(12) INTERNATIONAL APPLICATION PUBLISHED UNDER THE PATENT COOPERATION TREATY (PCT)

(19) World Intellectual Property Organization
International Bureau

(43) International Publication Date
6 April 2006 (06.04.2006)

PCT

(10) International Publication Number
WO 2006/036331 A2

(51) International Patent Classification:
H04B 1/10 (2006.01)

(21) International Application Number:
PCT/US2005/028799

(22) International Filing Date: 11 August 2005 (11.08.2005)

(25) Filing Language: English

(26) Publication Language: English

(30) Priority Data:
10/948,634 23 September 2004 (23.09.2004) US

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(81) Designated States *(unless otherwise indicated, for every kind of national protection available)*: AE, AG, AL, AM, AT, AU, AZ, BA, BB, BG, BR, BW, BY, BZ, CA, CH, CN, CO, CR, CU, CZ, DE, DK, DM, DZ, EC, EE, EG, ES, FI, GB, GD, GE, GH, GM, HR, HU, ID, IL, IN, IS, JP, KE, KG, KM, KP, KR, KZ, LC, LK, LR, LS, LT, LU, LV, MA, MD, MG, MK, MN, MW, MX, MZ, NA, NG, NI, NO, NZ, OM, PG, PH, PL, PT, RO, RU, SC, SD, SE, SG, SK, SL, SM, SY, TJ, TM, TN, TR, TT, TZ, UA, UG, US, UZ, VC, VN, YU, ZA, ZM, ZW.

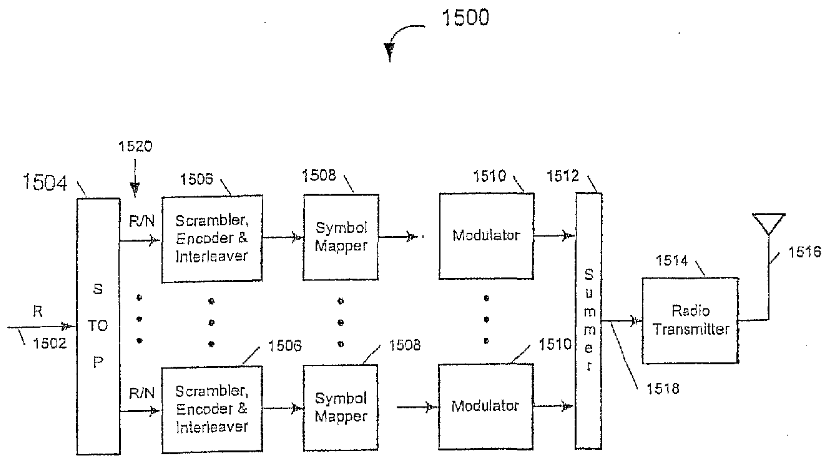
(84) Designated States *(unless otherwise indicated, for every kind of regional protection available)*: ARIPO (BW, GH, GM, KE, LS, MW, MZ, NA, SD, SL, SZ, TZ, UG, ZM, ZW), Eurasian (AM, AZ, BY, KG, KZ, MD, RU, TJ, TM), European (AT, BE, BG, CH, CY, CZ, DE, DK, EE, ES, FI, FR, GB, GR, HU, IE, IS, IT, LT, LU, LV, MC, NL, PL, PT, RO, SE, SI, SK, TR), OAPI (BF, BJ, CF, CG, CI, CM, GA, GN, GQ, GW, ML, MR, NE, SN, TD, TG).

Declaration under Rule 4.17:
— *as to applicant's entitlement to apply for and be granted a patent (Rule 4.17(ii))*

Published:
— *without international search report and to be republished upon receipt of that report*

For two-letter codes and other abbreviations, refer to the "Guidance Notes on Codes and Abbreviations" appearing at the beginning of each regular issue of the PCT Gazette.

(54) Title: SYSTEMS AND METHODS FOR EQUALIZATION OF RECEIVED SIGNALS IN A WIRELESS COMMUNICATION NETWORK



(57) Abstract: A method of communicating over a wideband communication channel is provided. In one embodiment an equalizer comprises a hard decision block which feeds an forward-error-correction (FEC) block and a shift register, which uses past estimates during a first iteration. The output of the FEC block is fed back to the shift register during a subsequent iteration to fully remove any inter-chip-interference, or inter-symbol-interference. This Abstract is provided for the sole purpose of complying with the Abstract requirement rules that allows a reader to quickly ascertain the subject matter of the disclosure contained herein This Abstract is submitted with the explicit that understanding that it will not be used to interpret or to limit the scope or the meaning of the claims.

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SYSTEMS AND METHODS FOR EQUALIZATION OF RECEIVED SIGNALS IN A WIRELESS COMMUNICATION NETWORK

1. Field of the Invention

The invention relates generally to wireless communication and more particularly to systems and methods for wireless communication over a wide bandwidth channel using a plurality of sub-channels.

2. Background

Wireless communication systems are proliferating at the Wide Area Network (WAN), Local Area Network (LAN), and Personal Area Network (PAN) levels. These wireless communication systems use a variety of techniques to allow simultaneous access to multiple users. The most common of these techniques are Frequency Division Multiple Access (FDMA), which assigns specific frequencies to each user, Time Division Multiple Access (TDMA), which assigns particular time slots to each user, and Code Division Multiple Access (CDMA), which assigns specific codes to each user. But these wireless communication systems and various modulation techniques are afflicted by a host of problems that limit the capacity and the quality of service provided to the users. The following paragraphs briefly describe a few of these problems for the purpose of illustration.

One problem that can exist in a wireless communication system is multipath interference. Multipath interference, or multipath, occurs because some of the energy in a transmitted wireless signal bounces off of obstacles, such as buildings or mountains, as it travels from source to destination. The obstacles in effect create reflections of the transmitted signal and the more obstacles there are, the more reflections they generate. The reflections then travel along their own transmission paths to the destination (or receiver). The reflections will contain the same information as the original signal; however, because of the differing transmission path lengths, the reflected signals will be out of phase with the original signal. As a result, they will often combine destructively with the original signal in the receiver. This is referred to as fading. To combat fading, current systems typically try to estimate the multipath effects and then compensate for them in the receiver using an equalizer. In practice, however, it is very difficult to achieve effective multipath compensation.

A second problem that can affect the operation of wireless communication systems is interference from adjacent communication cells within the system. In FDMA/TDMA systems, this type of interference is prevent through a frequency reuse plan. Under a frequency reuse plan, available communication frequencies are allocated to communication cells within the communication system such that the same frequency will not be used in adjacent cells. Essentially, the available frequencies are split into groups. The number of groups is termed the reuse factor. Then the communication cells are grouped into clusters, each cluster containing the same number of cells as there are frequency groups. Each frequency group is then assigned to a cell in each cluster. Thus, if a frequency reuse factor of 7 is used, for example, then a particular communication frequency will be used only once in every seven communication cells. Thus, in any group of seven communication cells, each cell can only use 1/7th of the available frequencies, i.e., each cell is only able to use 1/7th of the available bandwidth.

In a CDMA communication system, each cell uses the same wideband communication channel. In order to avoid interference with adjacent cells, each communication cell uses a particular set of spread spectrum codes to differentiate communications within the cell from those originating outside of the cell. Thus, CDMA systems preserve the bandwidth in the sense that they avoid reuse planning. But as will be discussed, there are other issues that limit the bandwidth in CDMA systems as well. Thus, in overcoming interference, system bandwidth is often sacrificed. Bandwidth is becoming a very valuable commodity as wireless communication systems continue to expand by adding more and more users. Therefore, trading off bandwidth for system performance is a costly, albeit necessary, proposition that is inherent in all wireless communication systems.

The foregoing are just two examples of the types of problems that can affect conventional wireless communication systems. The examples also illustrate that there are many aspects of wireless communication system performance that can be improved through systems and methods that, for example, reduce interference, increase bandwidth, or both.

Not only are conventional wireless communication systems effected by problems, such as those described in the preceding paragraphs, but also different types of systems are effected in different ways and to different degrees. Wireless communication systems can be split into three types: 1) line-of-sight systems, which can include point-to-point or point-to-multipoint systems; 2) indoor non-line of sight systems; and 3) outdoor systems such as wireless WANs. Line-of-sight systems are least affected by the problems described above, while indoor systems are more affected, due for example to signals bouncing off of building walls. Outdoor systems are by far the most affected of the three systems. Because these types of problems are limiting factors in the design of wireless transmitters and receivers, such designs must be tailored to the specific types of system in which it will operate. In practice, each type of system implements unique communication standards that address the issues unique to the particular type of system. Even if an indoor system used the same communication protocols and modulation techniques as an outdoor system, for example, the receiver designs would still be different because multipath and other problems are unique to a given type of system and must be addressed with unique solutions. This would not necessarily be the case if cost efficient and effective methodologies can be developed to combat such problems as described above that build in programmability so that a device can be reconfigured for different types of systems and still maintain superior performance.

SUMMARY OF THE INVENTION

In order to combat the above problems, the systems and methods described herein provide a novel channel access technology that provides a cost efficient and effective methodology that builds in programmability so that a device can be reconfigured for different types of systems and still maintain superior performance. In one aspect of the invention, a method of communicating over a wideband-communication channel divided into a plurality of sub-channels is provided. The method comprises dividing a single serial message intended for one of the plurality of communication devices into a plurality of parallel messages, encoding each of the plurality of parallel messages onto at least some of the plurality of sub-channels, and transmitting the encoded plurality of parallel messages to the communication device over the wideband communication channel.

When symbols are restricted to particular range of values, the transmitters and receivers can be simplified to eliminate high power consuming components such as a local oscillator, synthesizer and phase locked loops. Thus, in one aspect a transmitter comprises a plurality of pulse converters and differential amplifiers, to convert a balanced trinary data stream into a pulse sequence which can be filtered to reside in the desired frequency ranges and phase. The use of the balanced trinary data stream allows conventional components to be replaced by less costly, smaller components that consume less power.

Similarly, in another aspect, a receiver comprises detection of the magnitude and phase of the symbols, which can be achieved with an envelope detector and sign detector respectively. Thus, conventional receiver components can be replaced by less costly, smaller components that consume less power.

Other aspects, advantages, and novel features of the invention will become apparent from the following Detailed Description of Preferred Embodiments, when considered in conjunction with the accompanying drawings.

BRIEF DESCRIPTION OF THE DRAWINGS

Preferred embodiments of the present inventions taught herein are illustrated by way of example, and not by way of limitation, in the figures of the accompanying drawings, in which:

Figure 1 is a diagram illustrating an example embodiment of a wideband channel divided into a plurality of sub-channels in accordance with the invention;

Figure 2 is a diagram illustrating the effects of multipath in a wireless communication system;

Figure 3 is a diagram illustrating another example embodiment of a wideband communication channel divided into a plurality of sub-channels in accordance with the invention;

Figure 4 is a diagram illustrating the application of a roll-off factor to the sub-channels of figures 1 and 2;

Figure 5A is a diagram illustrating the assignment of sub-channels for a wideband communication channel in accordance with the invention;

Figure 5B is a diagram illustrating the assignment of time slots for a wideband communication channel in accordance with the invention;

Figure 6 is a diagram illustrating an example embodiment of a wireless communication in accordance with the invention;

Figure 7 is a diagram illustrating the use of synchronization codes in the wireless communication system of figure 5 in accordance with the invention;

Figure 8 is a diagram illustrating a correlator that can be used to correlate synchronization codes in the wireless communication system of figure 5;

Figure 9 is a diagram illustrating synchronization code correlation in accordance with the invention;

Figure 10 is a diagram illustrating the cross-correlation properties of synchronization codes configured in accordance with the invention;

Figure 11 is a diagram illustrating another example embodiment of a wireless communication system in accordance with the invention;

Figure 12A is a diagram illustrating how sub-channels of a wideband communication channel according to the present invention can be grouped in accordance with the present invention;

Figure 12B is a diagram illustrating the assignment of the groups of sub- channels of figure 12A in accordance with the invention;

Figure 13 is a diagram illustrating the group assignments of figure 12B in the time domain;

Figure 14 is a flow chart illustrating the assignment of sub-channels based on SIR measurements in the wireless communication system of figure 11 in accordance with the invention;

Figure 15 is a logical block diagram of an example embodiment of transmitter configured in accordance with the invention;

Figure 16 is a logical block diagram of an example embodiment of a modulator configured in accordance with the present invention for use in the transmitter of figure 15;

Figure 17 is a diagram illustrating an example embodiment of a rate controller configured in accordance with the invention for use in the modulator of figure 16;

Figure 18 is a diagram illustrating another example embodiment of a rate controller configured in accordance with the invention for use in the modulator of figure 16;

Figure 19 is a diagram illustrating an example embodiment of a frequency encoder configured in accordance with the invention for use in the modulator of figure 16;

Figure 20 is a logical block diagram of an example embodiment of a TDM/FDM block configured in accordance with the invention for use in the modulator of figure 16;

Figure 21 is a logical block diagram of another example embodiment of a TDM/FDM block configured in accordance with the invention for use in the modulator of figure 16;

Figure 22 is a logical block diagram of an example embodiment of a frequency shifter configured in accordance with the invention for use in the modulator of figure 16;

Figure 23 is a logical block diagram of a receiver configured in accordance with the invention;

Figure 24 is a logical block diagram of an example embodiment of a demodulator configured in accordance with the invention for use in the receiver of figure 23;

Figure 25 is a logical block diagram of an example embodiment of an equalizer configured in accordance with the present invention for use in the demodulator of figure 24;

Figure 26 is a logical block diagram of an example embodiment of a wireless communication device configured in accordance with the invention;

Figure 27 is a flow chart illustrating an exemplary method for recovering bandwidth in a wireless communication network in accordance with the invention;

Figure 28 is a diagram illustrating an exemplary wireless communication network in which the method of figure 27 can be implemented;

Figure 29 is a logical block diagram illustrating an exemplary transmitter that can be used in the network of figure 28 to implement the method of figure 27;

Figure 30 is a logical block diagram illustrating another exemplary transmitter that can be used in the network of figure 28 to implement the method of figure 27;

Figure 31 is a diagram illustrating another exemplary wireless communication network in which the method of figure 27 can be implemented;

Figure 32 is a diagram illustrating an exemplary receive chain that illustrated various stages of a receiver;

Figure 33 is a diagram illustrating an embodiment of an equalizer;

Figure 34 is a diagram illustrating an example configuration;

Figure 35 is a diagram illustrating an example implementation of DFE;

Figure 36 is a diagram illustrating an example embodiment of a matched filter 3600 that can be used, for example, in the receive chain of figure 32; and

Figure 37 is a diagram illustrating an example embodiment of a DFE that can be used to implement an iterative equalization.

DETAILED DESCRIPTION OF THE PREFERRED EMBODIMENTS

1. Introduction

In order to improve wireless communication system performance and allow a single device to move from one type of system to another, while still maintaining superior performance, the systems and methods described herein provide various communication methodologies that enhance performance of transmitters and receivers with regard to various common problems that afflict such systems and that allow the transmitters and/or receivers to be reconfigured for optimal performance in a variety of systems. Accordingly, the systems and methods described herein define a channel access protocol that uses a common wideband communication channel for all communication cells. The wideband channel, however, is then divided into a plurality of sub-channels. Different sub-channels are then assigned to one or more users within each cell. But the base station, or service access point, within each cell transmits one message that occupies the entire bandwidth of the wideband channel. Each user's communication device receives the entire message, but only decodes those portions of the message that reside in sub-channels assigned to the user. For a point-to-point system, for example, a single user may be assigned all sub-channels and, therefore, has the full wide band channel available to them. In a wireless WAN, on the other hand, the sub-channels may be divided among a plurality of users.

In the descriptions of example embodiments that follow, implementation differences, or unique concerns, relating to different types of systems will be pointed out to the extent possible. But it should be understood that the systems and methods described herein are applicable to any type of communication systems. In addition, terms such as communication cell, base station, service access point, etc. are used interchangeably to refer to the common aspects of networks at these different levels. To begin illustrating the advantages of the systems and methods described herein, one can start by looking at the multipath effects for a single wideband communication channel 100 of bandwidth *B* as shown in figure 1. Communications sent over channel 100 in a traditional wireless communication system will comprise digital data bits, or symbols, that are encoded and modulated onto a RF carrier that is centered at frequency f_c and occupies bandwidth *B*. Generally, the width of the symbols (or the symbol duration) *T* is defined as *1/B*. Thus, if the bandwidth *B* is equal to 100MHz, then the symbol duration *T* is defined by the following equation:

$$T = 1/B = 1/100\ megahertz\ (MHZ) = 10\ nanoseconds\ (ns). \quad (1)$$

When a receiver receives the communication, demodulates it, and then decodes it, it will recreate a stream 104 of data symbols 106 as illustrated in figure 2. But the receiver will also receive multipath versions 108 of the same data stream. Because multipath data streams 108 are delayed in time relative to the data stream 104 by delays d1, d2, d3, and d4, for example, they may combine destructively with data stream 104.

A delay spread d_s is defined as the delay from reception of data stream 104 to the reception of the last multipath data stream 108 that interferes with the reception of data stream 104. Thus, in the example illustrated in figure 2, the delay spread d_s is equal to delay d4. The delay spread d_s will vary for different environments. An environment with a lot of obstacles will create a lot of multipath reflections. Thus, the delay spread d_s will be longer. Experiments have shown that for outdoor WAN type environments, the delay spread d_s can be as long as 20 microseconds. Using the 10ns symbol duration of equation (1), this translates to 2000 symbols. Thus, with a very large bandwidth, such as 100MHz, multipath interference can cause a significant amount of interference at the symbol level for which adequate compensation is difficult to achieve. This is true even for indoor environments. For indoor LAN type systems, the delay spread d_s is significantly shorter, typically about 1 microsecond. For a 10ns symbol duration, this is equivalent to 100 symbols, which is more manageable but still significant. By segmenting the bandwidth *B* into a plurality of sub-channels 202, as illustrated in figure 2, and generating a distinct data stream for each sub-channel, the multipath effect can be reduced to a much more manageable level. For example, if the bandwidth *b* of each sub-channel 202 is 500KHz, then the symbol duration is 2 microseconds. Thus, the delay spread d_s for each sub-channel is equivalent to only 10 symbols (outdoor) or half a symbol (indoor). Thus, by breaking up a message that occupies the entire bandwidth *B* into discrete messages, each occupying the bandwidth *b* of sub-channels 202, a very wideband signal that suffers from relatively minor multipath effects is created.

Before discussing further features and advantages of using a wideband communication channel segmented into a plurality of sub-channels as described, certain aspects of the sub-channels will be explained in more detail. Referring back to figure 3, the overall bandwidth *B* is segmented into *N* sub-channels center at frequencies f_0 *to* f_{N-1}. Thus, the sub-channel 202 that is immediately to the right of *fc* is offset from *fc* by *b/2*, where *b* is the bandwidth of each sub-channel 202. The next sub-channel 202 is offset by *3b/2*, the next by *5b/2*, and so on. To the left of *fc*, each sub-channel 202 is offset by $-b/2$, $-3b/2$, $-5b/2$, etc.

Preferably, sub-channels 202 are non-overlapping as this allows each sub-channel to be processed independently in the receiver. To accomplish this, a roll-off factor is preferably applied to the signals in each sub-channel in a pulse-shaping step. The effect of such a pulse-shaping step is illustrated in figure 3 by the non-rectangular shape of the pulses in each sub-channel 202. Thus, the bandwidth b of each sub-channel can be represented by an equation such as the following:

$$b=(1+r)/T; \qquad (2)$$

Where r = the roll-off factor; and T = the symbol duration.

Without the roll-off factor, i.e., $b = 1/T$, the pulse shape would be rectangular in the frequency domain, which corresponds to a $(\sin x)/x$ function in the time domain. The time domain signal for a $(\sin x)/x$ signal 400 is shown in figure 4 in order to illustrate the problems associated with a rectangular pulse shape and the need to use a roll-off factor.

As can be seen, main lobe 402 comprises almost all of signal 400. But some of the signal also resides in side lobes 404, which stretch out indefinitely in both directions from main lobe 402. Side lobes 404 make processing signal 400 much more difficult, which increases the complexity of the receiver. Applying a roll-off factor r, as in equation (2), causes signal 400 to decay faster, reducing the number of side lobes 404. Thus, increasing the roll-off factor decreases the length of signal 400, i.e., signal 400 becomes shorter in time. But including the roll-off factor also decreases the available bandwidth in each sub-channel 202. Therefore, r must be selected so as to reduce the number of side lobes 404 to a sufficient number, e.g., 15, while still maximizing the available bandwidth in each sub-channel 202.

Thus, the overall bandwidth B for communication channel 200 is given by the following equation:

$$B=N(1+r)/T; \qquad (3)$$

or

$$B=M/T; \qquad (4)$$

Where

$$M=(1+r)N. \qquad (5)$$

For efficiency purposes related to transmitter design, it is preferable that r is chosen so that M in equation (5) is an integer. Choosing r so that M is an integer allows for more efficient transmitters designs using, for example, Inverse Fast Fourier Transform (IFFT) techniques. Since $M = N + N(r)$, and N is always an integer, this means that r must be chosen so that $N(r)$ is an integer. Generally, it is preferable for r to be between 0.1 and 0.5. Therefore, if N is 16, for example, then 0.5 could be selected for r so that $N(r)$ is an integer. Alternatively, if a value for r is chosen in the above example so that $N(r)$ is not an integer, B can be made slightly wider than M/T to compensate. In this case, it is still preferable that r be chosen so that $N(r)$ is approximately an integer.

2. Example Embodiment of a Wireless Communication System

With the above in mind, figure 6 illustrates an example communication system 600 comprising a plurality of cells 602 that each use a common wideband communication channel to communicate with communication devices 604 within each cell 602. The common communication channel is a wideband communication channel as described above. Each communication cell 602 is defined as the coverage area of a base station, or service access point, 606 within the cell. One such base station 606 is shown for illustration in figure 6. For purposes of this specification and the claims that follow, the term base station will be used generically to refer to a device that provides wireless access to the wireless communication system for a plurality of communication devices, whether the system is a line of sight, indoor, or outdoor system.

Because each cell 602 uses the same communication channel, signals in one cell 602 must be distinguishable from signals in adjacent cells 602. To differentiate signals from one cell 602 to another, adjacent base stations 606 use different synchronization codes according to a code reuse plan. In figure 6, system 600 uses a synchronization code reuse factor of 4, although the reuse factor can vary depending on the application.

Preferably, the synchronization code is periodically inserted into a communication from a base station 606 to a communication device 604 as illustrated in figure 7. After a predetermined number of data packets 702, in this case two, the particular synchronization code 704 is inserted into the information being transmitted by each base station 606. A synchronization code is a sequence of data bits known to both the base station 606 and any communication devices 604 with which it is communicating. The synchronization code allows such a communication device 604 to synchronize its timing to that of base station 606, which, in turn, allows device 604 to decode the data properly. Thus, in cell 1 (see lightly shaded cells 602 in figure 6), for example, synchronization code 1 (SYNC1) is inserted into data stream 706, which is generated by base station 606 in cell 1, after every two packets 702; in cell 2 SYNC2 is inserted after every two packets 702; in cell 3 SYNC3 is inserted; and in cell 4 SYNC4 is inserted. Use of the synchronization codes is discussed in more detail below.

In figure 5A, an example wideband communication channel 500 for use in communication system 600 is divided into 16 sub-channels 502, centered at frequencies f_0 to f_{15}. A base station 606 at the center of each communication cell 602 transmits a single packet occupying the whole bandwidth B of wideband channel 500. Such a packet is illustrated by packet 504 in figure 5B. Packet 504 comprises sub-packets 506 that are encoded with a frequency offset corresponding to one of sub-channels 502. Sub-packets 506 in effect define available time slots in packet 504. Similarly, sub-channels 502 can be said to define available frequency bins in communication channel 500. Therefore, the resources available in communication cell 602 are time slots 506 and frequency bins 502, which can be assigned to different communication devices 604 within each cell 602. Thus, for example, frequency bins 502 and time slots 506 can be assigned to 4 different communication devices 604 within a cell 602 as shown in figure 5. Each communication device 604 receives the entire packet 504, but only processes those frequency bins 502 and/or timeslots 506 that are assigned to it. Preferably, each device 604 is assigned non-adjacent frequency bins 502, as in figure 5B. This way, if interference corrupts the information in a portion of communication channel 500, then the effects are spread across all devices 604 within a cell 602. Hopefully, by spreading out the effects of interference in this manner the effects are minimized and the entire information sent to each device 604 can still be recreated from the unaffected information received in other frequency bins. For example, if interference, such as fading, corrupted the information in bins f_0-f_4, then each user 1-4 loses one packet of data. But each user potentially receives three unaffected packets from the other bins assigned to them. Hopefully, the unaffected data in the other three bins provides enough information to recreate the entire message for each user. Thus, frequency diversity can be achieved by assigning non-adjacent bins to each of multiple users.

Ensuring that the bins assigned to one user are separated by more than the coherence bandwidth ensures frequency diversity. As discussed above, the coherence bandwidth is approximately equal to $1/d_s$. For outdoor systems, where *ds* is typically 1 microsecond, $1/d_s$ = 1/1 microsecond = 1 Mega Hertz (MHz). Thus, the non-adjacent frequency bands assigned to a user are preferably separated by at least 1MHz. It is even more preferable, however, if the coherence bandwidth plus some guard band to ensure sufficient frequency diversity separate the non-adjacent bins assigned to each user. For example, it is preferable in certain implementations to ensure that at least 5 times the coherence bandwidth, or 5MHz in the above example, separates the non-adjacent bins.

Another way to provide frequency diversity is to repeat blocks of data in frequency bins assigned to a particular user that are separated by more than the coherence bandwidth. In other words, if 4 sub-channels 202 are assigned to a user, then data block *a* can be repeated in the first and third sub-channels 202 and data block *b* can be repeated in the second and fourth sub-channels 202, provided the sub-channels are sufficiently separated in frequency. In this case, the system can be said to be using a diversity length factor of 2. The system can similarly be configured to implement other diversity lengths, e.g., 3, 4,..., *l*.

It should be noted that spatial diversity can also be included depending on the embodiment. Spatial diversity can comprise transmit spatial diversity, receive spatial diversity, or both. In transmit spatial diversity, the transmitter uses a plurality of separate transmitters and a plurality of separate antennas to transmit each message. In other words, each transmitter transmits the same message in parallel. The messages are then received from the transmitters and combined in the receiver. Because the parallel transmissions travel different paths, if one is affected by fading, the others will likely not be affected. Thus, when they are combined in the receiver, the message should be recoverable even if one or more of the other transmission paths experienced severe fading.

