box C.   ☐ See patent family annex.

*	Special categories of cited documents:	"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
"A"	document defining the general state of the art which is not considered to be part of particular relevance		
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"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	"&"	document member of the same patent family

Date of the actual completion of the international search	Date of mailing of the international search report
01 September 1993	16 SEP 1993
Name and mailing address of the ISA/US Commissioner of Patents and Trademarks Box PCT Washington, D.C. 20231	Authorized officer BERNARR EARL GREGORY
Facsimile No. NOT APPLICABLE	Telephone No. (703) 308-0479

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