Receive spatial diversity uses a plurality of separate receivers and a plurality of separate antennas to receive a single message. If an adequate distance separates the antennas, then the transmission path for the signals received by the antennas will be different. Again, this difference in the transmission paths will provide imperviousness to fading when the signals from the receivers are combined. Transmit and receive spatial diversity can also be combined within a system such as system 600 so that two antennas are used to transmit and two antennas are used to receive. Thus, each base station 606 transmitter can include two antennas, for transmit spatial diversity, and each communication device 604 receiver can include two antennas, for receive spatial diversity. If only transmit spatial diversity is implemented in system 600, then it can be implemented in base stations 606 or in communication devices 604. Similarly, if only receive spatial diversity is included in system 600, then it can be implemented in base stations 606 or communication devices 604.

The number of communication devices 604 assigned frequency bins 502 and/or time slots 506 in each cell 602 is preferably programmable in real time. In other words, the resource allocation within a communication cell 602 is preferably programmable in the face of varying external conditions, i.e., multipath or adjacent cell interference, and varying requirements, i.e., bandwidth requirements for various users within the cell. Thus, if user 1 requires the whole bandwidth to download a large video file, for example, then the allocation of bins 502 can be adjust to provide user 1 with more, or even all, of bins 502. Once user 1 no longer requires such large amounts of bandwidth, the allocation of bins 502 can be readjusted among all of users 1-4.

It should also be noted that all of the bins assigned to a particular user can be used for both the forward and reverse link. Alternatively, some bins 502 can be assigned as the forward link and some can be assigned for use on the reverse link, depending on the implementation.

To increase capacity, the entire bandwidth B is preferably reused in each communication cell 602, with each cell 602 being differentiated by a unique synchronization code (see discussion below). Thus, system 600 provides increased immunity to multipath and fading as well as increased bandwidth due to the elimination of frequency reuse requirements.

3. Synchronization

Figure 8 illustrates an example embodiment of a synchronization code correlator 800. When a device 604 in cell 1 (see figure 6), for example, receives an incoming communication from the cell 1 base station 606, it compares the incoming data with SYNCl in correlator 800. Essentially, the device scans the incoming data trying to correlate the data with the known synchronization code, in this case SYNCl. Once correlator 800 matches the incoming data to SYNCl it generates a correlation peak 804 at the output. Multipath versions of the data will also generate correlation peaks 806, although these peaks 806 are generally smaller than correlation peak 804. The device can then use the correlation peaks to perform channel estimation, which allows the device to adjust for the multipath using an equalizer. Thus, in cell 1, if correlator 800 receives a data stream comprising SYNC1, it will generate correlation peaks 804 and 806. If, on the other hand, the data stream comprises SYNC2, for example, then no peaks will be generated and the device will essentially ignore the incoming communication.

Even though a data stream that comprises SYNC2 will not create any correlation peaks, it can create noise in correlator 800 that can prevent detection of correlation peaks 804 and 806. Several steps can be taken to prevent this from occurring. One way to minimize the noise created in correlator 800 by signals from adjacent cells 602, is to configure system 600 so that each base station 606 transmits at the same time. This way, the synchronization codes can preferably be generated in such a manner that only the synchronization codes 704 of adjacent cell data streams, e.g., streams 708, 710, and 712, as opposed to packets 702 within those streams, will interfere with detection of the correct synchronization code 704, e.g., SYNC1. The synchronization codes can then be further configured to eliminate or reduce the interference.

For example, the noise or interference caused by an incorrect synchronization code is a function of the cross correlation of that synchronization code with respect to the correct code. The better the cross correlation between the two, the lower the noise level. When the cross correlation is ideal, then the noise level will be virtually zero as illustrated in figure 9 by noise level 902. Therefore, a preferred embodiment of system 600 uses synchronization codes that exhibit ideal cross correlation, i.e., zero. Preferably, the ideal cross correlation of the synchronization codes covers a period *l* that is sufficient to allow accurate detection of multipath 906 as well as multipath correlation peaks 904. This is important so that accurate channel estimation and equalization can take place. Outside of period *l*, the noise level 908 goes up, because the data in packets 702 is random and will exhibit low cross correlation with the synchronization code, e.g., SYNC1. Preferably, period *l* is actually slightly longer then the multipath length in order to ensure that the multipath can be detected.

a. <u>Synchronization code generation</u>

Conventional systems use orthogonal codes to achieve cross correlation in correlator 800. In system 600 for example, SYNC1, SYNC2, SYNC3, and SYNC4, corresponding to cells 1-4 (see lightly shaded cells 602 of figure 6) respectively, will all need to be generated in such a manner that they will have ideal cross correlation with each other. In one embodiment, if the data streams involved comprise high and low data bits, then the value "1" can be assigned to the high data bits and "-1" to the low data bits. Orthogonal data sequences are then those that produce a "0" output when they are exclusively ORed (XORed) together in correlator 800. The following example illustrates this point for orthogonal sequences 1 and 2:

sequence 1: 11-1 1
sequence 2: <u>11 1-1</u>
1 1-1-1 =0

Thus, when the results of XORing each bit pair are added, the result is "0".

But in system 600, for example, each code must have ideal, or zero, cross correlation with each of the other codes used in adjacent cells 602. Therefore, in one example embodiment of a method for generating synchronization codes exhibiting the properties described above, the process begins by selecting a "perfect sequence" to be used as the basis for the codes. A perfect sequence is one that when correlated with itself produces a number equal to the number of bits in the sequence. For example:

Perfect sequence 1: 11-11
<u>11-11</u>
1111=4

But each time a perfect sequence is cyclically shifted by one bit, the new sequence is orthogonal with the original sequence. Thus, for example, if perfect sequence 1 is cyclically shifted by one bit and then correlated with the original, the correlation produces a "0" as in the following example;

Perfect sequence 1: 11-1 1
<u>11 1-1</u>
11 -1-1 =0

If the perfect sequence 1 is again cyclically shifted by one bit, and again correlated with the original, then it will produce a "0". In general, you can cyclically shift a perfect sequence by any number of bits up to its length and correlate the shifted sequence with the original to obtain a "0".

Once a perfect sequence of the correct length is selected, the first synchronization code is preferably generated in one embodiment by repeating the sequence 4 times. Thus, if perfect sequence 1 is being used, then a first synchronization code y would be the following:

y=1 1-1 1 11-11 11-11 11-11.

Or in generic form:

y=x(0)x(l)x(2)x(3) x(0)x(l)x(2)x(3) x(0)x(l)x(2)x(3) x(0)x(l)x(2)x(3).

For a sequence of length L:

y=x(0)x(l)...x(L)x(0)x(l)...x(L)x(0)x(l)...x(L)x(0)x(l)...x(L)

Repeating the perfect sequence allows correlator 800 a better opportunity to detect the synchronization code and allows generation of other uncorrelated frequencies as well. Repeating has the effect of sampling in the frequency domain. This effect is illustrated by the graphs in figure 10. Thus, in TRACE 1, which corresponds to synchronization code y, a sample 1002 is generated every fourth sample bin 1000. Each sample bin is separated by $1/(4LxT)$, where T is the symbol duration. Thus, in the above example, where $L = 4$, each sample bin is separated by $1/(16xT)$ in the frequency domain. TRACES 2-4 illustrate the next three synchronization codes. As can be seen, the samples for each subsequent synchronization code are shifted by one sample bin relative to the samples for the previous sequence. Therefore, none of the sequences interfere with each other.

To generate the subsequent sequences, corresponding to TRACES 2-4, sequence y must be shifted in frequency. This can be accomplished using the following equation:

$$z^r(m) = y(m)*exp(j*2*\pi*r*m/(n*L)), \qquad (5)$$

for $r = 1$ to L (# of sequences) and m = 0 to $4*L$-1 (time); and

where: $z^r(m)$ = each subsequent sequence;

$y(m)$ = the first sequence; and

n = the number of times the sequence is repeated.

It will be understood that multiplying by an $exp(j2\pi(r*m/N))$ factor, where N is equal to the number of times the sequence is repeated n multiplied by the length of the underlying perfect sequence L, in the time domain results in a shift in the frequency domain. Equation (5) results in the desired shift as illustrated in figure 9 for each of synchronization codes 2-4, relative to synchronization code 1. The final step in generating each synchronization code is to append the copies of the last M samples, where M is the length of the multipath, to the front of each code. This is done to make the convolution with the multipath cyclic and to allow easier detection of the multipath.

It should be noted that synchronization codes can be generated from more than one perfect sequence using the same methodology. For example, a perfect sequence can be generated and repeated four times and then a second perfect sequence can be generated and repeated four times to get a n factor equal to eight. The resulting sequence can then be shifted as described above to create the synchronization codes.

b. Signal Measurements Using Synchronization Codes

Therefore, when a communication device is at the edge of a cell, it will receive signals from multiple base stations and, therefore, will be decoding several synchronization codes at the same time. This can be illustrated with the help of figure 11, which illustrates another example embodiment of a wireless communication system 1100 comprising communication cells 1102, 1104, and 1106 as well as communication device 1108, which is in communication with base station 1110 of cell 1102 but also receiving communication from base stations 1112 and 1114 of cells 1104 and 1106, respectively. If communications from base station 1110 comprise synchronization code SYNC1 and communications from base station 1112 and 1114 comprise SYNC2 and SYNC3 respectively, then device 1108 will effectively receive the sum of these three synchronization codes. This is because, as explained above, base stations 1110, 1112, and 1114 are configured to transmit at the same time. Also, the synchronization codes arrive at device 1108 at almost the same time because they are generated in accordance with the description above.

Again as described above, the synchronization codes SYNC1, SYNC2, and SYNC3 exhibit ideal cross correlation. Therefore, when device 1108 correlates the sum x of codes SYNC1, SYNC2, and SYNC3, the latter two will not interfere with proper detection of SYNC1 by device 1108. Importantly, the sum x can also be used to determine important signal characteristics, because the sum x is equal to the sum of the synchronization code signal in accordance with the following equation:

$$x = SYNC1 + SYNC2 + SYNC3. \quad (6)$$

Therefore, when SYNC1 is removed, the sum of SYNC2 and SYNC3 is left, as shown in the following:

$$x - SYNC1 = SYNC2 + SYNC3. \quad (7)$$

The energy computed from the sum (SYNC2 + SYNC3) is equal to the noise or interference seen by device 1108. Since the purpose of correlating the synchronization code in device 1106 is to extract the energy in SYNC 1, device 1108 also has the energy in the signal from base station 1110, i.e., the energy represented by SYNC1. Therefore, device 1106 can use the energy of SYNC1 and of (SYNC2 + SYNC3) to perform a signal-to-interference measurement for the communication channel over which it is communicating with base station 1110. The result of the measurement is preferably a signal-to-interference ratio (SIR). The SIR measurement can then be communicated back to base station 1110 for purposes that will be discussed below. The ideal cross correlation of the synchronization codes, also allows device 1108 to perform extremely accurate determinations of the Channel Impulse Response (CIR), or channel estimation, from the correlation produced by correlator 800. This allows for highly accurate equalization using low cost, low complexity equalizers, thus overcoming a significant draw back of conventional systems.

4. Sub-Channel Assignments

As mentioned, the SIR as determined by device 1108 can be communicated back to base station 1110 for use in the assignment of channels 502. In one embodiment, due to the fact that each sub-channel 502 is processed independently, the SIR for each sub-channel 502 can be measured and communicated back to base station 1110. In such an embodiment, therefore, sub-channels 502 can be divided into groups and a SIR measurement for each group can be sent to base station 1110. This is illustrated in figure 11A, which shows a wideband communication channel 1200 segmented into sub-channels f_0 to f_{15}. Sub-channels f_0 to f_{15} are then grouped into 8 groups Gl to G8. Thus, in one embodiment, device 1108 and base station 1110 communicate over a channel such as channel 1200.

Sub-channels in the same group are preferably separated by as many sub-channels as possible to ensure diversity. In figure 12A for example, sub-channels within the same group are 7 sub-channels apart, e.g., group Gl comprises f_0 and f_8. Device 1102 reports a SIR measurement for each of the groups G1 to G8. These SIR measurements are preferably compared with a threshold value to determine which sub-channels groups are useable by device 1108. This comparison can occur in device 1108 or base station 1110. If it occurs in device 1108, then device 1108 can simply report to base station 1110 which sub-channels groups are useable by device 1108.

SIR reporting will be simultaneously occurring for a plurality of devices within cell 1102. Thus, figure 12B illustrates the situation where two communication devices corresponding to User 1 and User 2 report SIR levels above the threshold for groups Gl, G3, G5, and G7. Base station 1110 preferably then assigns sub-channel groups to User 1 and User 2 based on the SIR reporting as illustrated in Figure 12B. When assigning the "good" sub-channel groups to User 1 and User 2, base station 1110 also preferably assigns them based on the principles of frequency diversity. In figure 12B, therefore, User 1 and User 2 are alternately assigned every other "good" sub-channel.

The assignment of sub-channels in the frequency domain is equivalent to the assignment of time slots in the time domain. Therefore, as illustrated in figure 13, two users, User 1 and User 2, receive packet 1302 transmitted over communication channel 1200. Figure 13 also illustrated the sub-channel assignment of figure 12B. While figures 12 and 13 illustrate sub-channel/time slot assignment based on SIR for two users, the principles illustrated can be extended for any number of users. Thus, a packet within cell 1102 can be received by 3 or more users. Although, as the number of available subchannels is reduced due to high SIR, so is the available bandwidth. In other words, as available channels are reduced, the number of users that can gain access to communication channel 1200 is also reduced.

Poor SIR can be caused for a variety of reasons, but frequently it results from a device at the edge of a cell receiving communication signals from adjacent cells. Because each cell is using the same bandwidth *B*, the adjacent cell signals will eventually raise the noise level and degrade SIR for certain sub-channels. In certain embodiments, therefore, sub-channel assignment can be coordinated between cells, such as cells 1102, 1104, and 1106 in figure 11, in order to prevent interference from adjacent cells.

Thus, if communication device 1108 is near the edge of cell 1102, and device 1118 is near the edge of cell 1106, then the two can interfere with each other. As a result, the SIR measurements that device 1108 and 1118 report back to base stations 1110 and 1114, respectively, will indicate that the interference level is too high. Base station 1110 can then be configured to assign only the odd groups, i.e., Gl, G3, G5, etc., to device 1108, while base station 1114 can be configured to assign the even groups to device 1118. The two devices 1108 and 1118 will then not interfere with each other due to the coordinated assignment of sub-channel groups.

Assigning the sub-channels in this manner reduces the overall bandwidth available to devices 1108 and 1118, respectively. In this case the bandwidth is reduced by a factor of two. But it should be remembered that devices operating closer to each base station 1110 and 1114, respectively, will still be able to use all channels if needed. Thus, it is only devices, such as device 1108, that are near the edge of a cell that will have the available bandwidth reduced. Contrast this with a CDMA system, for example, in which the bandwidth for all users is reduced, due to the spreading techniques used in such systems, by approximately a factor of 10 at all times. It can be seen, therefore, that the systems and methods for wireless communication over a wide bandwidth channel using a plurality of sub-channels not only improves the quality of service, but can also increase the available bandwidth significantly.

When there are three devices 1108, 1118, and 1116 near the edge of their respective adjacent cells 1102, 1104, and 1106, the sub-channels can be divided by three. Thus, device 1108, for example, can be assigned groups Gl, G4, etc., device 1118 can be assigned groups G2, G5, etc., and device 1116 can be assigned groups G3, G6, etc. In this case the available bandwidth for these devices, i.e., devices near the edges of cells 1102, 1104, and 1106, is reduced by a factor of 3, but this is still better than a CDMA system, for example.

The manner in which such a coordinated assignment of sub-channels can work is illustrated by the flow chart in figure 14. First in step 1402, a communication device, such as device 1108, reports the SIR for all sub-channel groups G1 to G8. The SIRs reported are then compared, in step 1404, to a threshold to determine if the SIR is sufficiently low for each group. Alternatively, device 1108 can make the determination and simply report which groups are above or below the SIR threshold. If the SIR levels are good for each group, then base station 1110 can make each group available to device 1108, in step 1406. Periodically, device 1108 preferably measures the SIR level and updates base station 1110 in case the SIR as deteriorated. For example, device 1108 may move from near the center of cell 1102 toward the edge, where interference from an adjacent cell may affect the SIR for device 1108.

If the comparison in step 1404 reveals that the SIR levels are not good, then base station 1110 can be preprogrammed to assign either the odd groups or the even groups only to device 1108, which it will do in step 1408. Device 1108 then reports the SIR measurements for the odd or even groups it is assigned in step 1410, and they are again compared to a SIR threshold in step 1412.

It is assumed that the poor SIR level is due to the fact that device 1108 is operating at the edge of cell 1102 and is therefore being interfered with by a device such as device 1118. But device 1108 will be interfering with device 1118 at the same time. Therefore, the assignment of odd or even groups in step 1408 preferably corresponds with the assignment of the opposite groups to device 1118, by base station 1114. Accordingly, when device 1108 reports the SIR measurements for whichever groups, odd or even, are assigned to it, the comparison in step 1410 should reveal that the SIR levels are now below the threshold level. Thus, base station 1110 makes the assigned groups available to device 1108 in step 1414. Again, device 1108 preferably periodically updates the SIR measurements by returning to step 1402.

It is possible for the comparison of step 1410 to reveal that the SIR levels are still above the threshold, which should indicate that a third device, e.g., device 1116 is still interfering with device 1108. In this case, base station 1110 can be preprogrammed to assign every third group to device 1108 in step 1416. This should correspond with the corresponding assignments of non-interfering channels to devices 1118 and 1116 by base stations 1114 and 1112, respectively. Thus, device 1108 should be able to operate on the sub-channel groups assigned, i.e., G1, G4, etc., without undue interference. Again, device 1108 preferably periodically updates the SIR measurements by returning to step 1402. Optionally, a third comparison step (not shown) can be implemented after step 1416, to ensure that the groups assigned to device 1408 posses an adequate SIR level for proper operation. Moreover, if there are more adjacent cells, i.e., if it is possible for devices in a 4th or even a 5th adjacent cell to interfere with device 1108, then the process of figure 13 would continue and the sub-channel groups would be divided even further to ensure adequate SIR levels on the sub-channels assigned to device 1108.

Even though the process of figure 14 reduces the bandwidth available to devices at the edge of cells 1102, 1104, and 1106, the SIR measurements can be used in such a manner as to increase the data rate and therefore restore or even increase bandwidth. To accomplish this, the transmitters and receivers used in base stations 1102, 1104, and 1106, and in devices in communication therewith, e.g., devices 1108, 1114, and 1116 respectively, must be capable of dynamically changing the symbol mapping schemes used for some or all of the sub-channel. For example, in some embodiments, the symbol mapping scheme can be dynamically changed among BPSK, QPSK, 8PSK, 16QAM, 32QAM, etc. As the symbol mapping scheme moves higher, i.e., toward 32QAM, the SIR level required for proper operation moves higher, i.e., less and less interference can be withstood. Therefore, once the SIR levels are determined for each group, the base station, e.g., base station 1110, can then determine what symbol mapping scheme can be supported for each sub-channel group and can change the modulation scheme accordingly. Device 1108 must also change the symbol mapping scheme to correspond to that of the base stations. The change can be effected for all groups uniformly, or it can be effected for individual groups. Moreover, the symbol mapping scheme can be changed on just the forward link, just the reverse link, or both, depending on the embodiment.

Thus, by maintaining the capability to dynamically assign sub-channels and to dynamically change the symbol mapping scheme used for assigned sub-channels, the systems and methods described herein provide the ability to maintain higher available bandwidths with higher performance levels than conventional systems. To fully realize the benefits described, however, the systems and methods described thus far must be capable of implementation in a cost effect and convenient manner. Moreover, the implementation must include reconfigurability so that a single device can move between different types of communication systems and still maintain optimum performance in accordance with the systems and methods described herein. The following descriptions detail example high level embodiments of hardware implementations configured to operate in accordance with the systems and methods described herein in such a manner as to provide the capability just described above.

5. Sample Transmitter Embodiments

Figure 15 is logical block diagram illustrating an example embodiment of a transmitter 1500 configured for wireless communication in accordance with the systems and methods described above. The transmitter could, for example be within a base station, e.g., base station 606, or within a communication device, such as device 604. Transmitter 1500 is provided to illustrate logical components that can be included in a transmitter configured in accordance with the systems and methods described herein. It is not intended to limit the systems and methods for wireless communication over a wide bandwidth channel using a plurality of sub-channels to any particular transmitter configuration or any particular wireless communication system.

With this in mind, it can be seen that transmitter 1500 comprises a serial-to-parallel converter 1504 configured to receive a serial data stream 1502 comprising a data rate *R*. Serial-to-parallel converter 1504 converts data stream 1502 into *N* parallel data streams 1504, where *N* is the number of sub-channels 202. It should be noted that while the discussion that follows assumes that a single serial data stream is used, more than one serial data stream can also be used if required or desired. In any case, the data rate of each parallel data stream 1504 is then *R/N*. Each data stream 1504 is then sent to a scrambler, encoder, and interleaver block 1506. Scrambling, encoding, and interleaving are common techniques implemented in many wireless communication transmitters and help to provide robust, secure communication. Examples of these techniques will be briefly explained for illustrative purposes.

Scrambling breaks up the data to be transmitted in an effort to smooth out the spectral density of the transmitted data. For example, if the data comprises a long string of "1"s, there will be a spike in the spectral density. This spike can cause greater interference within the wireless communication system. By breaking up the data, the spectral density can be smoothed out to avoid any such peaks. Often, scrambling is achieved by XORing the data with a random sequence.

Encoding, or coding, the parallel bit streams 1504 can, for example, provide Forward Error Correction (FEC). The purpose of FEC is to improve the capacity of a communication channel by adding some carefully designed redundant information to the data being transmitted through the channel. The process of adding this redundant information is known as channel coding. Convolutional coding and block coding are the two major forms of channel coding. Convolutional codes operate on serial data, one or a few bits at a time. Block codes operate on relatively large (typically, up to a couple of hundred bytes) message blocks. There are a variety of useful convolutional and block codes, and a variety of algorithms for decoding the received coded information sequences to recover the original data. For example, convolutional encoding or turbo coding with Viterbi decoding is a FEC technique that is particularly suited to a channel in which the transmitted signal is corrupted mainly by additive white gaussian noise (AWGN) or even a channel that simply experiences fading.

Convolutional codes are usually described using two parameters: the code rate and the constraint length. The code rate, *k/n*, is expressed as a ratio of the number of bits into the convolutional encoder *(k)* to the number of channel symbols *(n)* output by the convolutional encoder in a given encoder cycle. A common code rate is 1/2, which means that 2 symbols are produced for every 1-bit input into the coder. The constraint length parameter, *K*, denotes the "length" of the convolutional encoder, i.e. how many *k*-bit stages are available to feed the combinatorial logic that produces the output symbols. Closely related to *K* is the parameter *m*, which indicates how many encoder cycles an input bit is retained and used for encoding after it first appears at the input to the convolutional encoder. The *m* parameter can be thought of as the memory length of the encoder.

Interleaving is used to reduce the effects of fading. Interleaving mixes up the order of the data so that if a fade interferes with a portion of the transmitted signal, the overall message will not be effected. This is because once the message is de-interleaved and decoded in the receiver, the data lost will comprise non-contiguous portions of the overall message. In other words, the fade will interfere with a contiguous portion of the interleaved message, but when the message is de-interleaved, the interfered with portion is spread throughout the overall message. Using techniques such as FEC, the missing information can then be filled in, or the impact of the lost data may just be negligible.

After blocks 1506, each parallel data stream 1504 is sent to symbol mappers 1508. Symbol mappers 1508 apply the requisite symbol mapping, e.g., BPSK, QPSK, etc., to each parallel data stream 1504. Symbol mappers 1508 are preferably programmable so that the modulation applied to parallel data streams can be changed, for example, in response to the SIR reported for each sub-channel 202. It is also preferable, that each symbol mapper 1508 be separately programmable so that the optimum symbol mapping scheme for each sub-channel can be selected and applied to each parallel data stream 1504.

After symbol mappers 1508, parallel data streams 1504 are sent to modulators 1510. Important aspects and features of example embodiments of modulators 1510 are described below. After modulators 1510, parallel data streams 1504 are sent to summer 1512, which is configured to sum the parallel data streams and thereby generate a single serial data stream 1518 comprising each of the individually processed parallel data streams 1504. Serial data stream 1518 is then sent to radio module 1512, where it is modulated with an RF carrier, amplified, and transmitted via antenna 1516 according to known techniques.

The transmitted signal occupies the entire bandwidth B of communication channel 100 and comprises each of the discrete parallel data streams 1504 encoded onto their respective sub-channels 102 within bandwidth B. Encoding parallel data streams 1504 onto the appropriate sub-channels 102 requires that each parallel data stream 1504 be shifted in frequency by an appropriate offset. This is achieved in modulator 1510. Figure 16 is a logical block diagram of an example embodiment of a modulator 1600 in accordance with the systems and methods described herein. Importantly, modulator 1600 takes parallel data streams 1602 performs Time Division Modulation (TDM) or Frequency Division Modulation (FDM) on each data stream 1602, filters them using filters 1612, and then shifts each data stream in frequency using frequency shifter 1614 so that they occupy the appropriate sub-channel. Filters 1612 apply the required pulse shaping, i.e., they apply the roll-off factor described in section 1. The frequency shifted parallel data streams 1602 are then summed and transmitted. Modulator 1600 can also include rate controller 1604, frequency encoder 1606, and interpolators 1610. All of the components shown in figure 16 are described in more detail in the following paragraphs and in conjunction with figures 17-23.

Figure 17 illustrates one example embodiment of a rate controller 1700 in accordance with the systems and methods described herein. Rate control 1700 is used to control the data rate of each parallel data stream 1602. In rate controller 1700, the data rate is halved by repeating data streams *d(0)* to *d(7),* for example, producing streams *a(0)* to *a(15)* in which *a(0)* is the same as *a(8), a(1)* is the same as *a(9),* etc. Figure 17 also illustrates that the effect of repeating the data streams in this manner is to take the data streams that are encoded onto the first 8 sub-channels 1702, and duplicate them on the next 8 sub-channels 1702. As can be seen, 7 sub-channels separate sub-channels 1702 comprising the same, or duplicate, data streams. Thus, if fading effects one sub-channel 1702, for example, the other sub-channels 1702 carrying the same data will likely not be effected, i.e., there is frequency diversity between the duplicate data streams. So by sacrificing data rate, in this case half the data rate, more robust transmission is achieved. Moreover, the robustness provided by duplicating the data streams *d(0)* to *d(7)* can be further enhanced by applying scrambling to the duplicated data streams via scramblers 1708.

It should be noted that the data rate can be reduced by more than half, e.g., by four or more. Alternatively, the data rate can also be reduced by an amount other than half. For example if information from n data stream is encoded onto m sub-channels, where $m > n$. Thus, to decrease the rate by 2/3, information from one data stream can be encoded on a first sub-channel, information from a second data stream can be encoded on a second data channel, and the sum or difference of the two data streams can be encoded on a third channel. In which case, proper scaling will need to be applied to the power in the third channel. Otherwise, for example, the power in the third channel can be twice the power in the first two.

Preferably, rate controller 1700 is programmable so that the data rate can be changed responsive to certain operational factors. For example, if the SIR reported for sub-channels 1702 is low, then rate controller 1700 can be programmed to provide more robust transmission via repetition to ensure that no data is lost due to interference. Additionally, different types of wireless communication system, e.g., indoor, outdoor, line-of-sight, may require varying degrees of robustness. Thus, rate controller 1700 can be adjusted to provide the minimum required robustness for the particular type of communication system. This type of programmability not only ensures robust communication, it can also be used to allow a single device to move between communication systems and maintain superior performance.

Figure 18 illustrates an alternative example embodiment of a rate controller 1800 in accordance with the systems and methods described. In rate controller 1800 the data rate is increased instead of decreased. This is accomplished using serial-to- parallel converters 1802 to convert each data streams *d(0)* to *d(15)*, for example, into two data streams. Delay circuits 1804 then delay one of the two data streams generated by each serial-to-parallel converter 1802 by 1/2 a symbol. Thus, data streams *d(0)* to *d(15)* are transformed into data streams *a(0)* to *a(31)*. The data streams generated by a particular serial-to-parallel converter 1802 and associate delay circuit 1804 must then be summed and encoded onto the appropriate sub-channel. For example, data streams *a(0)* and *a(1)* must be summed and encoded onto the first sub-channel. Preferably, the data streams are summed subsequent to each data stream being pulsed shaped by a filter 1612.

Thus, rate controller 1604 is preferably programmable so that the data rate can be increased, as in rate controller 1800, or decreased, as in rate controller 1700, as required by a particular type of wireless communication system, or as required by the communication channel conditions or sub-channel conditions. In the event that the data rate is increased, filters 1612 are also preferably programmable so that they can be configured to apply pulse shaping to data streams *a(0)* to *a(31)*, for example, and then sum the appropriate streams to generate the appropriate number of parallel data streams to send to frequency shifter 1614.

The advantage of increasing the data rate in the manner illustrated in figure 18 is that higher symbol mapping rates can essentially be achieved, without changing the symbol mapping used in symbol mappers 1508. Once the data streams are summed, the summed streams are shifted in frequency so that they reside in the appropriate sub-channel. But because the number of bits per each symbol has been doubled, the symbol mapping rate has been doubled. Thus, for example, a 4QAM symbol mapping can be converted to a 16QAM symbol mapping, even if the SIR is too high for 16QAM symbol mapping to otherwise be applied. In other words, programming rate controller 1800 to increase the data rate in the manner illustrated in figure 18 can increase the symbol mapping even when channel conditions would otherwise not allow it, which in turn can allow a communication device to maintain adequate or even superior performance regardless of the type of communication system.

The draw back to increasing the data rate as illustrated in figure 18 is that interference is increased, as is receiver complexity. The former is due to the increased amount of data. The latter is due to the fact that each symbol cannot be processed independently because of the 1/2 symbol overlap. Thus, these concerns must be balanced against the increase symbol mapping ability when implementing a rate controller such as rate controller 1800.

Figure 19 illustrates one example embodiment of a frequency encoder 1900 in accordance with the systems and methods described herein. Similar to rate encoding, frequency encoding is preferably used to provide increased communication robustness. In frequency encoder 1900 the sum or difference of multiple data streams are encoded onto each sub-channel. This is accomplished using adders 1902 to sum data streams *d(0)* to *d(7)* with data streams *d(8)* to *d(15)*, respectively, while adders 1904 subtract data streams *d(0)* to *d(7)* from data streams *d(8)* to *d(15)*, respectively, as shown. Thus, data streams *a(0)* to *a(15)* generated by adders 1902 and 1904 comprise information related to more than one data streams *d(0)* to *d(15)*. For example, *a(0)* comprises the sum of *d(0)* and *d(8)*, i.e., $d(0) + d(8)$, while *a(8)* comprises $d(8) - d(0)$. Therefore, if either *a(0)* or *a(8)* is not received due to fading, for example, then both of data streams *d(0)* and *d(8)* can still be retrieved from data stream *a(8)*.

Essentially, the relationship between data stream *d(0)* to *d(15)* and *a(0)* to *a(15)* is a matrix relationship. Thus, if the receiver knows the correct matrix to apply, it can recover the sums and differences of *d(0)* to *d(15)* from *a(0) to a(15)*. Preferably, frequency encoder 1900 is programmable, so that it can be enabled and disabled in order to provided robustness when required. Preferable, adders 1902 and 1904 are programmable also so that different matrices can be applied to *d(0)* to *d(15)*.

After frequency encoding, if it is included, data streams 1602 are sent to TDM/FDM blocks 1608. TDM/FDM blocks 1608 perform TDM or FDM on the data streams as required by the particular embodiment. Figure 20 illustrates an example embodiment of a TDM/FDM block 2000 configured to perform TDM on a data stream. TDM/FDM block 2000 is provided to illustrate the logical components that can be included in a TDM/FDM block configured to perform TDM on a data stream. Depending on the actual implementation, some of the logical components may or may not be included. TDM/FDM block 2000 comprises a sub-block repeater 2002, a sub- block scrambler 2004, a sub-block terminator 2006, a sub-block repeater 2008, and a sync inserter 2010.

Sub-block repeater 2002 is configured to receive a sub-block of data, such as block 2012 comprising bits *a(0)* to *a(3)* for example. Sub-block repeater is then configured to repeat block 2012 to provide repetition, which in turn leads to more robust communication. Thus, sub-block repeater 2002 generates block 2014, which comprises 2 blocks 2012.

Sub-block scrambler 2004 is then configured to receive block 2014 and to scramble it, thus generating block 2016. One method of scrambling can be to invert half of block 2014 as illustrated in block 2016. But other scrambling methods can also be implemented depending on the embodiment.

Sub-block terminator 2006 takes block 2016 generated by sub-block scrambler 2004 and adds a termination block 2034 to the front of block 2016 to form block 2018. Termination block 2034 ensures that each block can be processed independently in the receiver. Without termination block 2034, some blocks may be delayed due to multipath, for example, and they would therefore overlap part of the next block of data. But by including termination block 2034, the delayed block can be prevented from overlapping any of the actual data in the next block.

Termination block 2034 can be a cyclic prefix termination 2036. A cyclic prefix termination 2036 simply repeats the last few symbols of block 2018. Thus, for example, if cyclic prefix termination 2036 is three symbols long, then it would simply repeat the last three symbols of block 2018. Alternatively, termination block 2034 can comprise a sequence of symbols that are known to both the transmitter and receiver. The selection of what type of block termination 2034 to use can impact what type of equalizer is used in the receiver. Therefore, receiver complexity and choice of equalizers must be considered when determining what type of termination block 2034 to use in TDM/FDM block 2000.

After sub-block terminator 2006, TDM/FDM block 2000 can include a sub-block repeater 2008 configured to perform a second block repetition step in which block 2018 is repeated to form block 2020. In certain embodiments, sub-block repeater can be configured to perform a second block scrambling step as well. After sub-block repeater 2008, if included, TDM/FDM block 2000 comprises a sync inserter 210 configured to periodically insert an appropriate synchronization code 2032 after a predetermined number of blocks 2020 and/or to insert known symbols into each block. The purpose of synchronization code 2032 is discussed in section 3.

Figure 21, on the other hand, illustrates an example embodiment of a TDM/FDM block 2100 configured for FDM, which comprises sub-block repeater 2102, sub-block scrambler 2104, block coder 2106, sub-block transformer 2108, sub-block terminator 2110, and sync inserter 2112. As with TDM/FDM block 2000, sub-block repeater 2102 repeats block 2114 and generates block 2116. Sub-block scrambler then scrambles block 2116, generating block 2118. Sub-block coder 2106 takes block 2118 and codes it, generating block 2120. Coding block correlates the data symbols together and generates symbols *b*. This requires joint demodulation in the receiver, which is more robust but also more complex. Sub-block transformer 2108 then performs a transformation on block 2120, generating block 2122. Preferably, the transformation is an IFFT of block 2120, which allows for more efficient equalizers to be used in the receiver. Next, sub-block terminator 2110 terminates block 2122, generating block 2124 and sync inserter 2112 periodically inserts a synchronization code 2126 after a certain number of blocks 2124 and/or insert known symbols into each block. Preferably, sub-block terminator 2110 only uses cyclic prefix termination as described above. Again this allows for more efficient receiver designs.

TDM/FDM block 2100 is provided to illustrate the logical components that can be included in a TDM/FDM block configured to perform FDM on a data stream. Depending on the actual implementation, some of the logical components may or may not be included. Moreover, TDM/FDM block 2000 and 2100 are preferably programmable so that the appropriate logical components can be included as required by a particular implementation. This allows a device that incorporates one of blocks 2000 or 2100 to move between different systems with different requirements. Further, it is preferable that TDM/FDM block 1608 in figure 16 be programmable so that it can be programmed to perform TDM, such as described in conjunction with block 2000, or FDM, such as described in conjunction with block 2100, as required by a particular communication system.

After TDM/FDM blocks 1608, in figure 15, the parallel data streams are preferably passed to interpolators 1610. After Interpolators 1610, the parallel data streams are passed to filters 1612, which apply the pulse shaping described in conjunction with the roll-off factor of equation (2) in section 1. Then the parallel data streams are sent to frequency shifter 1614, which is configured to shift each parallel data stream by the frequency offset associated with the sub-channel to which the particular parallel data stream is associated. Figure 22 illustrates an example embodiment of a frequency shifter 2200 in accordance with the systems and methods described herein. As can be seen, frequency shifter 2200 comprises multipliers 2202 configured to multiply each parallel data stream by the appropriate exponential to achieve the required frequency shift. Each exponential is of the form: $exp(j2\pi f_c nT/rM)$, where c is the corresponding sub-channel, e.g., $c = 0$ to $N-1$, and n is time. Preferably, frequency shifter 1614 in figure 16 is programmable so that various channel/sub-channel configurations can be accommodated for various different systems. Alternatively, an IFFT block can replace shifter 1614 and filtering can be done after the IFFT block. This type of implementation can be more efficient depending on the implementation.

After the parallel data streams are shifted, they are summed, e.g., in summer 1512 of figure 15. The summed data stream is then transmitted using the entire bandwidth B of the communication channel being used. But the transmitted data stream also comprises each of the parallel data streams shifted in frequency such that they occupy the appropriate sub-channel. Thus, each sub-channel may be assigned to one user, or each sub-channel may carry a data stream intended for different users. The assignment of sub-channels is described in section 3b. Regardless of how the sub-channels are assigned, however, each user will receive the entire bandwidth, comprising all the sub-channels, but will only decode those sub-channels assigned to the user.

6. <u>Sample Receiver Embodiments</u>

Figure 23 illustrates an example embodiment of a receiver 2300 that can be configured in accordance with the present invention. Receiver 2300 comprises an antenna 2302 configured to receive a message transmitted by a transmitter, such as transmitter 1500. Thus, antenna 2302 is configured to receive a wide band message comprising the entire bandwidth B of a wide band channel that is divided into sub- channels of bandwidth b. As described above, the wide band message comprises a plurality of messages each encoded onto each of a corresponding sub-channel. All of the sub-channels may or may not be assigned to a device that includes receiver 2300; Therefore, receiver 2300 may or may not be required to decode all of the sub-channels. After the message is received by antenna 2300, it is sent to radio receiver 2304, which is configured to remove the carrier associated with the wide band communication channel and extract a baseband signal comprising the data stream transmitted by the transmitter. The baseband signal is then sent to correlator 2306 and demodulator 2308. Correlator 2306 is configured to correlated with a synchronization code inserted in the data stream as described in section 3. It is also preferably configured to perform SIR and multipath estimations as described in section 3(b). Demodulator 2308 is configured to extract the parallel data streams from each sub-channel assigned to the device comprising receiver 2300 and to generate a single data stream therefrom.

Figure 24 illustrates an example embodiment of a demodulator 2400 in accordance with the systems and methods described herein. Demodulator 2402 comprises a frequency shifter 2402, which is configured to apply a frequency offset to the baseband data stream so that parallel data streams comprising the baseband data stream can be independently processed in receiver 2400. Thus, the output of frequency shifter 2402 is a plurality of parallel data streams, which are then preferably filtered by filters 2404. Filters 2404 apply a filter to each parallel data stream that corresponds to the pulse shape applied in the transmitter, e.g., transmitter 1500. Alternatively, an IFFT block can replace shifter 1614 and filtering can be done after the IFFT block. This type of implementation can be more efficient depending on the implementation.

Next, receiver 2400 preferably includes decimators 2406 configured to decimate the data rate of the parallel bit streams. Sampling at higher rates helps to ensure accurate recreation of the data. But the higher the data rate, the larger and more complex equalizer 2408 becomes. Thus, the sampling rate, and therefore the number of samples, can be reduced by decimators 2406 to an adequate level that allows for a smaller and less costly equalizer 2408.

Equalizer 2408 is configured to reduce the effects of multipath in receiver 2300. Its operation will be discussed more fully below. After equalizer 2408, the parallel data streams are sent to de-scrambler, decoder, and de-interleaver 2410, which perform the opposite operations of scrambler, encoder, and interleaver 1506 so as to reproduce the original data generated in the transmitter. The parallel data streams are then sent to parallel to serial converter 2412, which generates a single serial data stream from the parallel data streams.

Equalizer 2408 uses the multipath estimates provided by correlator 2306 to equalize the effects of multipath in receiver 2300. In one embodiment, equalizer 2408 comprises Single-In Single-Out (SISO) equalizers operating on each parallel data stream in demodulator 2400. In this case, each SISO equalizer comprising equalizer 2408 receives a single input and generates a single equalized output. Alternatively, each equalizer can be a Multiple-In Multiple-Out (MIMO) or a Multiple-In Single-Out (MISO) equalizer. Multiple inputs can be required for example, when a frequency encoder or rate controller, such as frequency encoder 1900, is included in the transmitter. Because frequency encoder 1900 encodes information from more than one parallel data stream onto each sub-channel, each equalizers comprising equalizer 2408 need to equalize more than one sub-channel. Thus, for example, if a parallel data stream in demodulator 2400 comprises $d(1) + d(8)$, then equalizer 2408 will need to equalize both $d(1)$ and $d(8)$ together. Equalizer 2408 can then generate a single output corresponding to $d(1)$ or $d(8)$ (MISO) or it can generate both $d(1)$ and $d(8)$ (MIMO).

Equalizer 2408 can also be a time domain equalizer (TDE) or a frequency domain equalizer (FDE) depending on the embodiment. Generally, equalizer 2408 is a TDE if the modulator in the transmitter performs TDM on the parallel data streams, and a FDE if the modulator performs FDM. But equalizer 2408 can be an FDE even if TDM is used in the transmitter. Therefore, the preferred equalizer type should be taken into consideration when deciding what type of block termination to use in the transmitter. Because of power requirements, it is often preferable to use FDM on the forward link and TDM on the reverse link in a wireless communication system.

As with transmitter 1500, the various components comprising demodulator 2400 are preferably programmable, so that a single device can operate in a plurality of different systems and still maintain superior performance, which is a primary advantage of the systems and methods described herein. Accordingly, the above discussion provides systems and methods for implementing a channel access protocol that allows the transmitter and receiver hardware to be reprogrammed slightly depending on the communication system.

Thus, when a device moves from one system to another, it preferably reconfigures the hardware, i.e. transmitter and receiver, as required and switches to a protocol stack corresponding to the new system. An important part of reconfiguring the receiver is reconfiguring, or programming, the equalizer because multipath is a main problem for each type of system. The multipath, however, varies depending on the type of system, which previously has meant that a different equalizer is required for different types of communication systems. The channel access protocol described in the preceding sections, however, allows for equalizers to be used that need only be reconfigured slightly for operation in various systems.

a. Sample Equalizer Embodiment

Figure 25 illustrates an example embodiment of a receiver 2500 illustrating one way to configure equalizers 2506 in accordance with the systems and methods described herein. Before discussing the configuration of receiver 2500, it should be noted that one way to configure equalizers 2506 is to simply include one equalizer per channel (for the systems and methods described herein, a channel is the equivalent of a sub-channel as described above). A correlator, such as correlator 2306 (figure 23), can then provide equalizers 2506 with an estimate of the number, amplitude, and phase of any multipaths present, up to some maximum number. This is also known as the Channel Impulse Response (CIR). The maximum number of multipaths is determined based on design criteria for a particular implementation. The more multipaths included in the CIR the more path diversity the receiver has and the more robust communication in the system will be. Path diversity is discussed a little more fully below.

If there is one equalizer 2506 per channel, the CIR is preferably provided directly to equalizers 2506 from the correlator (not shown). If such a correlator configuration is used, then equalizers 2506 can be run at a slow rate, but the overall equalization process is relatively fast. For systems with a relatively small number of channels, such a configuration is therefore preferable. The problem, however, is that there is large variances in the number of channels used in different types of communication systems. For example, an outdoor system can have has many as 256 channels. This would require 256 equalizers 2506, which would make the receiver design too complex and costly. Thus, for systems with a lot of channels, the configuration illustrated in figure 25 is preferable. In receiver 2500, multiple channels share each equalizer 2506. For example, each equalizer can be shared by 4 channels, e.g., Ch1-Ch4, Ch5-Ch8, etc., as illustrated in figure 25.

In which case, receiver 2500 preferably comprises a memory 2502 configured to store information arriving on each channel.

Memory 2502 is preferably divided into sub-sections 2504, which are each configured to store information for a particular subset of channels. Information for each channel in each subset is then alternately sent to the appropriate equalizer 2506, which equalizes the information based on the CIR provided for that channel. In this case, each equalizer must run much faster than it would if there was simply one equalizer per channel. For example, equalizers 2506 would need to run 4 or more times as fast in order to effectively equalize 4 channels as opposed to 1. In addition, extra memory 2502 is required to buffer the channel information. But overall, the complexity of receiver 2500 is reduced, because there are fewer equalizers. This should also lower the overall cost to implement receiver 2500.

Preferably, memory 2502 and the number of channels that are sent to a particular equalizer is programmable. In this way, receiver 2500 can be reconfigured for the most optimum operation for a given system. Thus, if receiver 2500 were moved from an outdoor system to an indoor system with fewer channels, then receiver 2500 can preferably be reconfigured so that there are fewer, even as few as 1, channel per equalizer. The rate at which equalizers 2506 are run is also preferably programmable such that equalizers 2506 can be run at the optimum rate for the number of channels being equalized.

In addition, if each equalizer 2506 is equalizing multiple channels, then the CIR for those multiple paths must alternately be provided to each equalizer 2506. Preferably, therefore, a memory (not shown) is also included to buffer the CIR information for each channel. The appropriate CIR information is then sent to each equalizer from the CIR memory (not shown) when the corresponding channel information is being equalized. The CIR memory (not shown) is also preferably programmable to ensure optimum operation regardless of what type of system receiver 2500 is operating in.

Returning to the issue of path diversity, the number of paths used by equalizers 2506 must account for the delay spread d_s in the system. For example, if the system is an outdoor system operating in the 5 Giga Hertz (GHz) range, the communication channel can comprise a bandwidth of 125 Mega Hertz (MHz), e.g., the channel can extend from 5.725 GHz to 5.85GHz. If the channel is divided into 512 sub-channels with a roll-off factor r of .125, then each subchannel will have a bandwidth of approximately 215 kilohertz (KHz), which provides approximately a 4.6 microsecond symbol duration. Since the worst case delay spread d_s is 20 microseconds, the number of paths used by equalizers 2504 can be set to a maximum of 5. Thus, there would be a first path P1 at zero microseconds, a second path P2 at 4.6 microseconds, a third path P3 at 9.2 microseconds, a fourth path P4 at 13.8 microseconds, and fifth path P5 at 18.4 microseconds, which is close to the delay spread d_s. In another embodiment, a sixth path can be included so as to completely cover the delay spread d_s; however, 20 microseconds is the worst case. In fact, a delay spread d_s of 3 microseconds is a more typical value. In most instances, therefore, the delay spread d_s will actually be shorter and an extra path is not needed. Alternatively, fewer sub-channels can be used, thus providing a larger symbol duration, instead of using an extra path. But again, this would typically not be needed.

As explained above, equalizers 2506 are preferably configurable so that they can be reconfigured for various communication systems. Thus, for example, the number of paths used must be sufficient regardless of the type of communication system. But this is also dependent on the number of sub-channels used. If, for example, receiver 2500 went from operating in the above described outdoor system to an indoor system, where the delay spread d_s is on the order of 1 microsecond, then receiver 2500 can preferably be reconfigured for 32 sub-channels and 5 paths. Assuming the same overall bandwidth of 125 MHz, the bandwidth of each sub-channel is approximately 4 MHz and the symbol duration is approximately 250 nanoseconds.

Therefore, there will be a first path P1 at zero microseconds and subsequent paths P2 to P5 at 250ns, 500ns, 750ns, and 1 microsecond, respectively. Thus, the delay spread *ds* should be covered for the indoor environment. Again, the 1 microsecond delay spread d_s is worst case so the 1 microsecond delay spread d_s provided in the above example will often be more than is actually required. This is preferable, however, for indoor systems, because it can allow operation to extend outside of the inside environment, e.g., just outside the building in which the inside environment operates. For campus style environments, where a user is likely to be traveling between buildings, this can be advantageous.

7. <u>Sample Embodiment of a Wireless Communication device</u>

Figure 26 illustrates an example embodiment of a wireless communication device in accordance with the systems and methods described herein. Device 2600 is, for example, a portable communication device configured for operation in a plurality of indoor and outdoor communication systems. Thus, device 2600 comprises an antenna 2602 for transmitting and receiving wireless communication signals over a wireless communication channel 2618. Duplexor 2604, or switch, can be included so that transmitter 2606 and receiver 2608 can both use antenna 2602, while being isolated from each other. Duplexors, or switches used for this purpose, are well known and will not be explained herein.

Transmitter 2606 is a configurable transmitter configured to implement the channel access protocol described above. Thus, transmitter 2606 is capable of transmitting and encoding a wideband communication signal comprising a plurality of sub-channels. Moreover, transmitter 2606 is configured such that the various sub-components that comprise transmitter 2606 can be reconfigured, or programmed, as described in section 5. Similarly, receiver 2608 is configured to implement the channel access protocol described above and is, therefore, also configured such that the various sub-components comprising receiver 2608 can be reconfigured, or reprogrammed, as described in section 6.

Transmitter 2606 and receiver 2608 are interfaced with processor 2610, which can comprise various processing, controller, and/or Digital Signal Processing (DSP) circuits. Processor 2610 controls the operation of device 2600 including encoding signals to be transmitted by transmitter 2606 and decoding signals received by receiver 2608. Device 2610 can also include memory 2612, which can be configured to store operating instructions, e.g., firmware/software, used by processor 2610 to control the operation of device 2600.

Processor 2610 is also preferably configured to reprogram transmitter 2606 and receiver 2608 via control interfaces 2614 and 2616, respectively, as required by the wireless communication system in which device 2600 is operating. Thus, for example, device 2600 can be configured to periodically ascertain the availability is a preferred communication system. If the system is detected, then processor 2610 can be configured to load the corresponding operating instruction from memory 2612 and reconfigure transmitter 2606 and receiver 2608 for operation in the preferred system.

For example, it may preferable for device 2600 to switch to an indoor wireless LAN if it is available. So device 2600 may be operating in a wireless WAN where no wireless LAN is available, while periodically searching for the availability of an appropriate wireless LAN. Once the wireless LAN is detected, processor 2610 will load the operating instructions, e.g., the appropriate protocol stack, for the wireless LAN environment and will reprogram transmitter 2606 and receiver 2608 accordingly. In this manner, device 2600 can move from one type of communication system to another, while maintaining superior performance.

It should be noted that a base station configured in accordance with the systems and methods herein will operate in a similar manner as device 2600; however, because the base station does not move from one type of system to another, there is generally no need to configure processor 2610 to reconfigure transmitter 2606 and receiver 2608 for operation in accordance with the operating instruction for a different type of system. But processor 2610 can still be configured to reconfigure, or reprogram the sub-components of transmitter 2606 and/or receiver 2608 as required by the operating conditions within the system as reported by communication devices in communication with the base station. Moreover, such a base station can be configured in accordance with the systems and methods described herein to implement more than one mode of operation. In which case, controller 2610 can be configured to reprogram transmitter 2606 and receiver 2608 to implement the appropriate mode of operation.

8. Bandwidth recovery

As described above in relation to figures 11-14, when a device, such as device 1118 is near the edge of a communication cell 1106, it may experience interference from base station 1112 of an adjacent communication cell 1104. In this case, device 1118 will report a low SIR to base station 1114, which will cause base station 1114 to reduce the number of sub-channels assigned to device 1118. As explained in relation to figures 12 and 13, this reduction can comprise base station 1114 assigning only even sub-channels to device 1118. Preferably, base station 1112 is correspondingly assigning only odd sub-channels to device 1116.

In this manner, base station 1112 and 1114 perform complementary reductions in the channels assigned to devices 1116 and 1118 in order to prevent interference and improve performance of devices 1116 and 1118. The reduction in assigned channels reduces the overall bandwidth available to devices 1116 and 1118. But as described above, a system implementing such a complementary reduction of sub-channels will still maintain a higher bandwidth than conventional systems. Still, it is preferable to recover the unused sub-channels, or unused bandwidth, created by the reduction of sub-channels in response to a low reported SIR.

One method for recovering the unused bandwidth is illustrated in the flow chart of figure 27. First, in step 2702, base station 1114 receives SIR reports for different groups of sub-channels from device 1118 as described above. If the group SIR reports are good, then base station 1114 can assign all sub-channels to device 1118 in step 2704. If, however, some of the group SIR reports received in step 2702 are poor, then base station 1114 can reduce the number of sub-channels assigned to device 1118, e.g., by assigning only even sub-channels, in step 2706. At the same time, base station 1112 is preferably performing a complementary reduction in the sub-channels assigned to device 1116, e.g., by assigning only odd sub-channels.

At this point, each base station has unused bandwidth with respect to devices 1116 and 1118. To recover this bandwidth, base station 1114 can, in step 2708, assign the unused odd sub-channels to device 1116 in adjacent cell 1104. It should be noted that even though cells 1102, 1104, and 1106 are illustrated as geometrically shaped, non-overlapping coverage areas, the actual coverage areas do not resemble these shapes. The shapes are essentially fictions used to plan and describe a wireless communication system 1100. Therefore, base station 1114 can in fact communicate with device 1116, even though it is in adjacent cell 1104.

Once base station 1114 has assigned the odd sub-channels to device 1116, in step 2708, base station 1112 and 1114 communicate with device 1116 simultaneously over the odd sub-channels in step 2710. Preferably, base station 1112 also assigns the unused even sub-channels to device 1118 in order to recover the unused bandwidth in cell 1104 as well.

In essence, spatial diversity is achieved by having both base station 1114 and 1112 communicate with device 1116 (and 1118) over the same sub-channels. Spatial diversity occurs when the same message is transmitted simultaneously over statistically independent communication paths to the same receiver. The independence of the two paths improves the overall immunity of the system to fading. This is because the two paths will experience different fading effects. Therefore, if the receiver cannot receive the signal over one path due to fading, then it will probably still be able to receive the signal over the other path, because the fading that effected the first path will not effect the second. As a result, spatial diversity improves overall system performance by improving the Bit Error Rate (BER) in the receiver, which effectively increases the deliverable data rate to the receiver, i.e., increase the bandwidth.

For effective spatial diversity, base stations 1112 and 1114 ideally transmit the same information at the same time over the same sub-channels. As mentioned above, each base station in system 1100 is configured to transmit simultaneously, i.e., system 1100 is a TDM system with synchronized base stations. Base stations 1112 and 1114 also assigned the same sub-channels to device 1116 in step 2708. Therefore, all that is left is to ensure that base stations 1112 and 1114 send the same information. Accordingly, the information communicated to device 1116 by base stations 1112 and 1114 is preferably coordinated so that the same information is transmitted at the same time. The mechanism for enabling this coordination is discussed more fully below. Such coordination, however, also allows encoding that can provide further performance enhancements within system 1100 and allow a greater percentage of the unused bandwidth to be recovered.

One example coordinated encoding scheme that can be implemented between base stations 1112 and 1114 with respect to communications with device 1116 is Space-Time-Coding (STC) diversity. STC is illustrated by system 2800 in figure 28. In system 2800, transmitter 2802 transmits a message over channel 2808 to receiver 2806. Simultaneously, transmitter 2804 transmits a message over channel 2810 to receiver 2806. Because channels 2808 and

2810 are independent, system 2800 will have spatial diversity with respect to communications from transmitters 2802 and 2804 to receiver 2806. In addition, however, the data transmitted by each transmitter 2802 and 2804 can be encoded to also provide time diversity. The following equations illustrate the process of encoding and decoding data in a STC system, such as system 2800.

First, channel 2808 can be denoted h_n and channel 2810 can be denoted g_n, where:

$$h_n = \alpha_h e^{j\theta h}\text{; and} \quad (1)$$
$$g_n = \alpha_g e^{j\theta g} \quad (2)$$

Second, we can look at two blocks of data 2812a and 2812b to be transmitted by transmitter 2802 as illustrated in figure 28. Block 2812a comprises N-symbols denoted as $a_0, a_1, a_2, ..., a_{N-1}$, or $a(0:N-1)$. Block 2812b transmits N-symbols of data denoted $b(0: N-1)$. Transmitter 2804 simultaneously transmits two block of data 2814a and 2814b. Block 2814a is the negative inverse conjugate of block 2812b and can therefore be described as $-b^*(N-1:0)$. Block 2814b is the inverse conjugate of block 2812a and can therefore be described as $a^*(N-1:0)$. It should be noted that each block of data in the forgoing description will preferably comprise a cyclical prefix as described above.

When blocks 2812a, 2812b, 2814a, and 2814b are received in receiver 2806, they are combined and decoded in the following manner: First, the blocks will be combined in the receiver to form the following blocks, after discarding the cyclical prefix:

$$Block1 = a(0:N-1) \otimes h_n - b^*(N-1:0) \otimes g_n\text{; and} \quad (3)$$

$$Block2 = b(0:-N-1) \otimes h_n + a^*(N-1:0) \otimes g_n. \quad (4)$$

Where the symbol $\otimes$ represents a cyclic convolution.

Second, by taking an IFFT of the blocks, the blocks can be described as:

$$Block1 = A_n * H_n - B_n^* * G_n\text{; and} \quad (5)$$

$$Block2 = B_n * H_n - A_n^* * G_n. \quad (6)$$

Where $n = 0$ to $N-1$.

In equations (5) and (6) H_n and G_n will be known, or can be estimated. But to solve the two equations and determine A_n and B_n, it is preferable to turn equations (5) and (6) into two equations with two unknowns. This can be achieved using estimated signals X_n and Y_n as follows:

$$X_n = A_n * H_n - B_n^* * G_n\text{; and} \quad (7)$$

$$Y_n = B_n * H_n + A_n^* * G_n. \quad (8)$$

To generate two equations and two unknowns, the conjugate of Y_n can be used to generate the following two equations:

$$X_n = A_n * H_n - B_n^* * G_n\text{; and} \quad (9)$$

$$Y_n^* = B_n^* * H_n^* + A_n * G_n^*. \quad (10)$$

Thus, the two unknowns are A_n and B_n* and equations (9) and (10) define a matrix relationship in terms of these two unknowns as follows:

$$\begin{bmatrix} X_n \\ Y_n^* \end{bmatrix} = \begin{bmatrix} H_n & -G_n \\ G_n^* & H_n^* \end{bmatrix} * \begin{bmatrix} A_n \\ B_n^* \end{bmatrix} \tag{11}$$

Which can be rewritten as:

$$\begin{bmatrix} A_n \\ B_n^* \end{bmatrix} = \frac{1}{|H_n|^2 + |G_n|^2} * \begin{bmatrix} H_n^* & G_n \\ -G_n^* & H_n \end{bmatrix} * \begin{bmatrix} X_n \\ Y_n^* \end{bmatrix} \tag{12}$$

Signals A_n and B_n can be determined using equation (12). It should be noted, that the process just described is not the only way to implement STC. Other methods can also be implemented in accordance with the systems and methods described herein. Importantly, however, by adding time diversity, such as described in the preceding equations, to the space diversity already achieved by using base stations 1112 and 1114 to communicate with device 1116 simultaneously, the BER can be reduced even further to recover even more bandwidth.

An example transmitter 2900 configured to communicate using STC in accordance with the systems and methods described herein is illustrated in figure 29. Transmitter 2900 includes a block storage device 2902, a serial-to-parallel converter 2904, encoder 2906, and antenna 2908. Block storage device 2902 is included in transmitter 2900 because a 1 block delay is necessary to implement the coding illustrated in figure 28. This is because transmitter 2804 first transmits $-b_n$* *(n = N-1 to 0)*. But b_n is the second block, so if transmitter 2900 is going to transmit $-b_n$* first, it must store two blocks, e.g., a_n and b_n, and then generate block 2814a and 2814b (see figure 28).

Serial-to-parallel converter 2904 generates parallel bit streams from the bits of blocks a_n and b_n. Encoder 2906 then encodes the bit streams as required, e.g., encoder 2906 can generate $-b_n$* and a_n* (see blocks 2814a and 2814b in figure 28). The encoded blocks are then combined into a single transmit signal as described above and transmitted via antenna 2908.

Transmitter 2900 preferably uses TDM to transmit messages to receiver 2806. An alternative transmitter 3000 embodiment that uses FDM is illustrated in figure 30. Transmitter 3000 also includes block storage device 3002, a serial-to-parallel converter 3004, encoder 3006, and antenna 3008, which are configured to perform in the same manner as the corresponding components in transmitter 2900. But in addition, transmitter 3000 includes IFFTs 3010 to take the IFFT of the blocks generated by encoder 2906. Thus, transmitter 3000 transmits $-B_n$* and A_n* as opposed to $-b_n$* and a_n*, which provides space, frequency, and time diversity.

Figure 31 illustrates an alternative system 3100 that also uses FDM but that eliminates the 1 block delay associated with transmitters 2900 and 3000. In system 3100, transmitter 3102 transmits over channel 3112 to receiver 3116. Transmitter 3106 transmits over channel 3114 to receiver 3116. As with transmitters 2802 and 2804, transmitters

3102 and 3106 implement an encoding scheme designed to recover bandwidth in system 3100. In system 3100, however, the coordinated encoding occurs at the symbol level instead of the block level.

Thus, for example, transmitter 3102 can transmit block 3104 comprising symbols a_0, a_1, a_2, and a_3. In which case, transmitter 3106 will transmit a block 3108 comprising symbols $-a_1^*$, a_0^* $-a_3^*$, and a_2^*. As can be seen, this is the same encoding scheme used by transmitters 2802 and 2804, but implemented at the symbol level instead of the block level. As such, there is no need to delay one block before transmitting. An IFFT of each block 3104 and 3108 can then be taken and transmitted using FDM. An IFFT 3110 of block 3104 is shown in figure 31 for purposes of illustration.

Channels 3112 and 3114 can be described by H_n and G_n, respectively. Thus, in receiver 3116 the following symbols will be formed:

$$(A_0 * H_0) - (A_1^* * G_0)$$

$$(A_1 * H_1) + (A_0^* * G_1)$$

$$(A_2 * H_2) - (A_3^* * G_2)$$

$$(A_3 * H_3) + (A_2^* * G_3).$$

In time, each symbol a_n $(n = 0 \text{ to } 3)$ occupies a slightly different time location. In frequency, each symbol A_n $(n = 0 \text{ to } 3)$ occupies a slightly different frequency. Thus, each symbol A_n is transmitted over a slightly different channel, i.e., $H_n (n = 0 \text{ to } 3)$ or G_n $(n = 0 \text{ to } 3)$, which results in the combinations above.

As can be seen, the symbol combinations formed in the receiver are of the same form as equations (5) and (6) and, therefore, can be solved in the same manner, but without the one block delay.

In order to implement STC or Space Frequency Coding (SFC) diversity as described above, bases stations 1112 and 1114 must be able to coordinate encoding of the symbols that are simultaneously sent to a particular device, such as device 1116 or 1118. Fortunately, base stations 1112 and 1114 are preferably interfaced with a common network interface server. For example, in a LAN, base stations 1112 and 1114 (which would actually be service access points in the case of a LAN) are interfaced with a common network interface server that connects the LAN to a larger network such as a Public Switched Telephone Network (PSTN). Similarly, in a wireless WAN, base stations 1112 and 1114 are typically interfaced with a common base station control center or mobile switching center. Thus, coordination of the encoding can be enabled via the common connection with the network interface server. Bases station 1112 and 1114 can then be configured to share information through this common connection related to communications with devices at the edge of cells 1104 and 1106. The sharing of information, in turn, allows time or frequency diversity coding as described above.

It should be noted that other forms of diversity, such as polarization diversity or delay diversity, can also be combined with the spatial diversity in a communication system designed in accordance with the systems and methods described herein. The goal being to combine alternative forms of diversity with the spatial diversity in order to recover larger amounts of bandwidth. It should also be noted, that the systems and methods described can be applied regardless of he number of base stations, devices, and communication cells involved.

Briefly, delay diversity can preferably be achieved in accordance with the systems and methods described herein by cyclical shifting the transmitted blocks. For example, one transmitter can transmit a block comprising A_0, A_1, A_2, and A_3 in that order, while the other transmitter transmits the symbols in the following order A_3, A_0, A_1, and A_2. Therefore, it can be seen that the second transmitter transmits a cyclically shifted version of the block transmitted by the first transmitter. Further, the shifted block can be cyclically shifted by more then one symbol of required by a particular implementation.

9. Equalization

As explained above, when data is received in a receiver configured in accordance with the systems and methods described herein, the data can be received, split into a plurality of bands, filtered, and equalized. Because of the high data rates involved, however, it can be preferable to perform the equalization at a lower data rate than that of the received data.

An exemplary receive chain 3200 that illustrated various stages of a receiver configured in accordance with the systems and methods described herein, including equalizer 3210, is illustrated in figure 32. Thus, as can be seen, a received data stream 3212 is received and split into two parallel data streams 3214 and 3216, respectively, by serial to parallel converter 3202. Each data stream will have a lower data rate than the original data stream 3212. The number of parallel data streams can be equal to the spreading factor used in the transmitter, and the spreading factor can be related to the number of communication bands used. For purposes of simplicity, it is assumed that the spreading factor is 2 and, therefore, there are two parallel data streams 3214 and 3216. As can be seen, one of these channels, channel 3214, will comprise even data bits, while the other will comprise odd data bits.

Even data stream 3214 and odd data stream 3216 can then each be passed to a matched filter 3204 and 3206 respectively.. The output of the matched filters can then be combined into combined data stream 3218 and sent to equalizer 3210, which will attempt equalize combined data stream 3218, thus generating equalized data stream 3220. The goal of equalizer 3210 is to provide as accurate an estimate of the original data as possible. The estimated data 3220 can then be passed to subsequent processing blocks, such as error correction blocks, in order to detect in errors in the estimates. If errors are detected, various remedial measures can be taken. But the goal should be to limit errors to a rate that still results in acceptable performance.

One way to allow equalization at a lower rate is to use a simplified Decision Feedback Equalization (DFE) configuration. The signal 3212 will actually see the sum of 2 or more autocorrelation functions. As a result:

$$Z_n = (A_0 \text{ x } d_n) + (A_1 \text{ x } d_{n-2} + (A_2 \text{ x } d_{n.2}) + (A_1^* \text{ x } d_{n+1}) + (A_2^* \text{ x } d_{n+2}) \qquad (13)$$

Where: A1* and A2* are the complex conjugates of A1 and A2.

The last four terms in equation (13) are what is known as Intra Symbol Interference (ISI) or Intra Chip Interference (ICI). The first term is the data. So a current data sample, or decision, actually depends on the current sample as well as two past samples and two future samples. Obtaining the two past samples should not drive the complexity of the equalizer; however, obtaining two future samples does drive the complexity.

Accordingly, in the embodiment of equalizer 3210 illustrated in figure 33, the two future samples are ignored. In this example, combined data stream 3318 passes through combiner 3302 and into a hard decision estimation block 3304. This is in contrast with most conventional equalizers, which often depend on soft decisions. The hard decision is then fed to registers 3306 and 3308 in order to obtain the past two samples, which are multiplied by the associated amplitude factors Al and A2 and combined in combiner 3314. The output of combiner 3314 is then subtracted from the combined data stream 3218 in combiner 3302.

Thus, equalizer 3210 implements the following:

$$Z_0 = (A0 \times d0) + (Al \times dl) + (A2 \times d2) \quad (14)$$

Which becomes:

$$Z_0 - [(Al \times dl) + (A2 \times d2)] = (A0 \times d0) \quad (15)$$

The output of the equalizer 3210 can then be passed to a Forward Error Correction (FEC) block 3316, which can clean the estimates up even further.

Figure 34 illustrates an example configuration, wherein an equalizer in the form of DFE 3400 is coupled with the FEC block 3316. The FEC block 3316 is then fed back to another equalizer in the form of DFE 3402 in order to generate a full estimation of the data. An example implementation of DFE 3400 is illustrated in figure 35. The DFE 3402 can be the same or similar. Again, combined data stream 3218 is passed through a hard decision estimation block 3504, which is passed to FEC block 3316. FEC block 3316 then supplies the past two estimates and the future two estimates, which are cleaner estimates. In the embodiment of figure 35, the current sample is re-estimated by block 3504 and used in the lower portion of DFE 3402. This iterative process significantly reduces the ISI or ICI.

Thus, the iterative equalization illustrated above includes one extra iteration; however, it is very simple to implement, especially in an Ultra-wideband system, e.g., approximately 1 GHz. This is in part due to the use of hard decision estimator block 3504. Since the output of hard decision estimator block 3504 is either a 1 or a -1, no multiplication is necessary. This allows for the elimination of the traditional multipliers and the combiner 3314 may then comprise summers. Further, FEC block 3316 can include parity check, or the like, in order to detect errors. If there are no errors, then there is no reason to feedback the data and perform the iteration.

Figure 36 is a diagram illustrating an example embodiment of a matched filter 3600 that can be used, e.g., in the receive chain of figure 32, The coefficients $g_{n,n}$ are determined from the channel estimations and the spreading code.

Figure 37 is a diagram illustrating an example embodiment of a DFE 3700 that can be used to implement the iterative equalization described above. As can be seen, DFE 3700 can include two switches 3702 and 3704 respectively. During the first iteration, the switches can be in position Y. Thus, only the past samples are used as output form hard decision estimation block 3706. During the second iteration, the switches can be switch to position X and shift register 3708 can be fed by FEC 3316. Thus, DFE 3700 comprises a past decision block on the left side, and a future decision block on the right side.

It should be noted that the switches can be in position X during the first iteration, which will improve results, but if speed is an issue then position Y should be used during the first iteration. Alternatively, switch 3702 can be in position Y during the first iteration, while switch 3704 is in position X.

While embodiments and implementations of the invention have been shown and described, it should be apparent that many more embodiments and implementations are within the scope of the invention. Accordingly, the invention is not to be restricted, except in light of the claims and their equivalents.

CLAIMS

1. An equalizer comprising:

a hard decision estimation block configured to generate a hard decision estimate of an input sample;

a past decision block configured to multiply at least one past decision estimate with an associated coefficient; and

a future decision block configured to multiply at least one future decision estimate with an associated coefficient.

2. The equalizer of claim 1, wherein the hard decision estimate is fed to the past decision block during a first iteration.

3. The equalizer of claim 2, wherein the past decision block is fed by the future decision block during a second iteration.

4. The equalizer of claim 1, wherein the hard decision estimate is fed to a forward error correction block.

5. The equalizer of claim 4, wherein the future decision block is fed by the forward error correction block.

6. The equalizer of claim 1, wherein the output of the past decision block is subtracted from the input of the hard decision estimation block during a first iteration.

7. The equalizer of claim 6, wherein the output of the future decision block is combined with the output of the past decision block, and wherein the combined outputs are subtracted from the input of the hard decision estimation block during a second iteration.

8. A receiver, comprising:

a serial to parallel converter configured to receive a serial data stream and convert it into a plurality of parallel data streams;

a plurality of matched filters coupled with the serial to parallel converter, the plurality of matched filters configured to match each of the parallel data streams;

a combiner configured to combine the matched parallel data streams; and

an equalizer comprising:

a hard decision estimation block configured to generate a hard decision estimate of an input sample;

a past decision block configured to multiply at least one past decision estimate with an associated coefficient; and

a future decision block configured to multiply at least one future decision estimate with an associated coefficient.

9. The receiver of claim 8, wherein the hard decision estimate is fed to the past decision block during a first iteration.

10. The receiver of claim 9, wherein the past decision block is fed by the future decision block during a second iteration.

11. The receiver of claim 8, further comprising a forward error correction block coupled with the equalizer, wherein the hard decision estimate is fed to a forward error correction block.

12. The receiver of claim 11, wherein the future decision block is fed by the forward error correction block.

13. The receiver of claim 8, wherein the output of the past decision block is subtracted from the input of the hard decision estimation block during a first iteration.

14. The receiver of claim 13, wherein the output of the future decision block is combined with the output of the past decision block, and wherein the combined outputs are subtracted from the input of the hard decision estimation block during a second iteration.

15. The receiver of claim 12, wherein the forward error correction block is configured to detect errors in the hard decision estimate produced by the hard decision estimation block, and wherein the forward error correction block is further configured to only feed data back to the future estimation block if errors are detected.

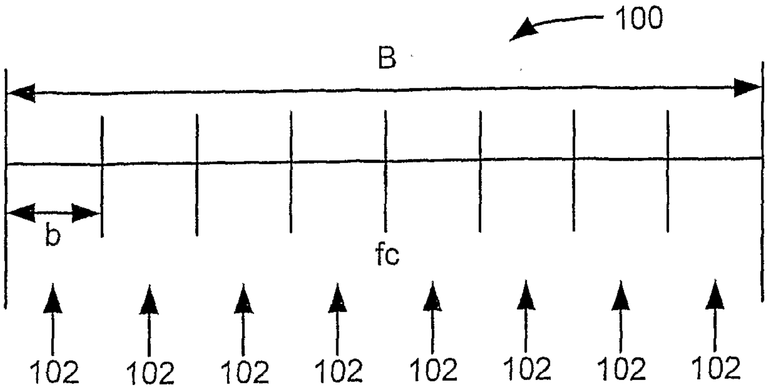

100
B
b
fc
102
102
102
102
102
102
102
102

FIG. 1

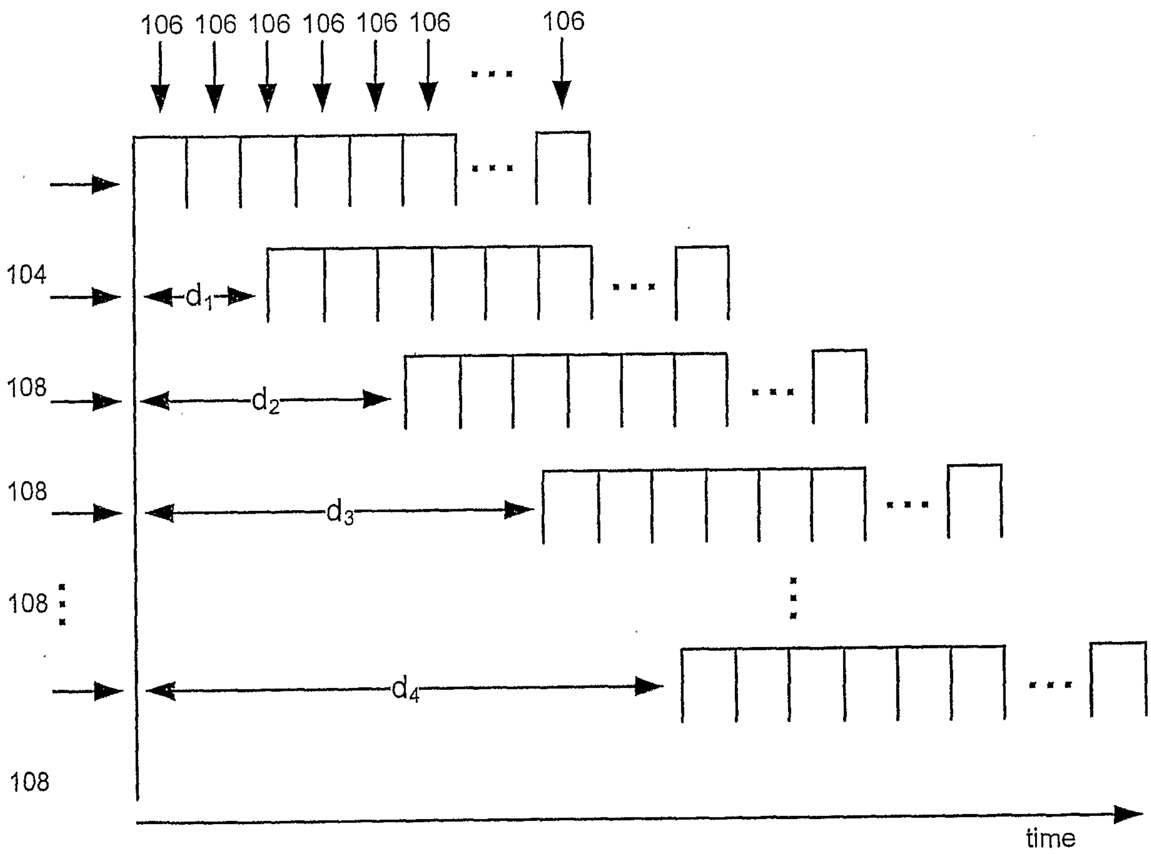

106
106
106
106
106
106
106
104
108
108
108
108
d1
d2
d3
d4
time

FIG. 2

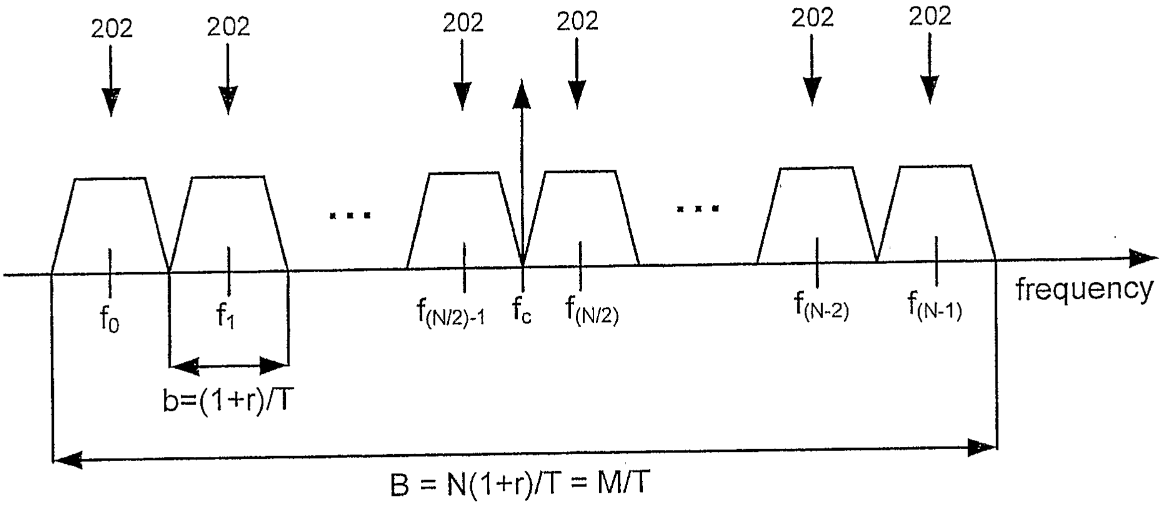

202
202
202
202
202
202
. . .
. . .
frequency
f_0
f_1
$f_{(N/2)-1}$
f_c
$f_{(N/2)}$
$f_{(N-2)}$
$f_{(N-1)}$
b=(1+r)/T
B = N(1+r)/T = M/T

FIG. 3

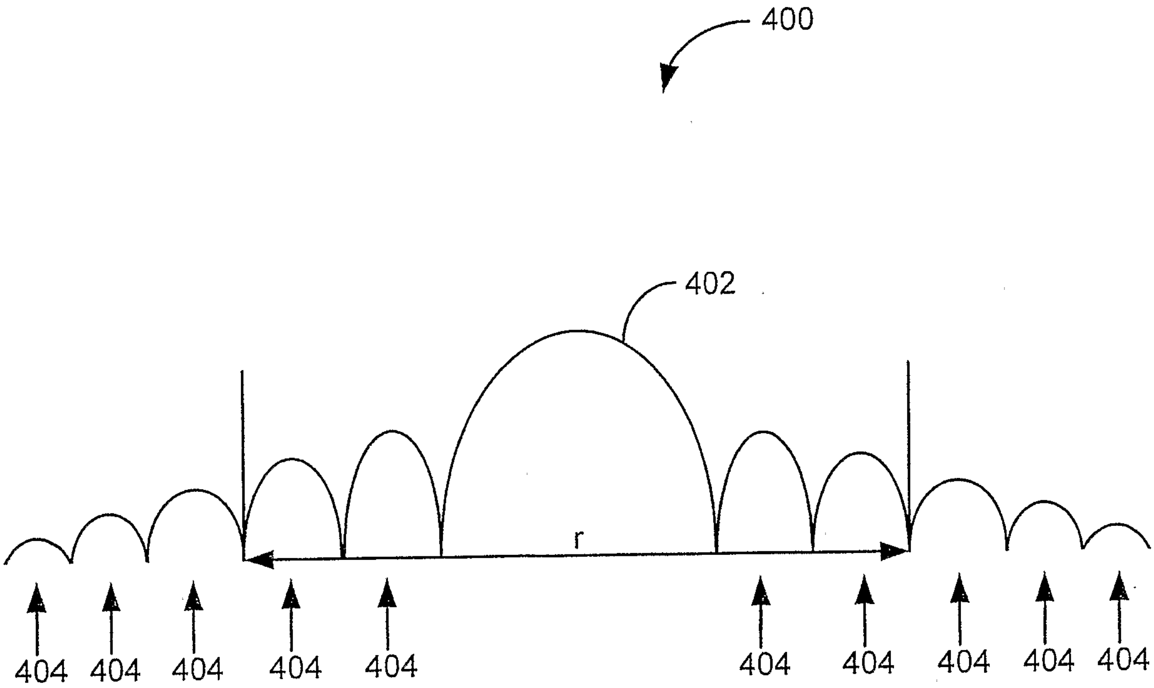


FIG. 4

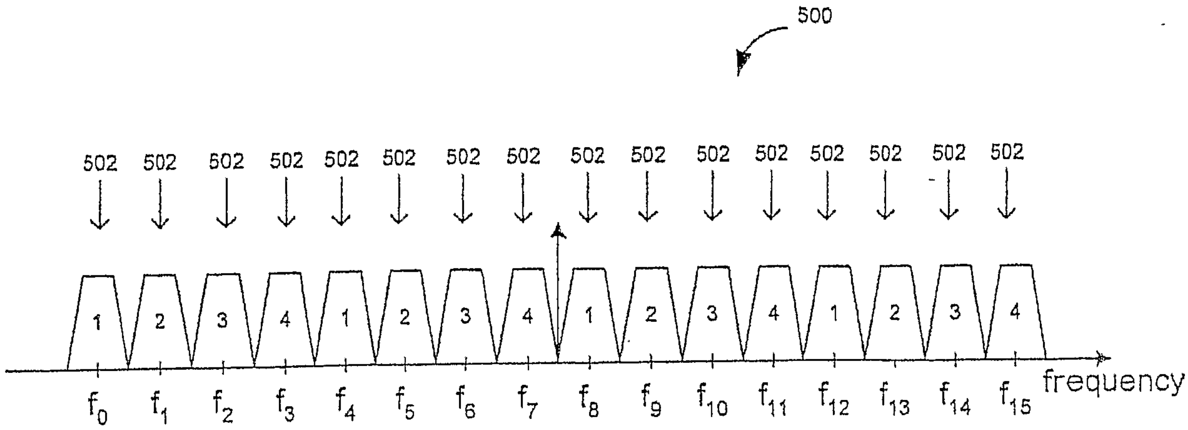


FIGURE 5A

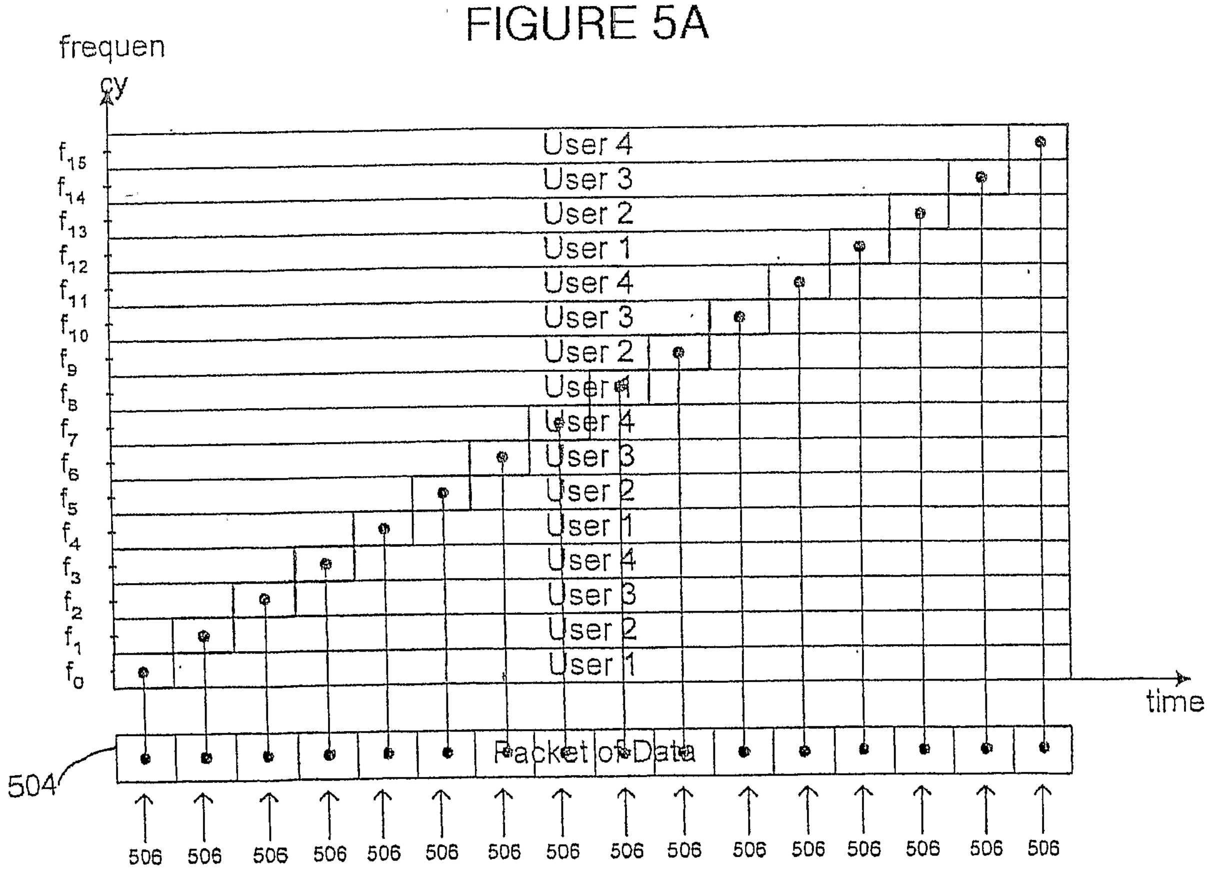


FIGURE 5B

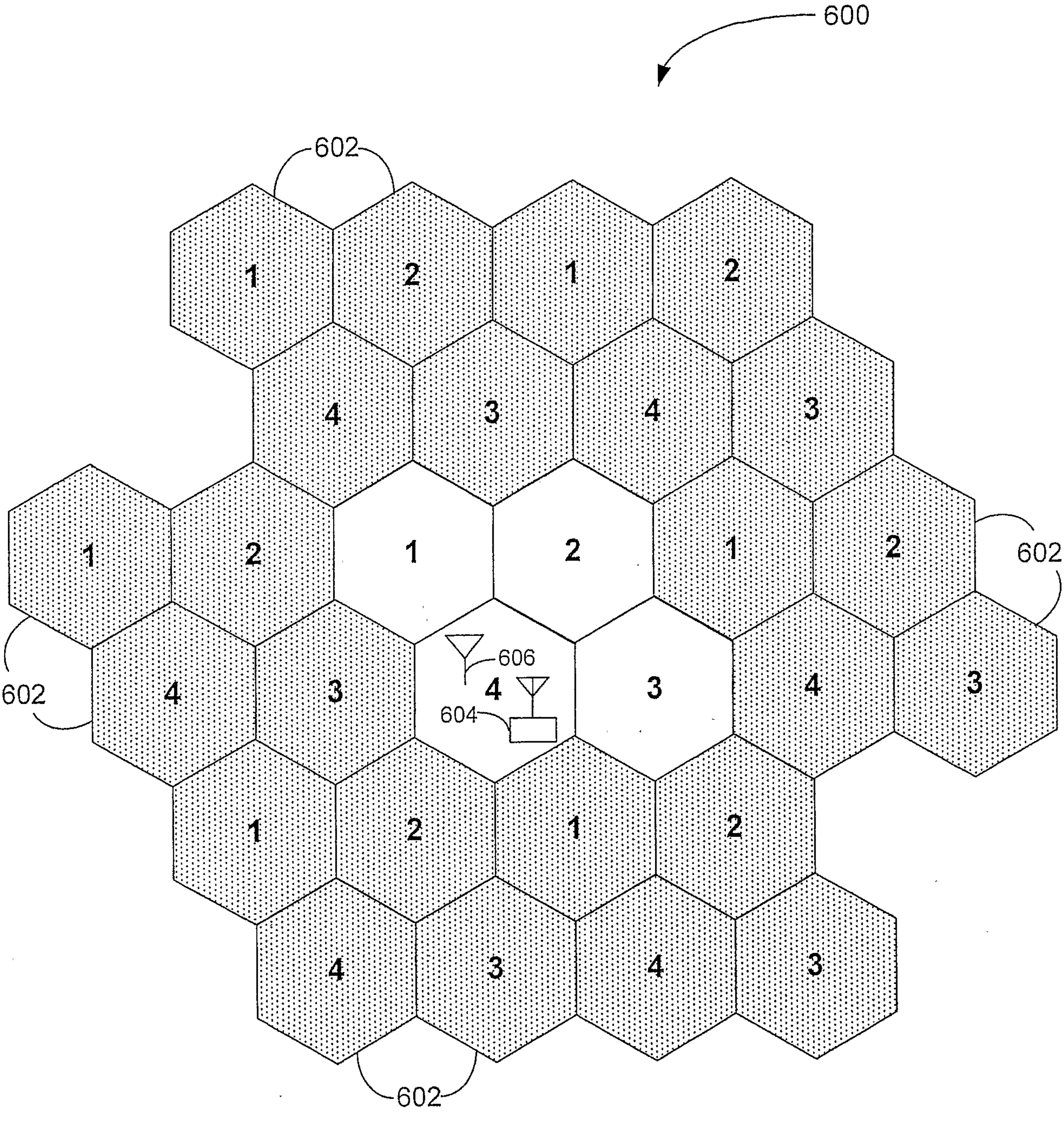


FIG. 6

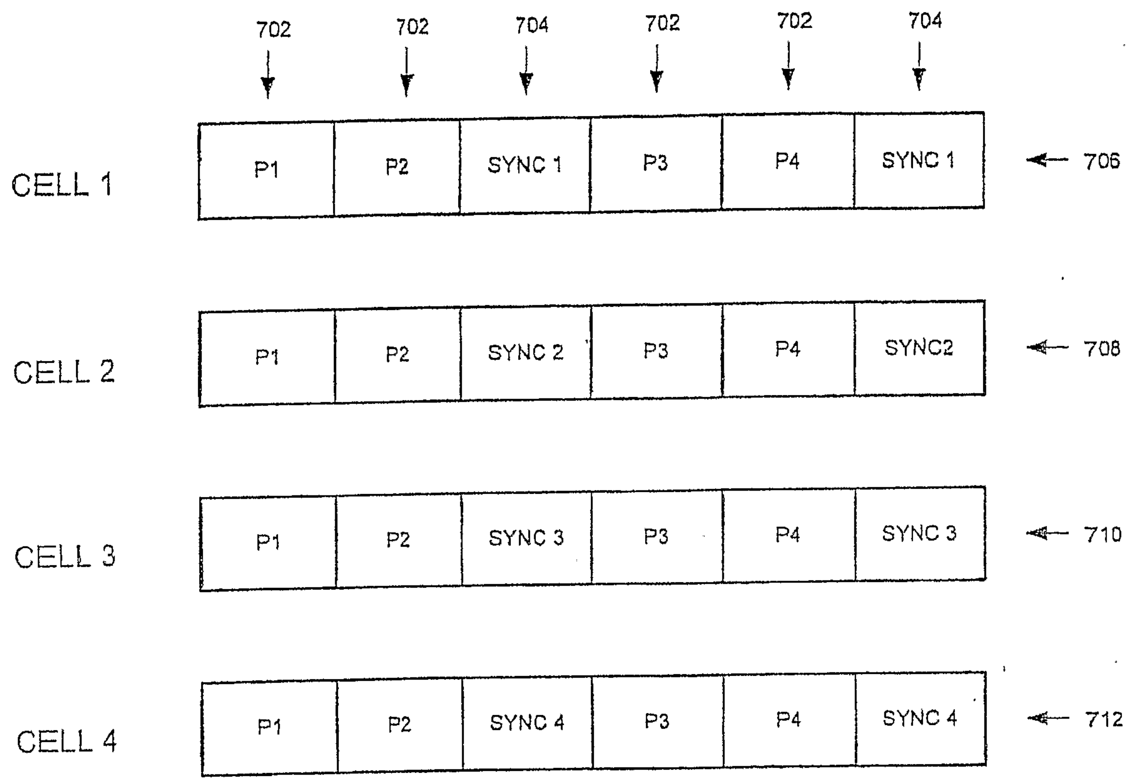

702
702
704
702
702
704
CELL 1
P1
P2
SYNC 1
P3
P4
SYNC 1
706
CELL 2
P1
P2
SYNC 2
P3
P4
SYNC2
708
CELL 3
P1
P2
SYNC 3
P3
P4
SYNC 3
710
CELL 4
P1
P2
SYNC 4
P3
P4
SYNC 4
712

FIGURE 7

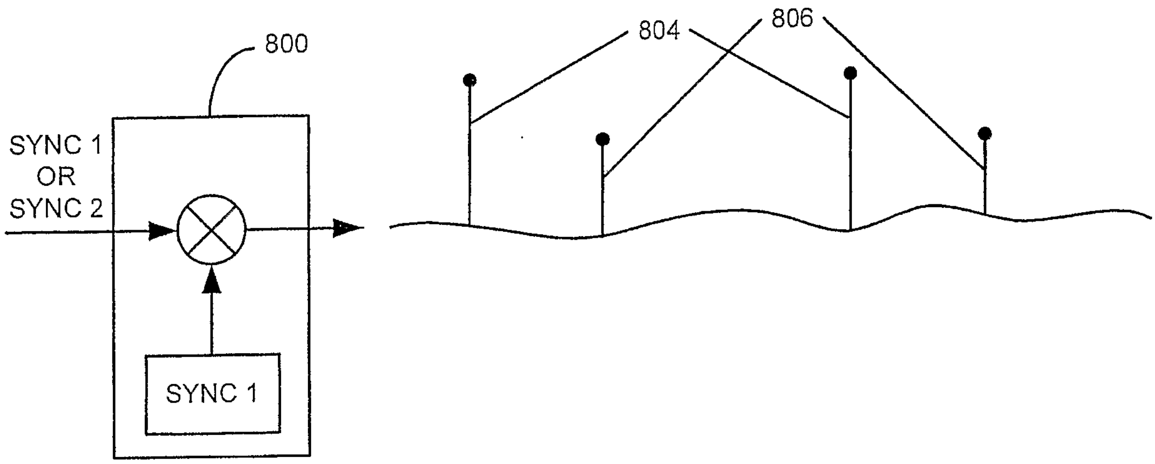


FIG. 8

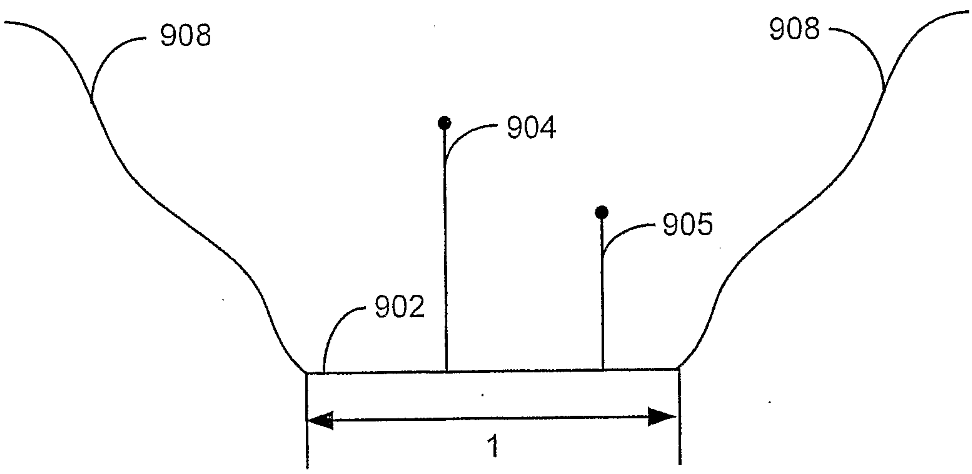


FIG. 9

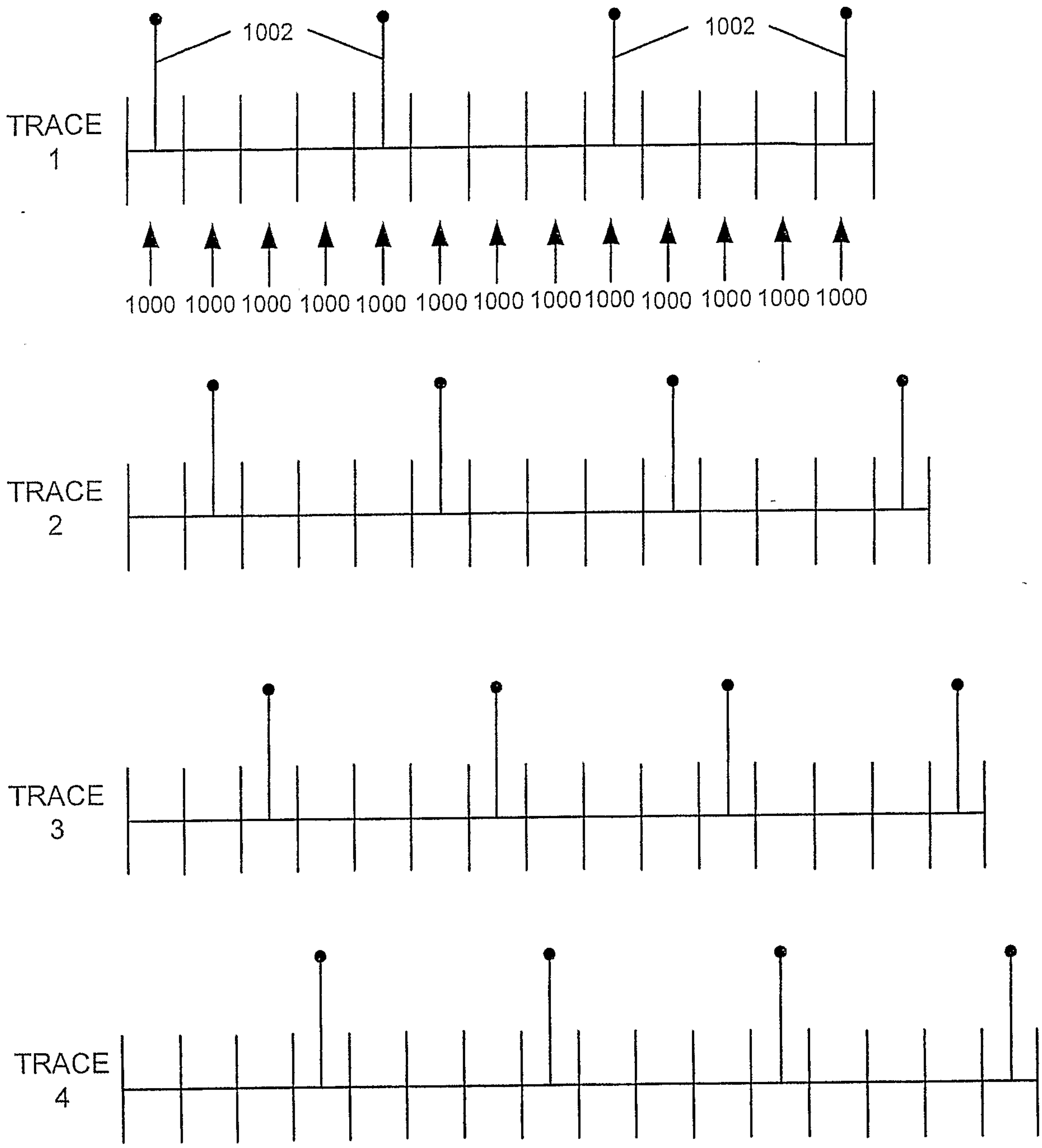

1002
1002
TRACE 1
1000 1000 1000 1000 1000 1000 1000 1000 1000 1000 1000 1000 1000
TRACE 2
TRACE 3
TRACE 4

FIG. 10

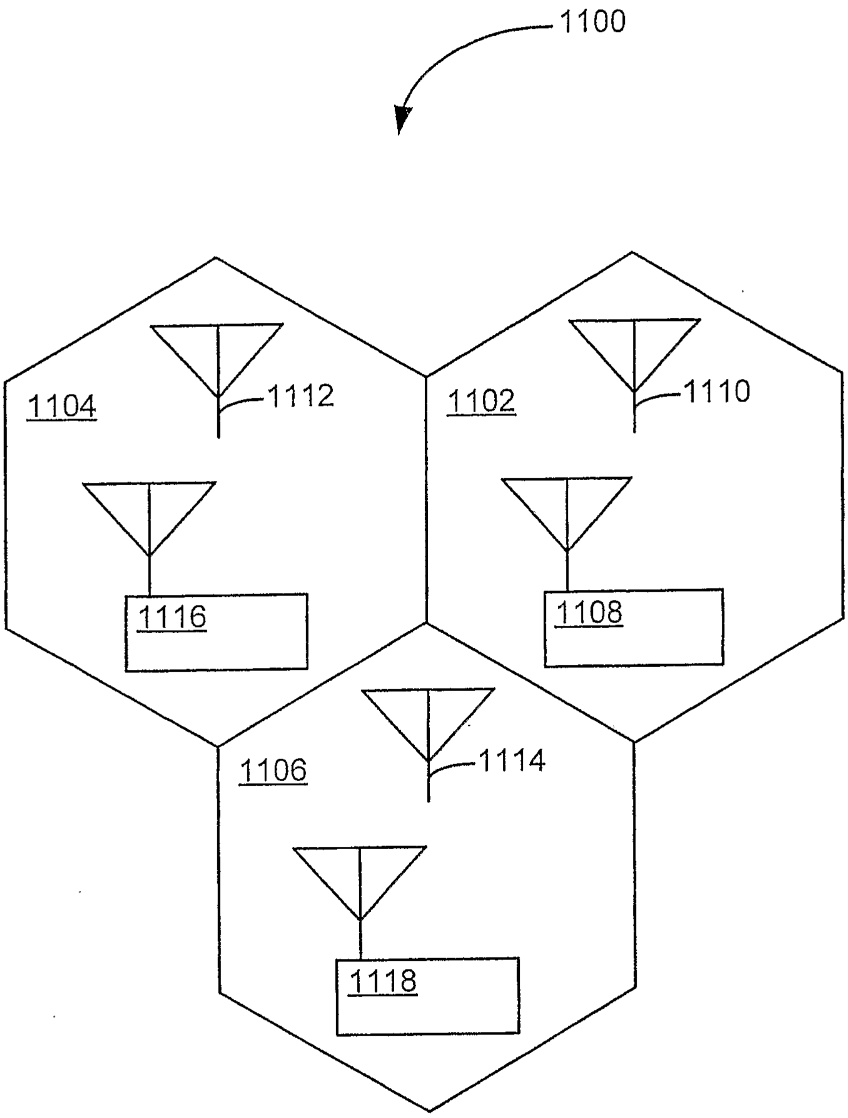


FIG. 11

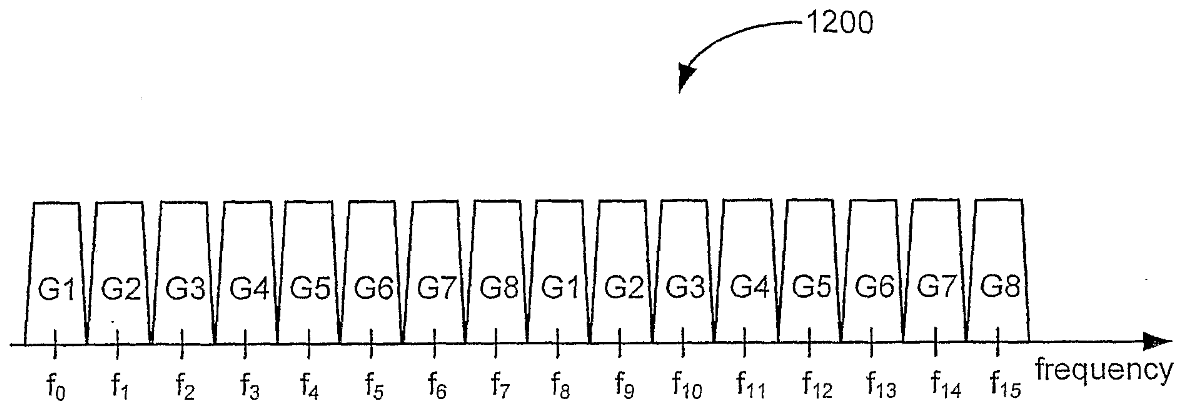


FIG. 12A

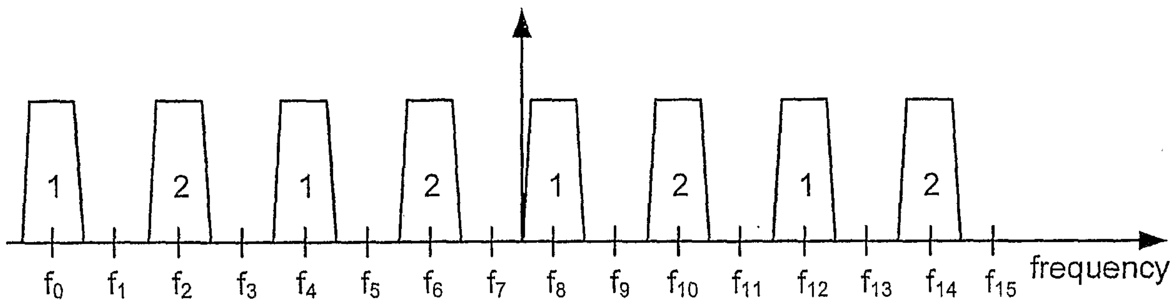


FIG. 12B

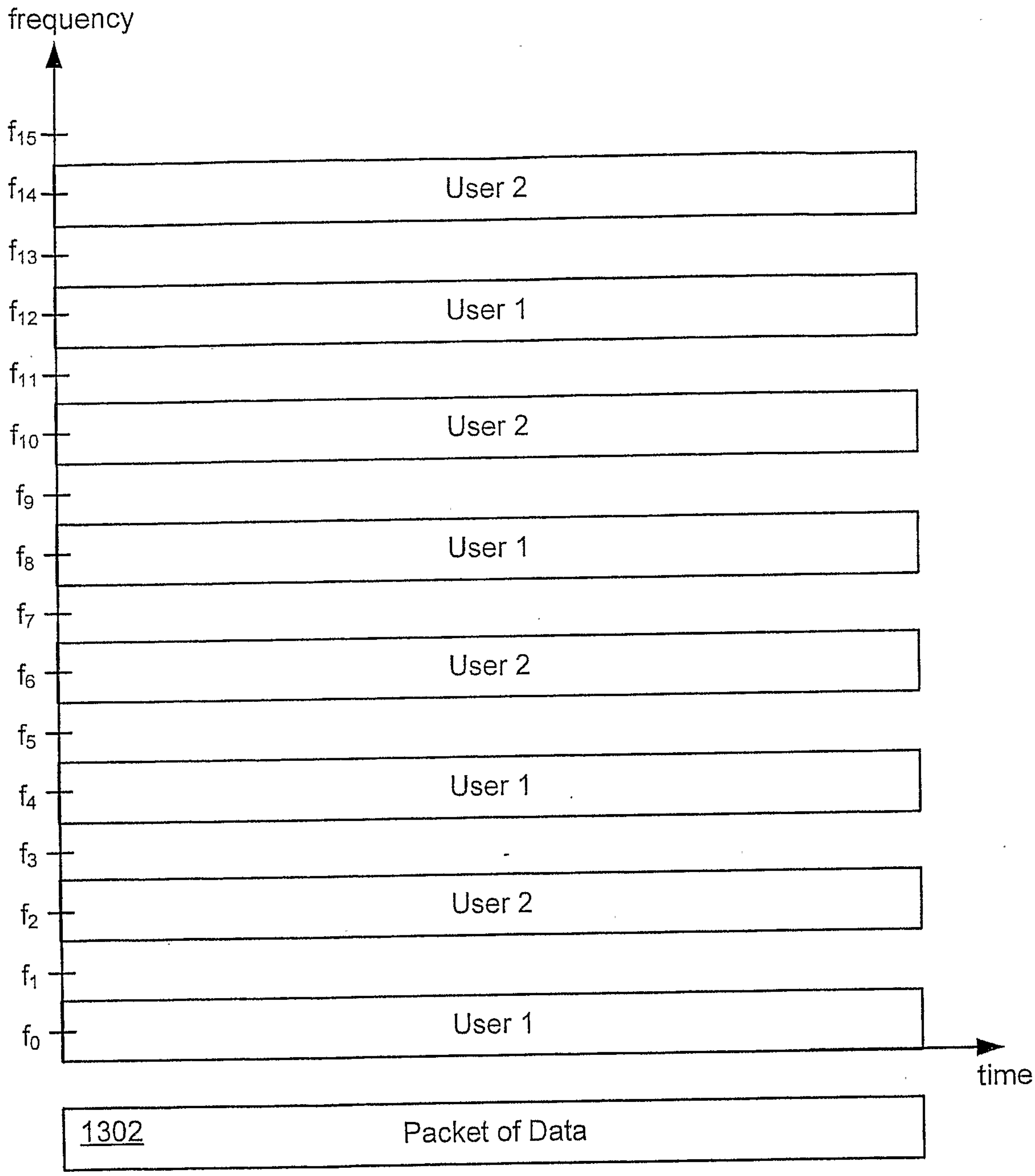

frequency
f15
f14
User 2
f13
f12
User 1
f11
f10
User 2
f9
f8
User 1
f7
f6
User 2
f5
f4
User 1
f3
f2
User 2
f1
f0
User 1
time
1302
Packet of Data

FIG. 13

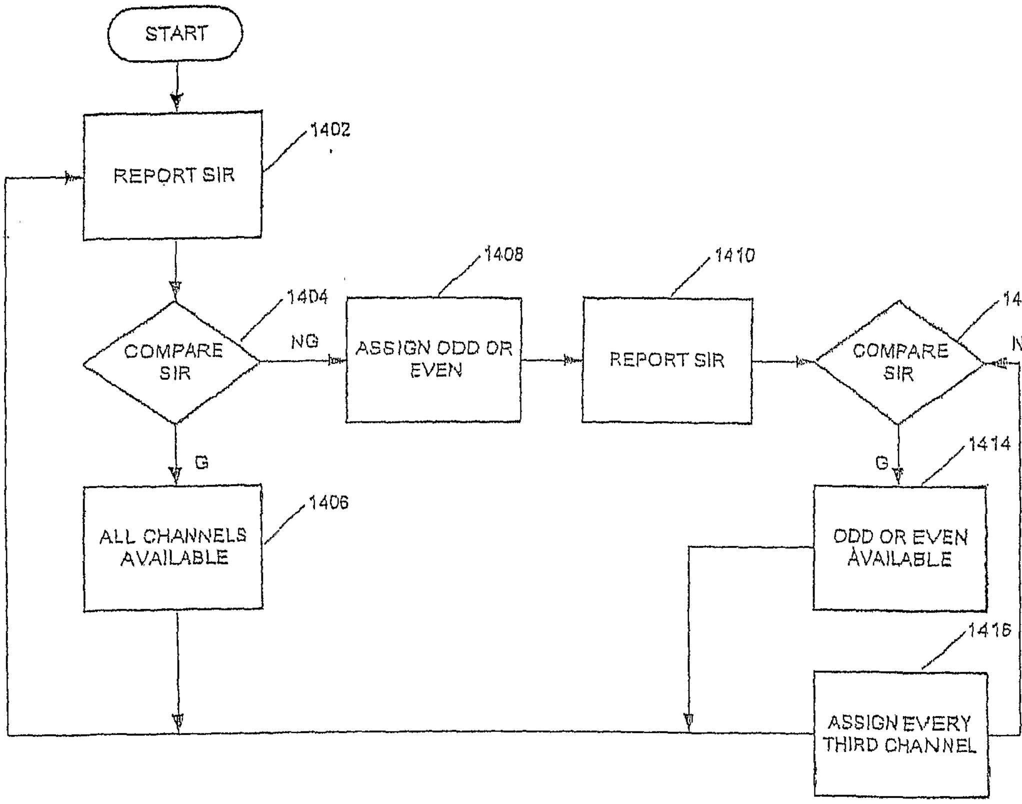

START
REPORT SIR
1402
COMPARE SIR
1404
NG
G
ALL CHANNELS AVAILABLE
1406
ASSIGN ODD OR EVEN
1408
REPORT SIR
1410
COMPARE SIR
1412
NG
G
ODD OR EVEN AVAILABLE
1414
ASSIGN EVERY THIRD CHANNEL
1416

Figure 14

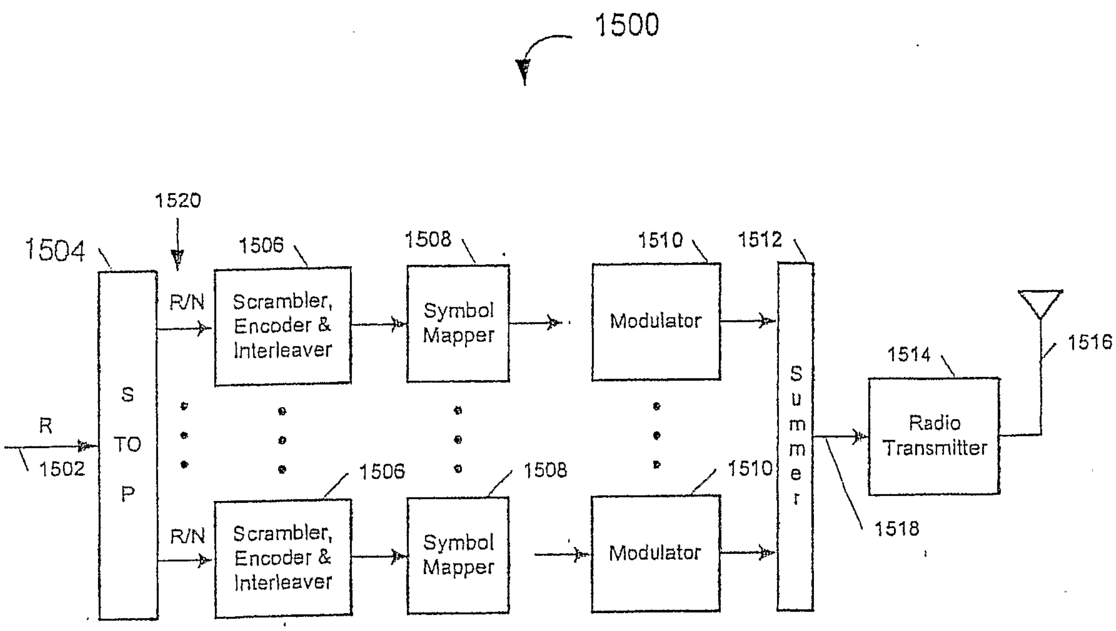

1500
1520
1504
1506
1508
1510
1512
R/N
Scrambler, Encoder & Interleaver
Symbol Mapper
Modulator
1516
1514
S TO P
R
1502
Summer
Radio Transmitter
1518

FIGURE 15

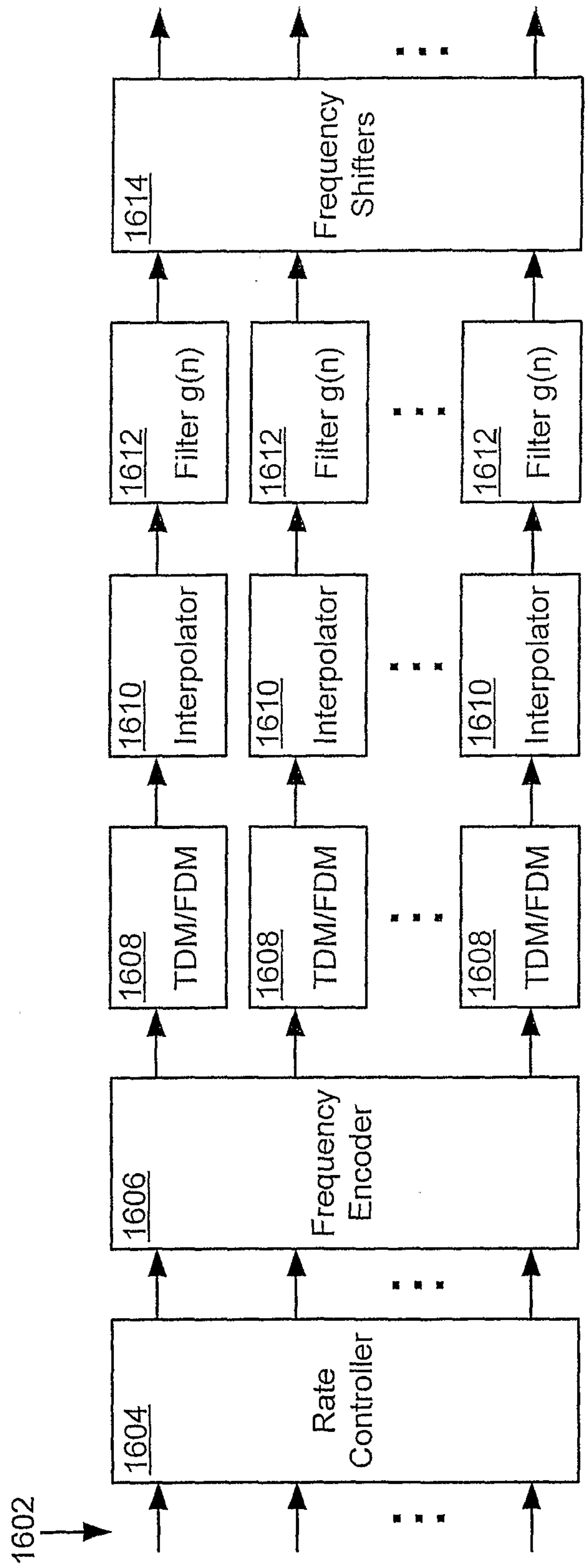

1602
1604
Rate Controller
1606
Frequency Encoder
1608
TDM/FDM
1610
Interpolator
1612
Filter g(n)
1614
Frequency Shifters

FIG. 16

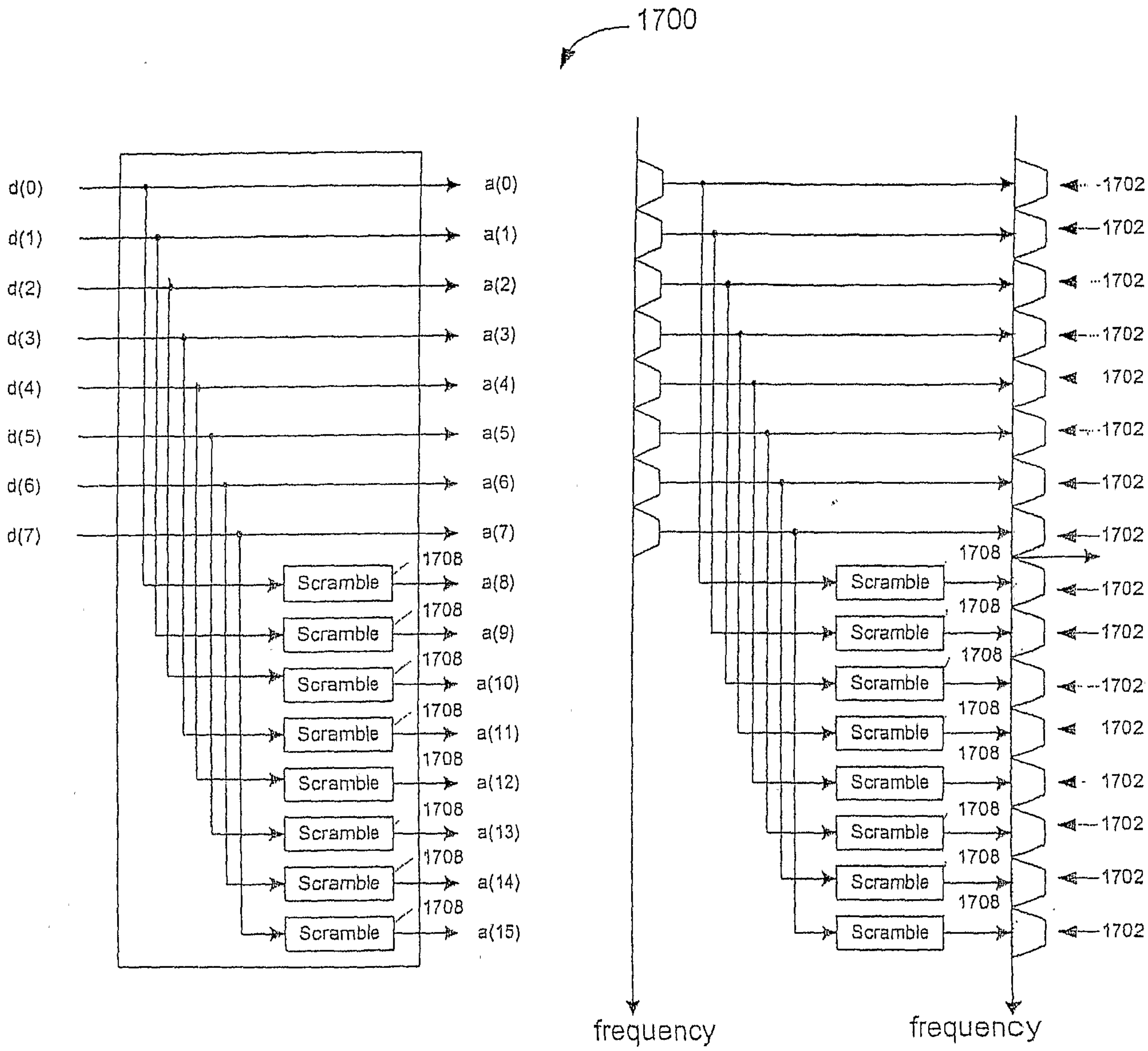
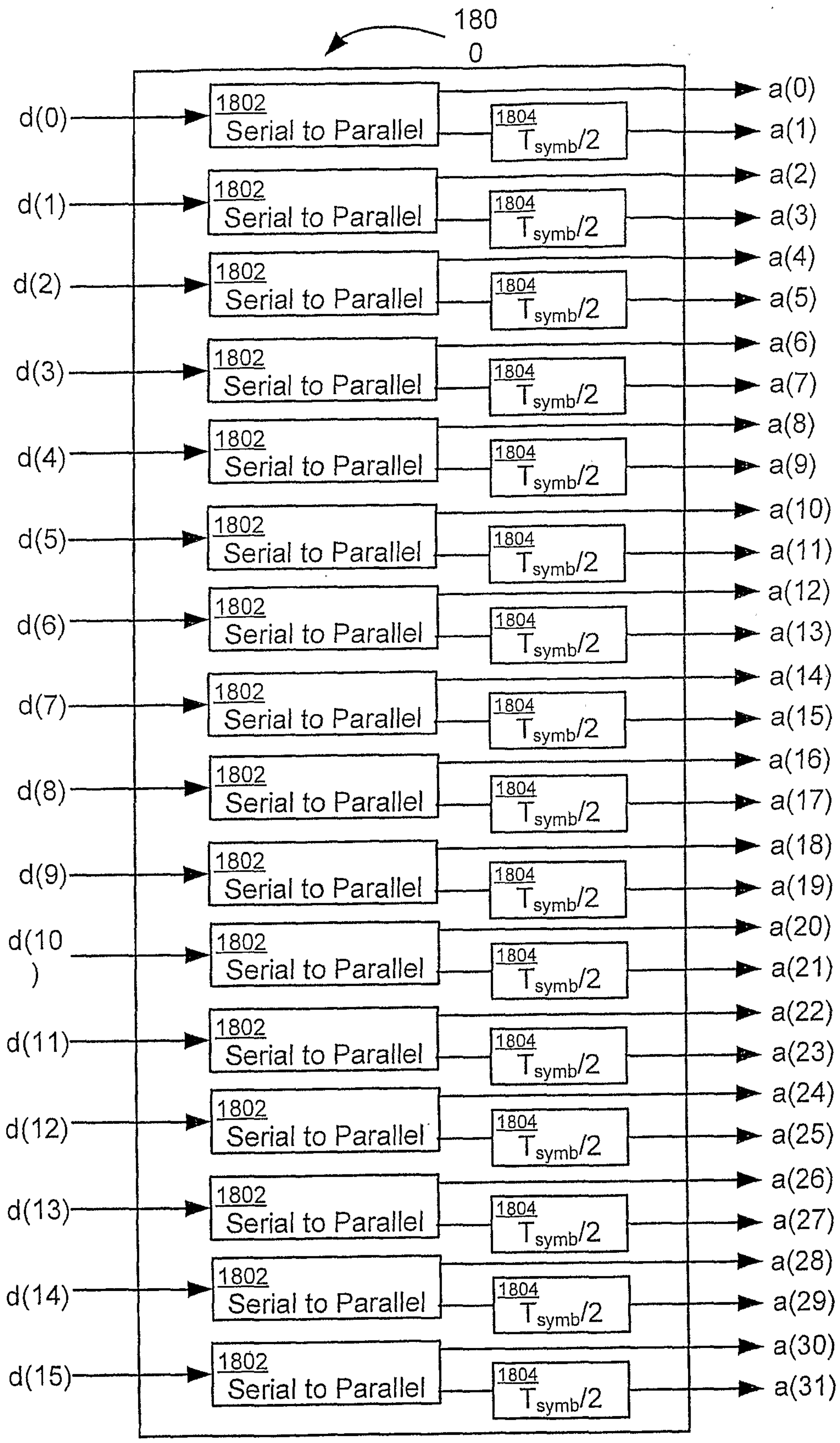

1700
d(0)
d(1)
d(2)
d(3)
d(4)
d(5)
d(6)
d(7)
a(0)
a(1)
a(2)
a(3)
a(4)
a(5)
a(6)
a(7)
a(8)
a(9)
a(10)
a(11)
a(12)
a(13)
a(14)
a(15)
Scramble
1708
1702
frequency
frequency

Figure 17



180
0
d(0)
d(1)
d(2)
d(3)
d(4)
d(5)
d(6)
d(7)
d(8)
d(9)
d(10
)
d(11)
d(12)
d(13)
d(14)
d(15)
1802
Serial to Parallel
1804
Tsymb/2
a(0)
a(1)
a(2)
a(3)
a(4)
a(5)
a(6)
a(7)
a(8)
a(9)
a(10)
a(11)
a(12)
a(13)
a(14)
a(15)
a(16)
a(17)
a(18)
a(19)
a(20)
a(21)
a(22)
a(23)
a(24)
a(25)
a(26)
a(27)
a(28)
a(29)
a(30)
a(31)

FIG. 18

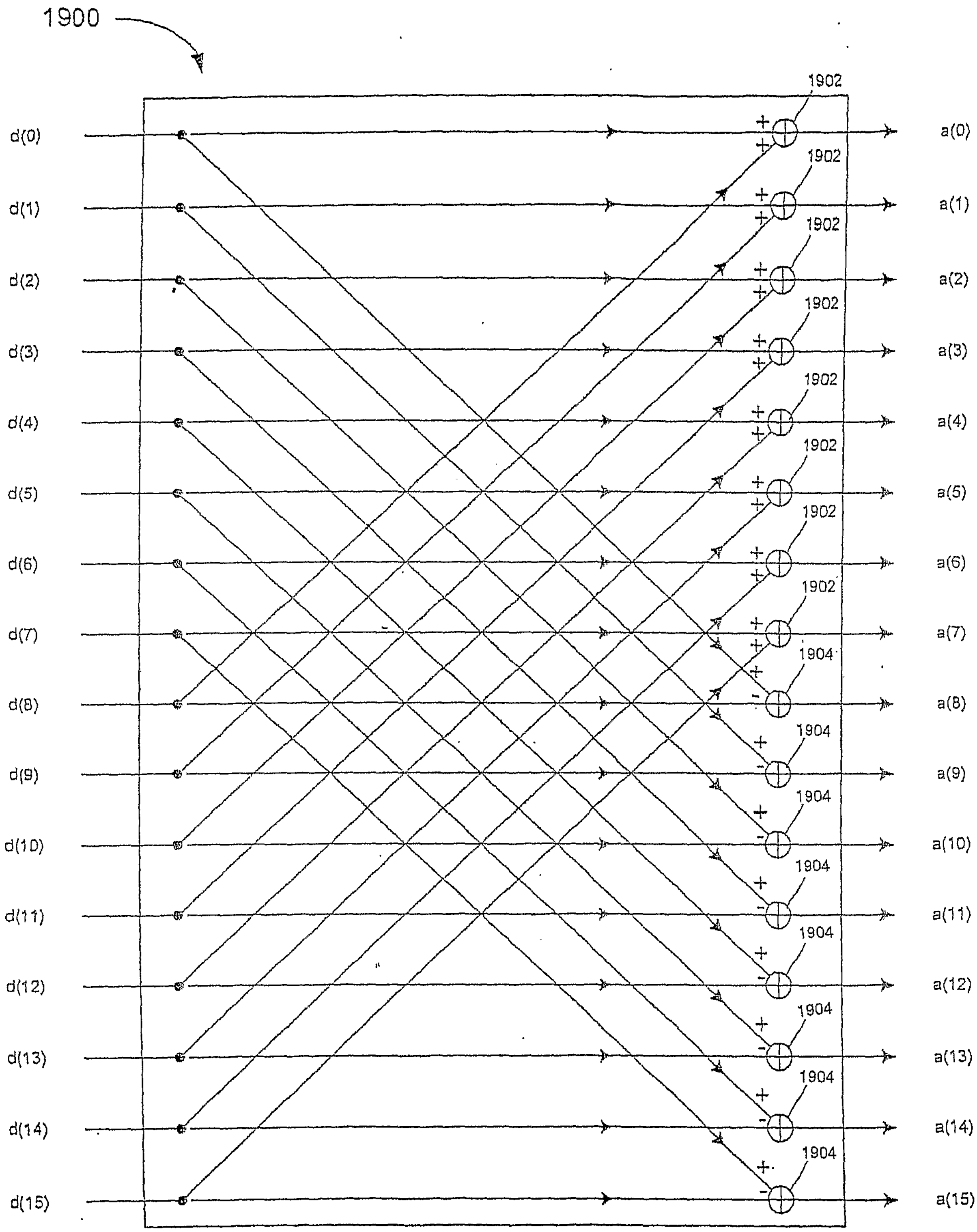

1900
d(0)
d(1)
d(2)
d(3)
d(4)
d(5)
d(6)
d(7)
d(8)
d(9)
d(10)
d(11)
d(12)
d(13)
d(14)
d(15)
1902
1904
a(0)
a(1)
a(2)
a(3)
a(4)
a(5)
a(6)
a(7)
a(8)
a(9)
a(10)
a(11)
a(12)
a(13)
a(14)
a(15)

FIGURE 19

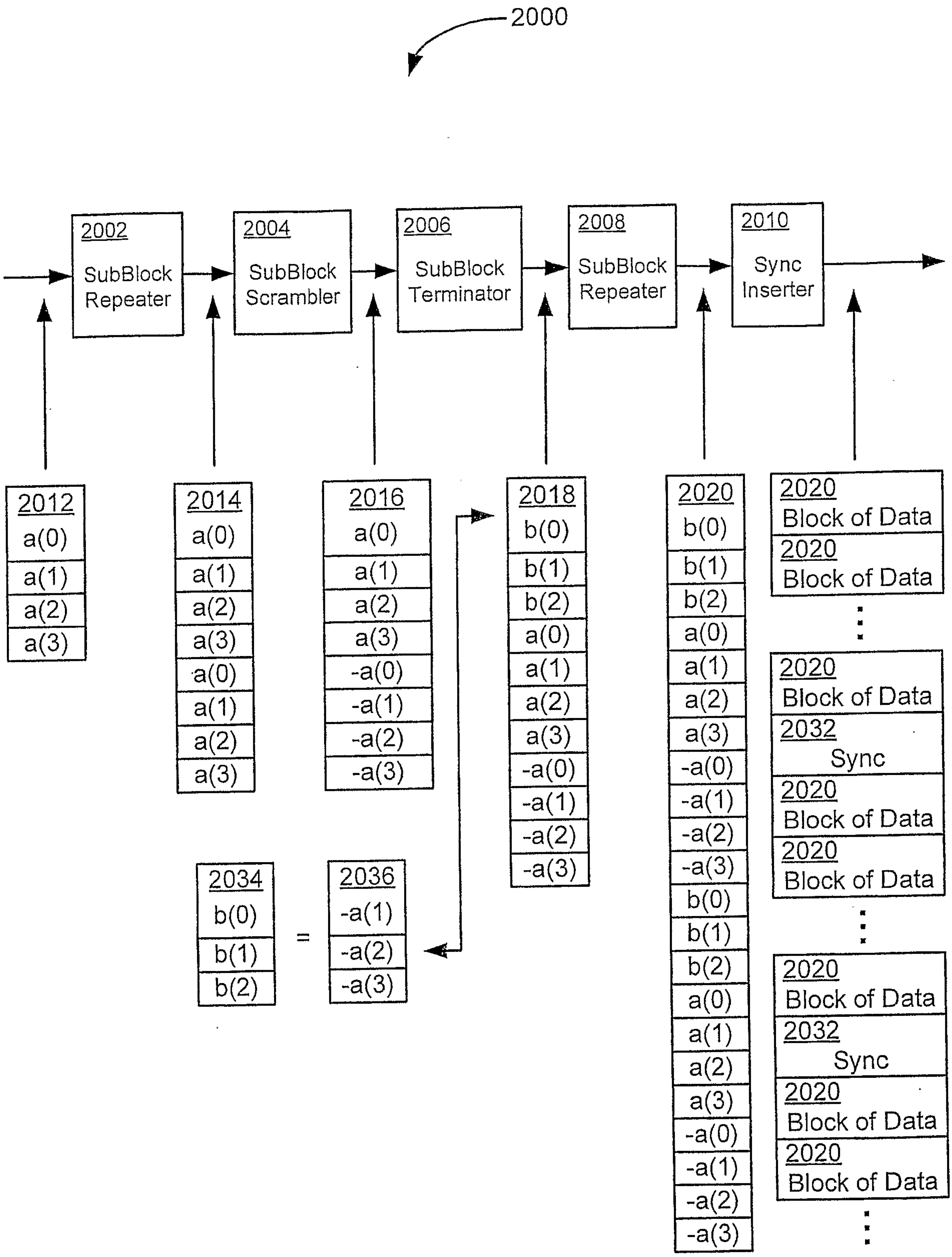

2000
2002
SubBlock Repeater
2004
SubBlock Scrambler
2006
SubBlock Terminator
2008
SubBlock Repeater
2010
Sync Inserter
2012
a(0)
a(1)
a(2)
a(3)
2014
a(0)
a(1)
a(2)
a(3)
a(0)
a(1)
a(2)
a(3)
2016
a(0)
a(1)
a(2)
a(3)
-a(0)
-a(1)
-a(2)
-a(3)
2018
b(0)
b(1)
b(2)
a(0)
a(1)
a(2)
a(3)
-a(0)
-a(1)
-a(2)
-a(3)
2020
b(0)
b(1)
b(2)
a(0)
a(1)
a(2)
a(3)
-a(0)
-a(1)
-a(2)
-a(3)
b(0)
b(1)
b(2)
a(0)
a(1)
a(2)
a(3)
-a(0)
-a(1)
-a(2)
-a(3)
2020 Block of Data
2020 Block of Data
2020 Block of Data
2032 Sync
2020 Block of Data
2020 Block of Data
2020 Block of Data
2032 Sync
2020 Block of Data
2020 Block of Data
2034
b(0)
b(1)
b(2)
=
2036
-a(1)
-a(2)
-a(3)

FIG. 20

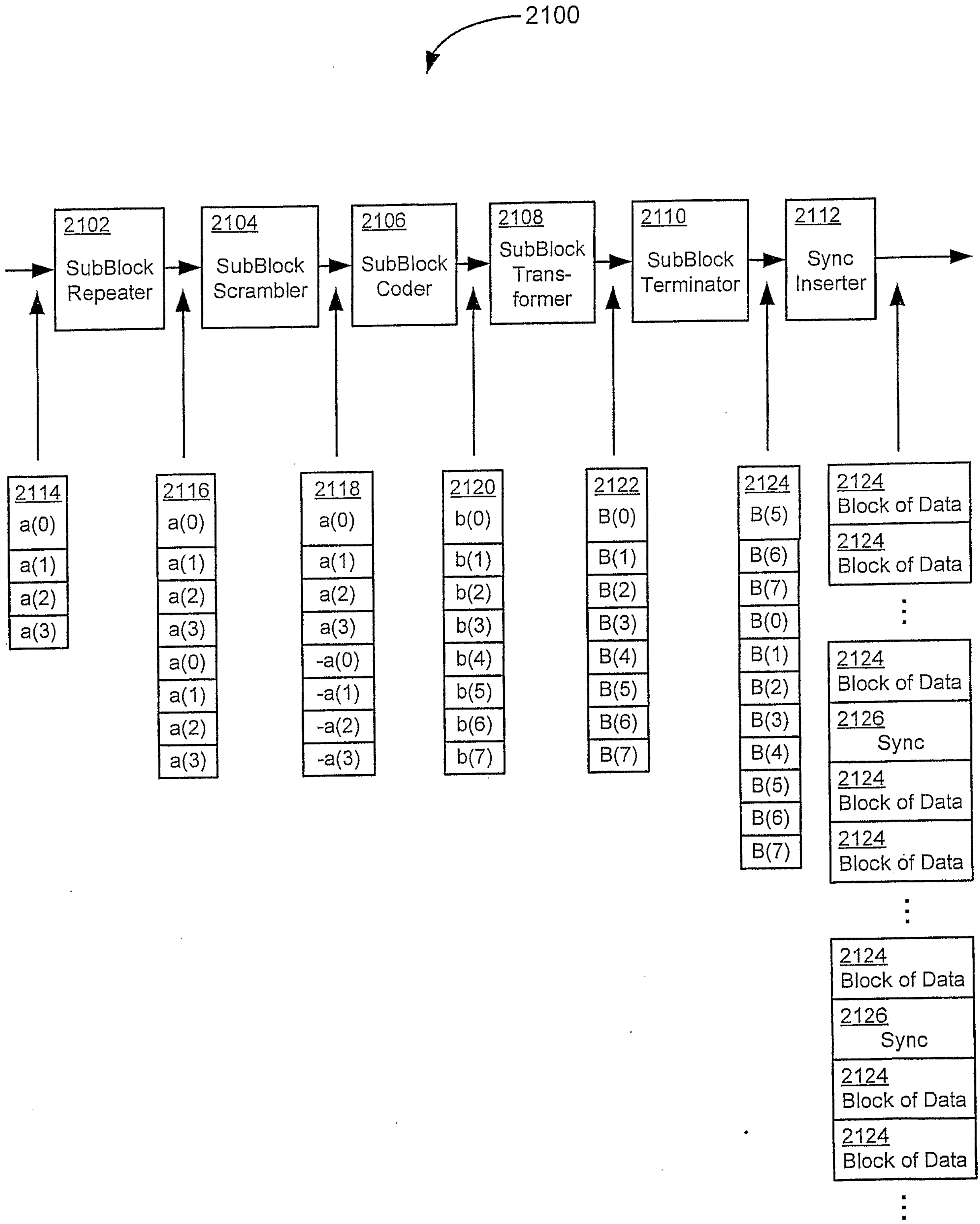

2100
2102
SubBlock Repeater
2104
SubBlock Scrambler
2106
SubBlock Coder
2108
SubBlock Trans-former
2110
SubBlock Terminator
2112
Sync Inserter
2114
a(0)
a(1)
a(2)
a(3)
2116
a(0)
a(1)
a(2)
a(3)
a(0)
a(1)
a(2)
a(3)
2118
a(0)
a(1)
a(2)
a(3)
-a(0)
-a(1)
-a(2)
-a(3)
2120
b(0)
b(1)
b(2)
b(3)
b(4)
b(5)
b(6)
b(7)
2122
B(0)
B(1)
B(2)
B(3)
B(4)
B(5)
B(6)
B(7)
2124
B(5)
B(6)
B(7)
B(0)
B(1)
B(2)
B(3)
B(4)
B(5)
B(6)
B(7)
2124 Block of Data
2124 Block of Data
2124 Block of Data
2126 Sync
2124 Block of Data
2124 Block of Data
2124 Block of Data
2126 Sync
2124 Block of Data
2124 Block of Data

FIG. 21

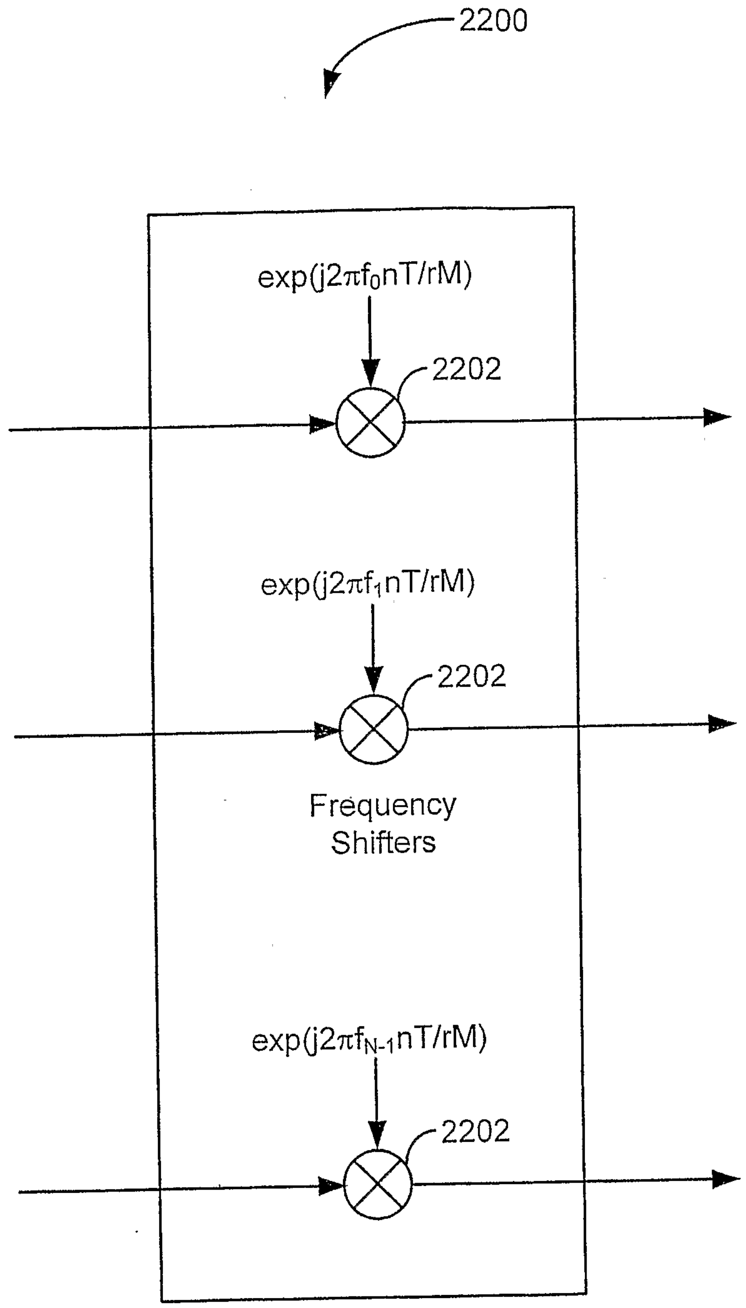

2200
exp(j2πf0nT/rM)
2202
exp(j2πf1nT/rM)
2202
Frequency
Shifters
exp(j2πfN-1nT/rM)
2202

FIG. 22

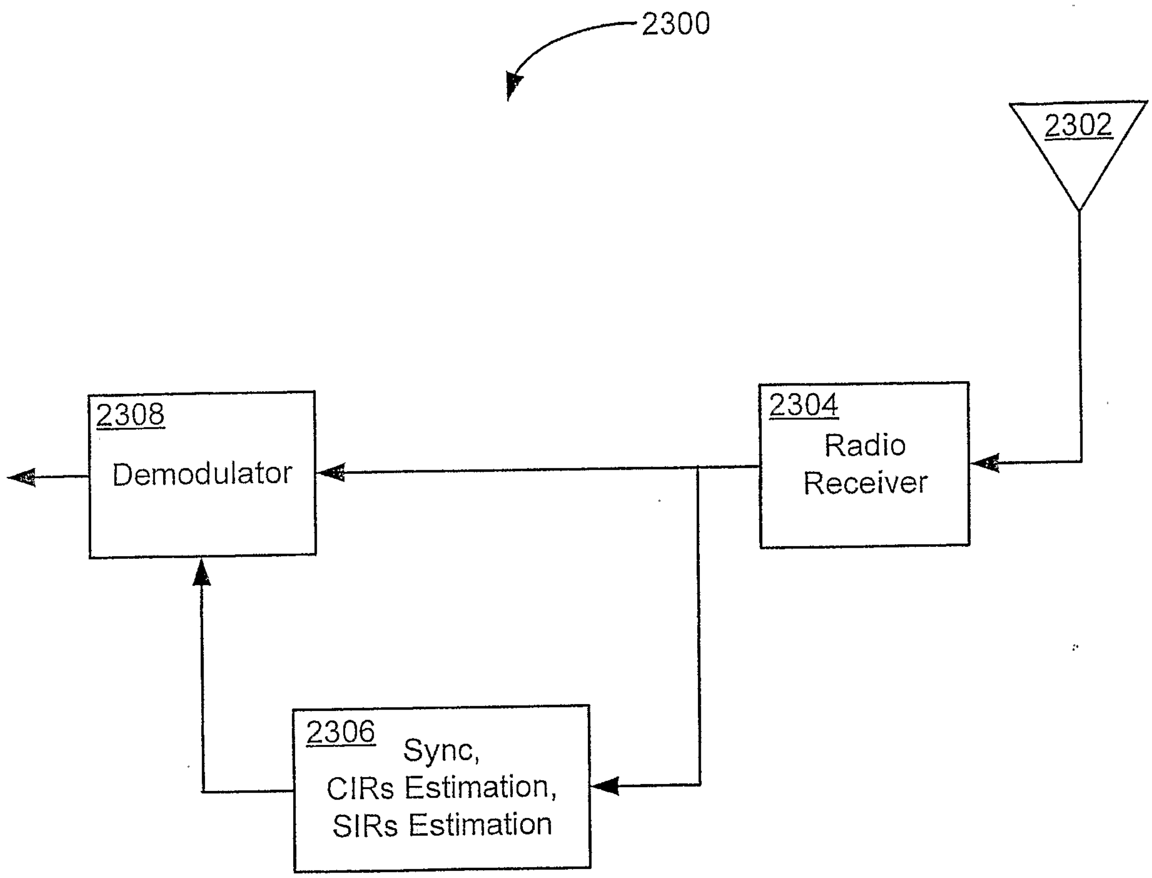

2300
2302
2304
Radio
Receiver
2308
Demodulator
2306
Sync,
CIRs Estimation,
SIRs Estimation

FIG. 23

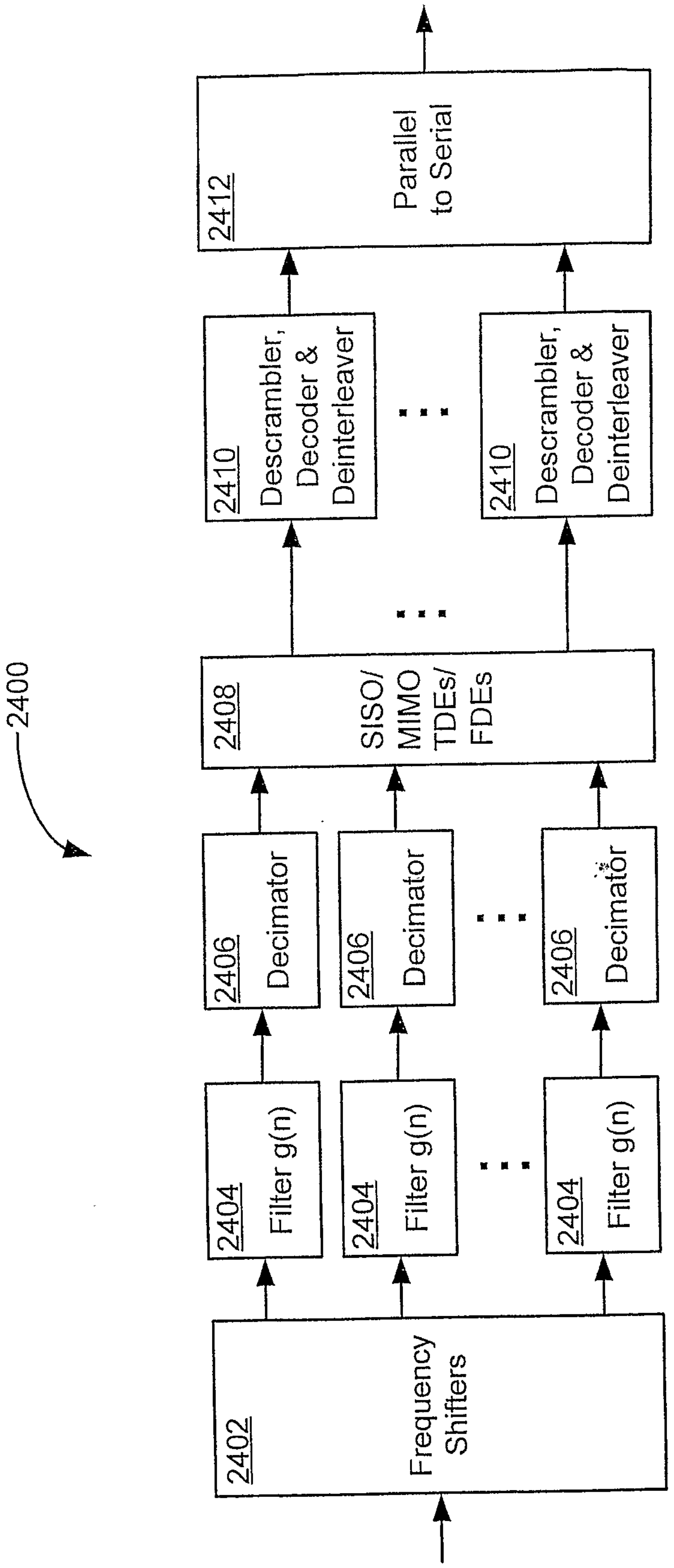

2400
2402
Frequency Shifters
2404
Filter g(n)
2404
Filter g(n)
2404
Filter g(n)
2406
Decimator
2406
Decimator
2406
Decimator
2408
SISO/
MIMO
TDEs/
FDEs
2410
Descrambler,
Decoder &
Deinterleaver
2410
Descrambler,
Decoder &
Deinterleaver
2412
Parallel
to Serial

FIG. 24

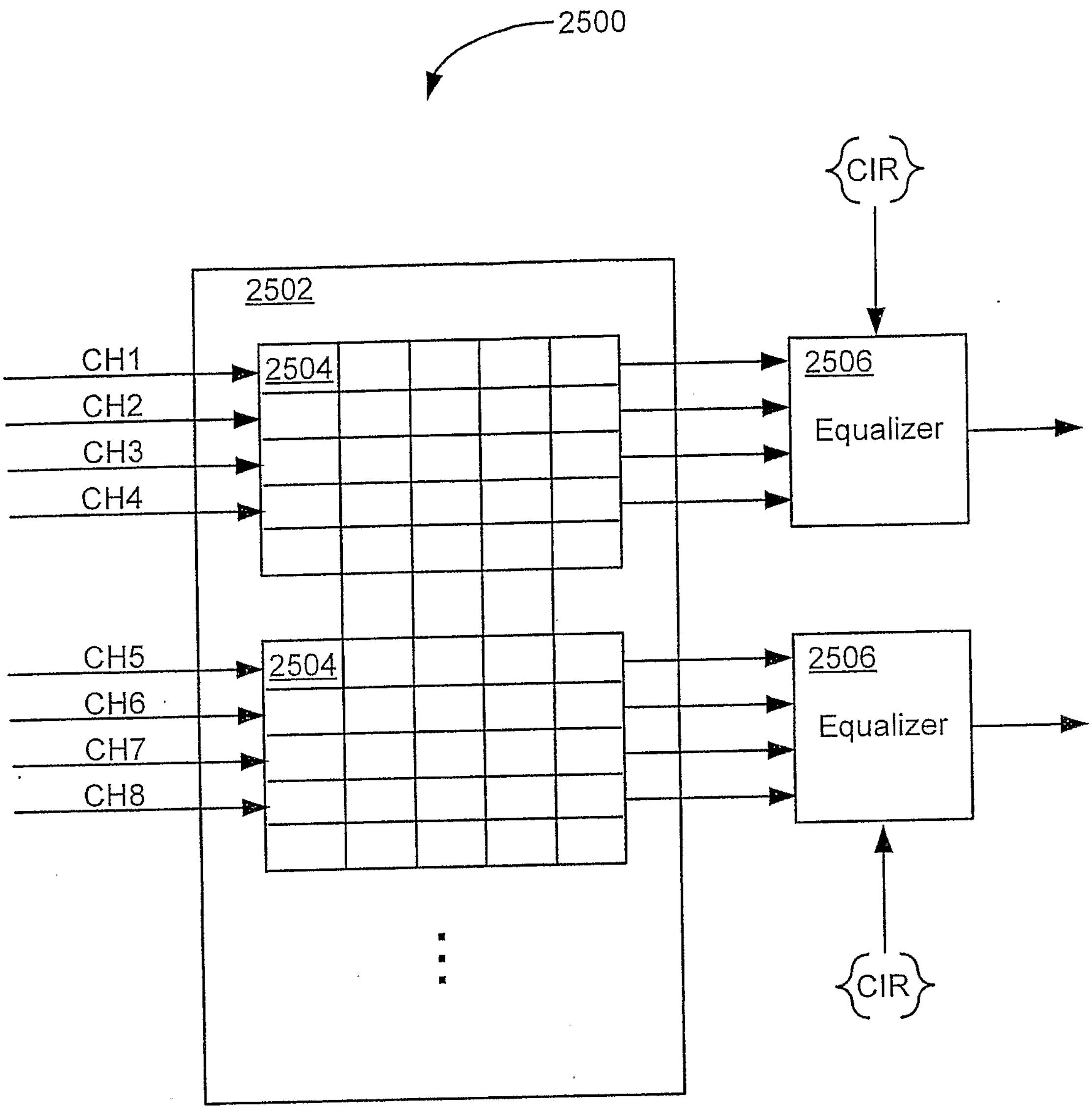

2500
CIR
2502
CH1
CH2
CH3
CH4
2504
2506
Equalizer
CH5
CH6
CH7
CH8
2504
2506
Equalizer
CIR

FIG. 25

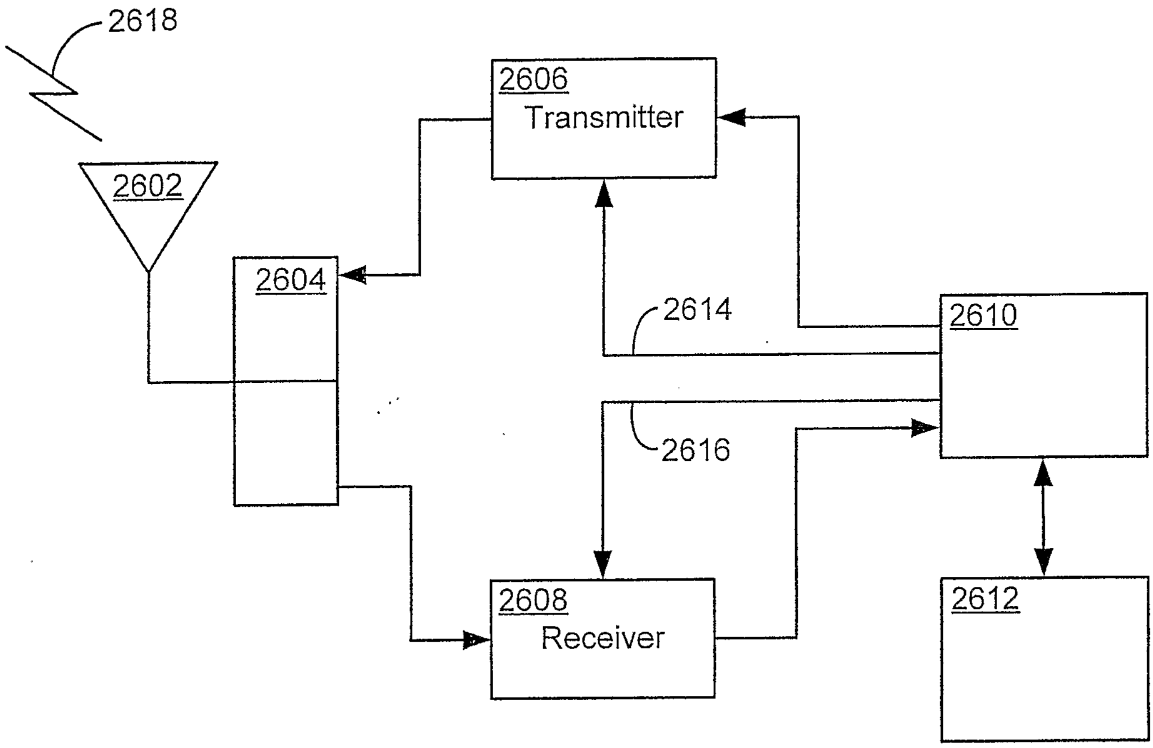

2618
2606
Transmitter
2602
2604
2614
2610
2616
2608
Receiver
2612

FIG. 26

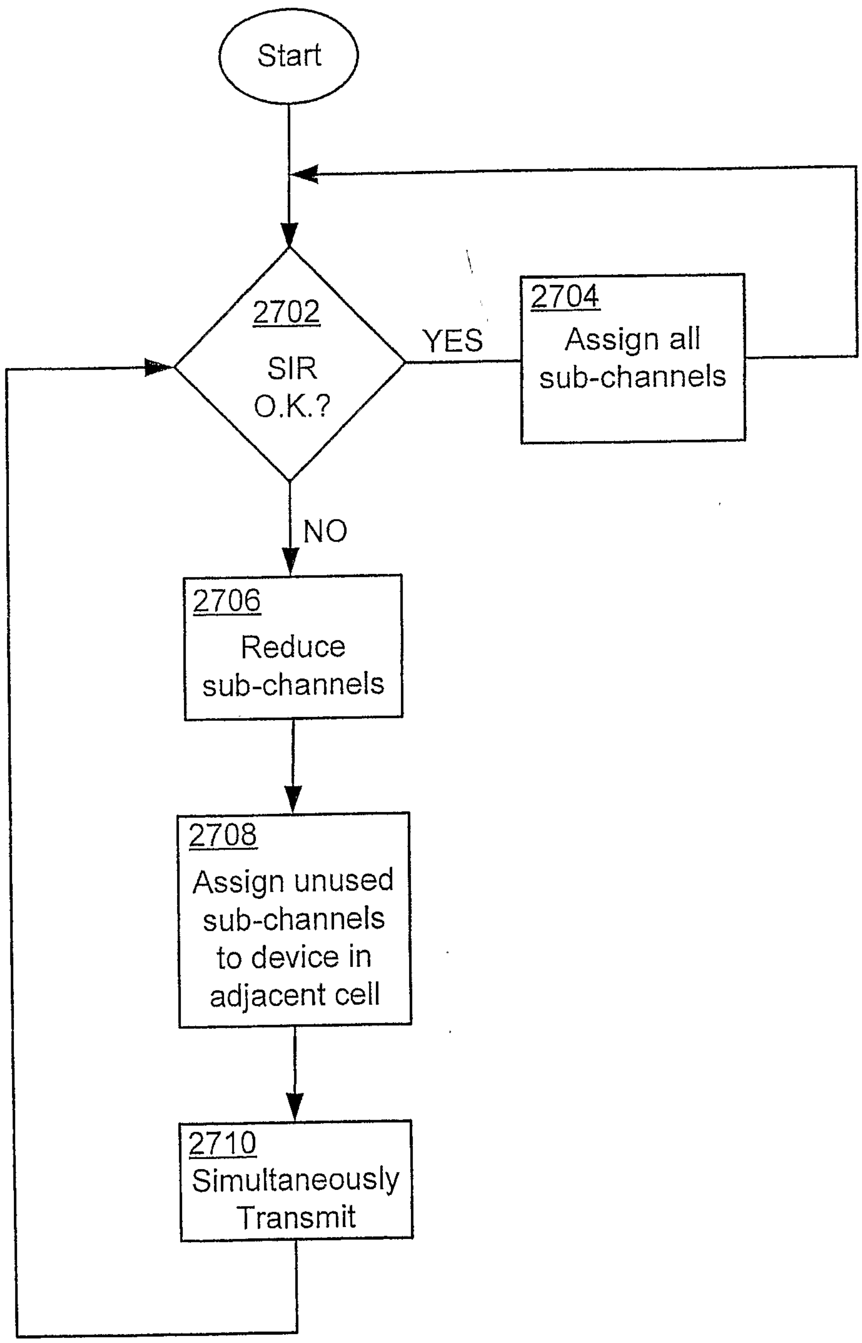

Start
2702
SIR
O.K.?
YES
2704
Assign all
sub-channels
NO
2706
Reduce
sub-channels
2708
Assign unused
sub-channels
to device in
adjacent cell
2710
Simultaneously
Transmit

FIG. 27

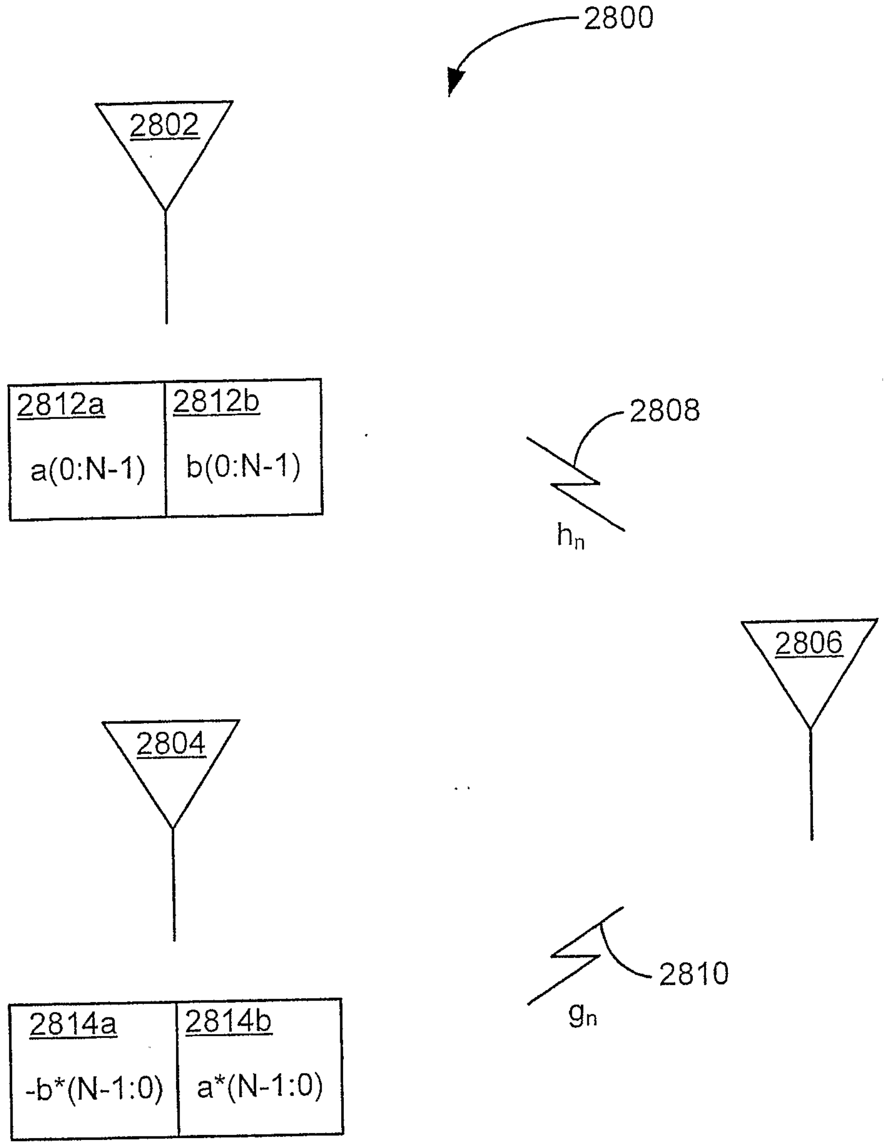

2800
2802
2812a
a(0:N-1)
2812b
b(0:N-1)
2808
h_n
2806
2804
2810
g_n
2814a
-b*(N-1:0)
2814b
a*(N-1:0)

FIG. 28

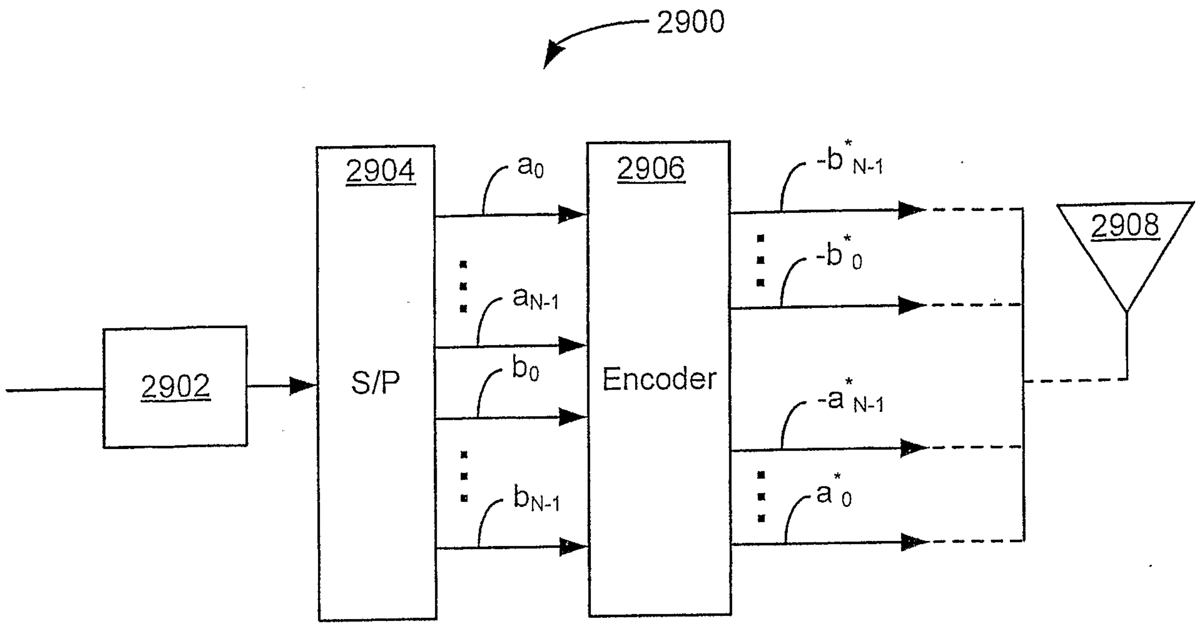

2900
2902
2904
S/P
a_0
a_{N-1}
b_0
b_{N-1}
2906
Encoder
$-b^*_{N-1}$
$-b^*_0$
$-a^*_{N-1}$
a^*_0
2908

FIG. 29

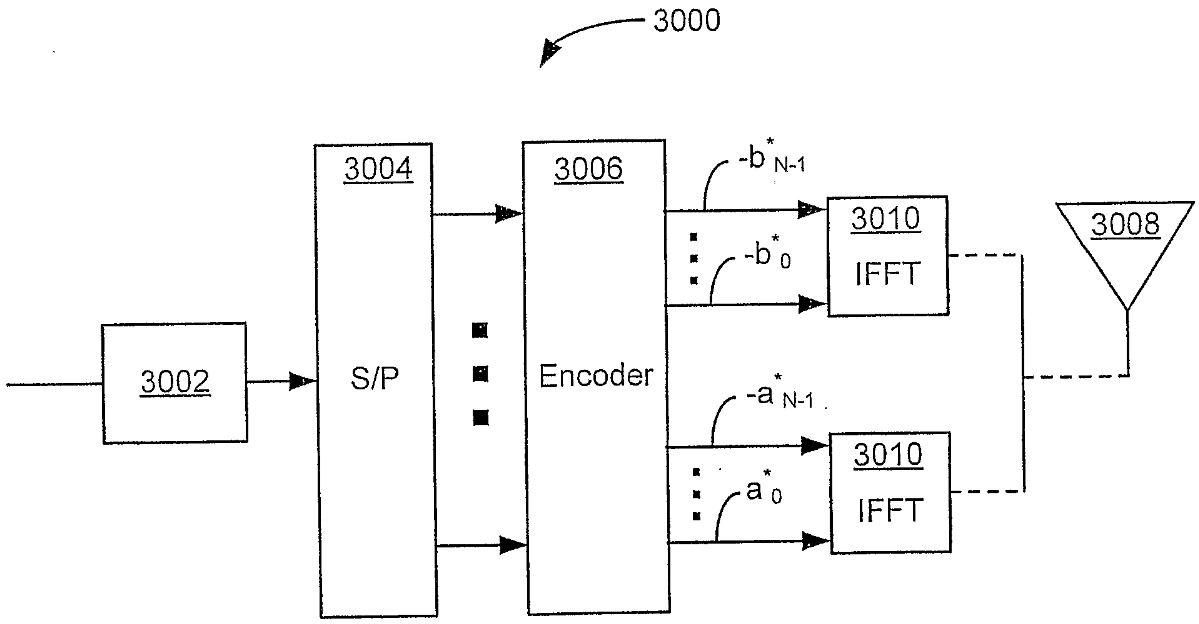

3000
3002
3004
S/P
3006
Encoder
$-b^*_{N-1}$
$-b^*_0$
$-a^*_{N-1}$
a^*_0
3010
IFFT
3010
IFFT
3008

FIG. 30

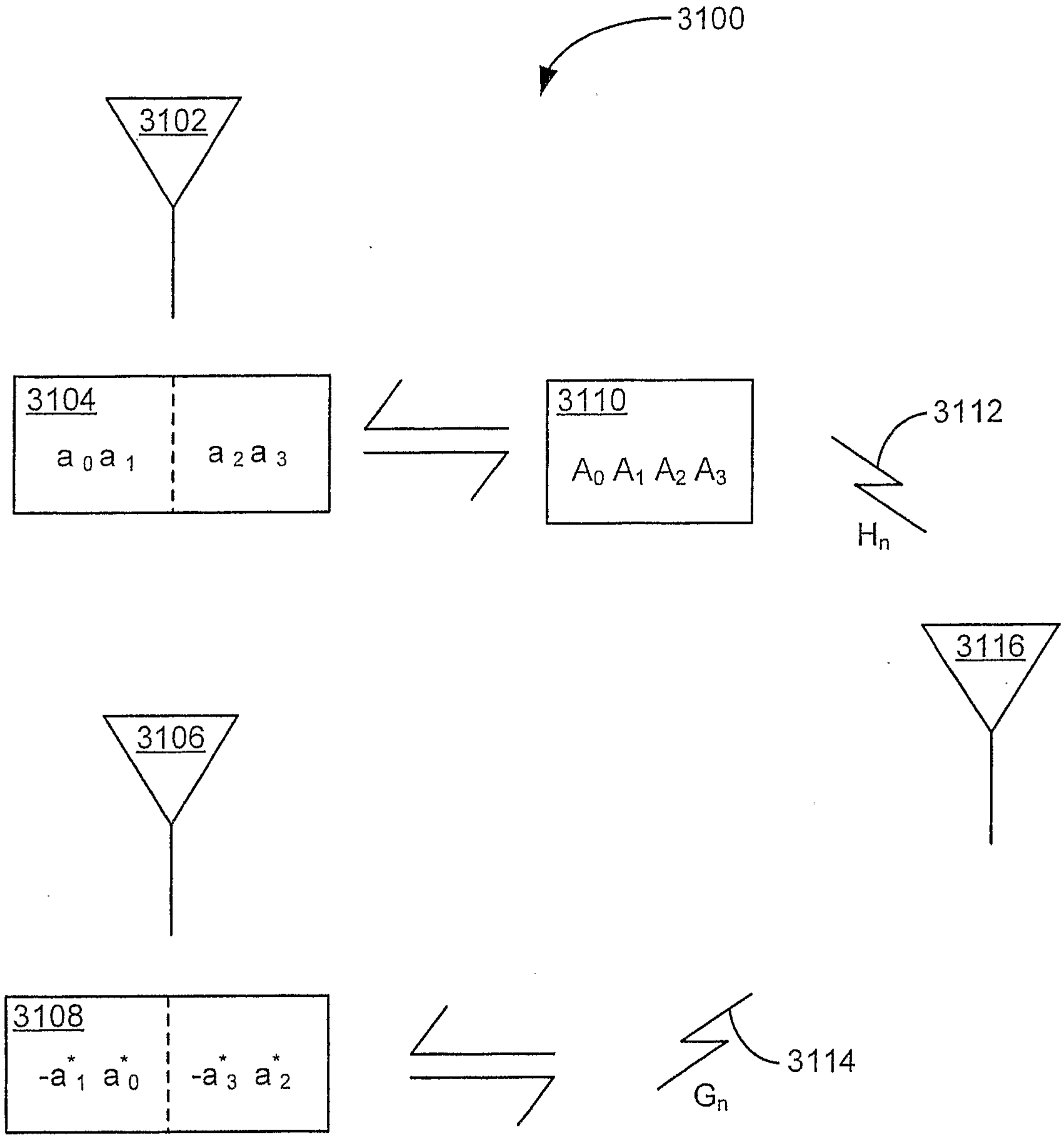

3100
3102
3104
a 0 a 1
a 2 a 3
3110
A0 A1 A2 A3
3112
Hn
3116
3106
3108
-a*1 a*0
-a*3 a*2
3114
Gn

FIG. 31

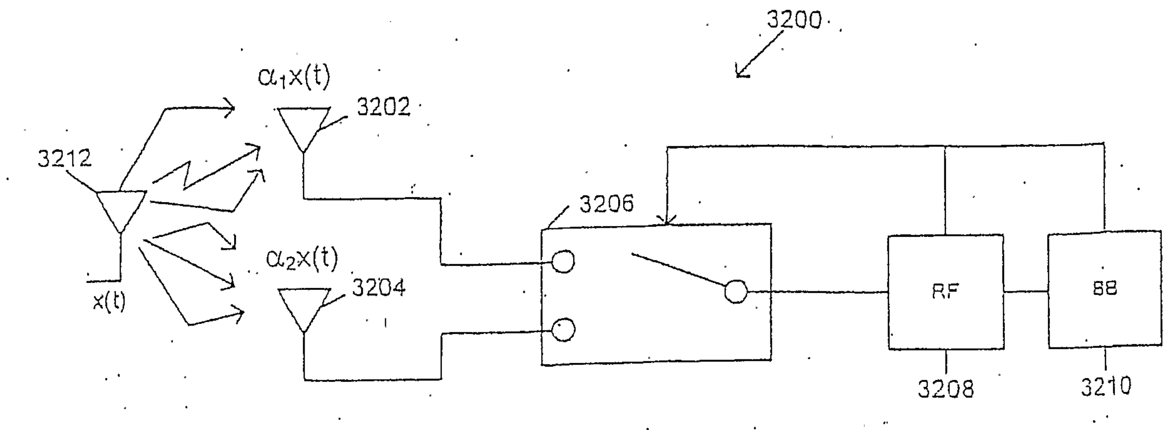

3200
$\alpha_1x(t)$
3202
3212
3206
x(t)
$\alpha_2x(t)$
3204
RF
BB
3208
3210

Fig. 32

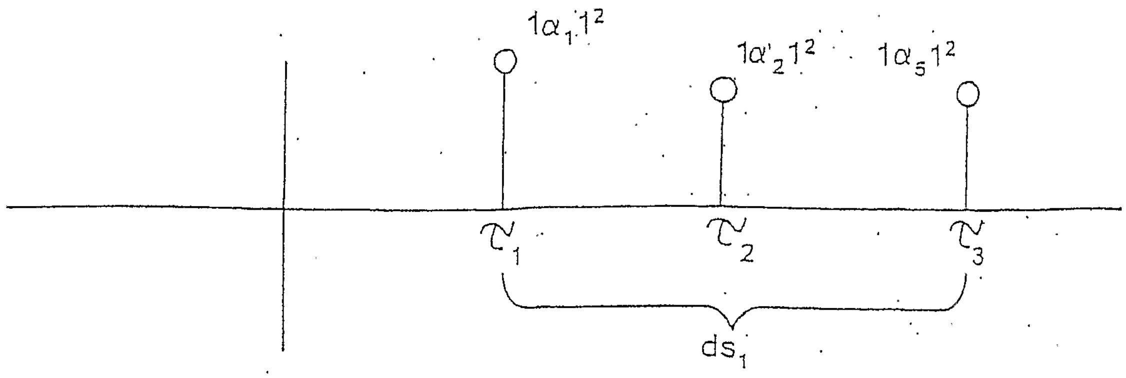


Fig. 33

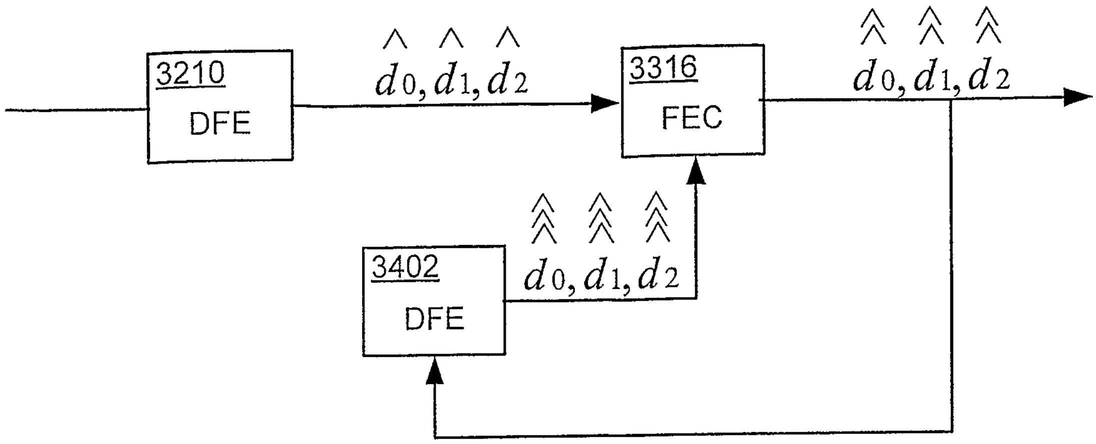

3210
DFE
d0, d1, d2
3316
FEC
d0, d1, d2
3402
DFE
d0, d1, d2

FIG. 34

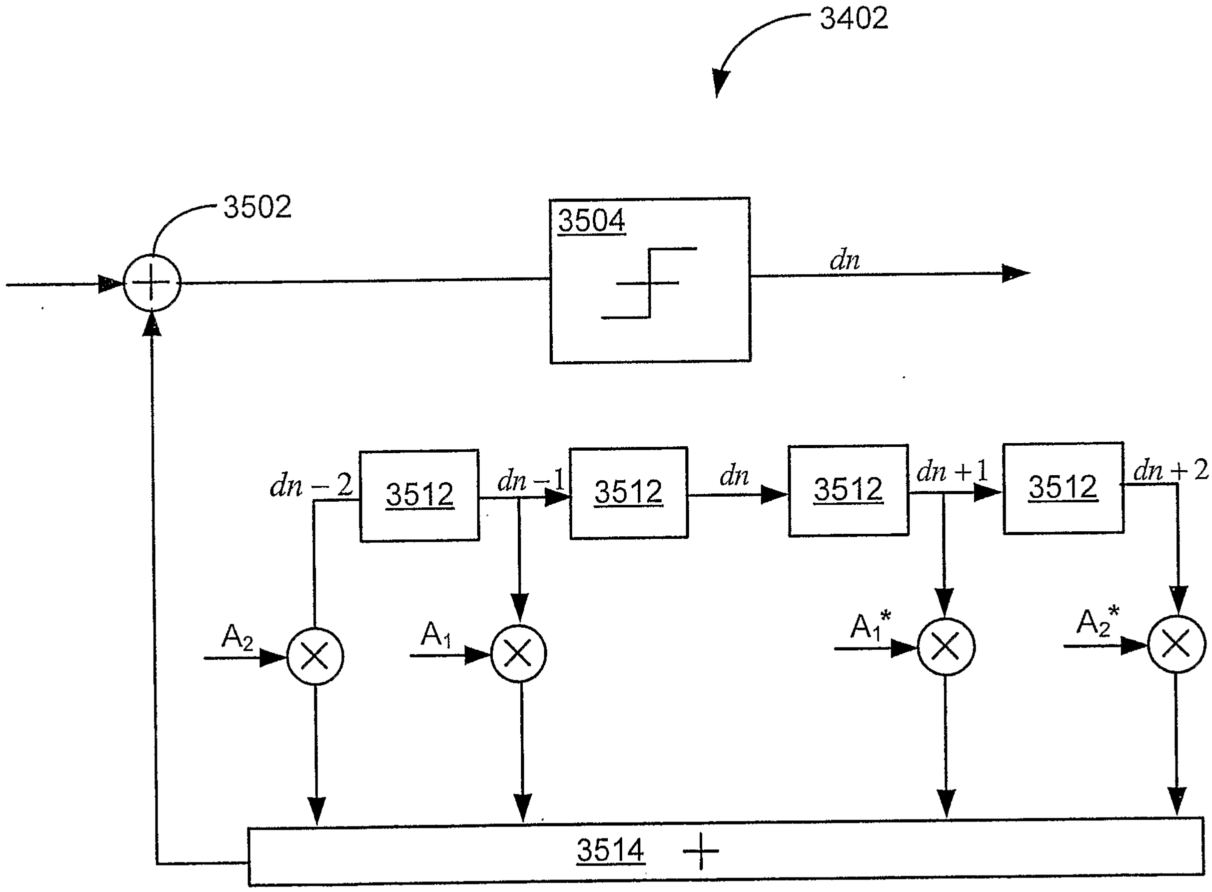

3402
3502
3504
dn
dn − 2
3512
dn − 1
3512
dn
3512
dn + 1
3512
dn + 2
A2
A1
A1*
A2*
3514 +

FIG. 35

MISO CIR Matched Filter

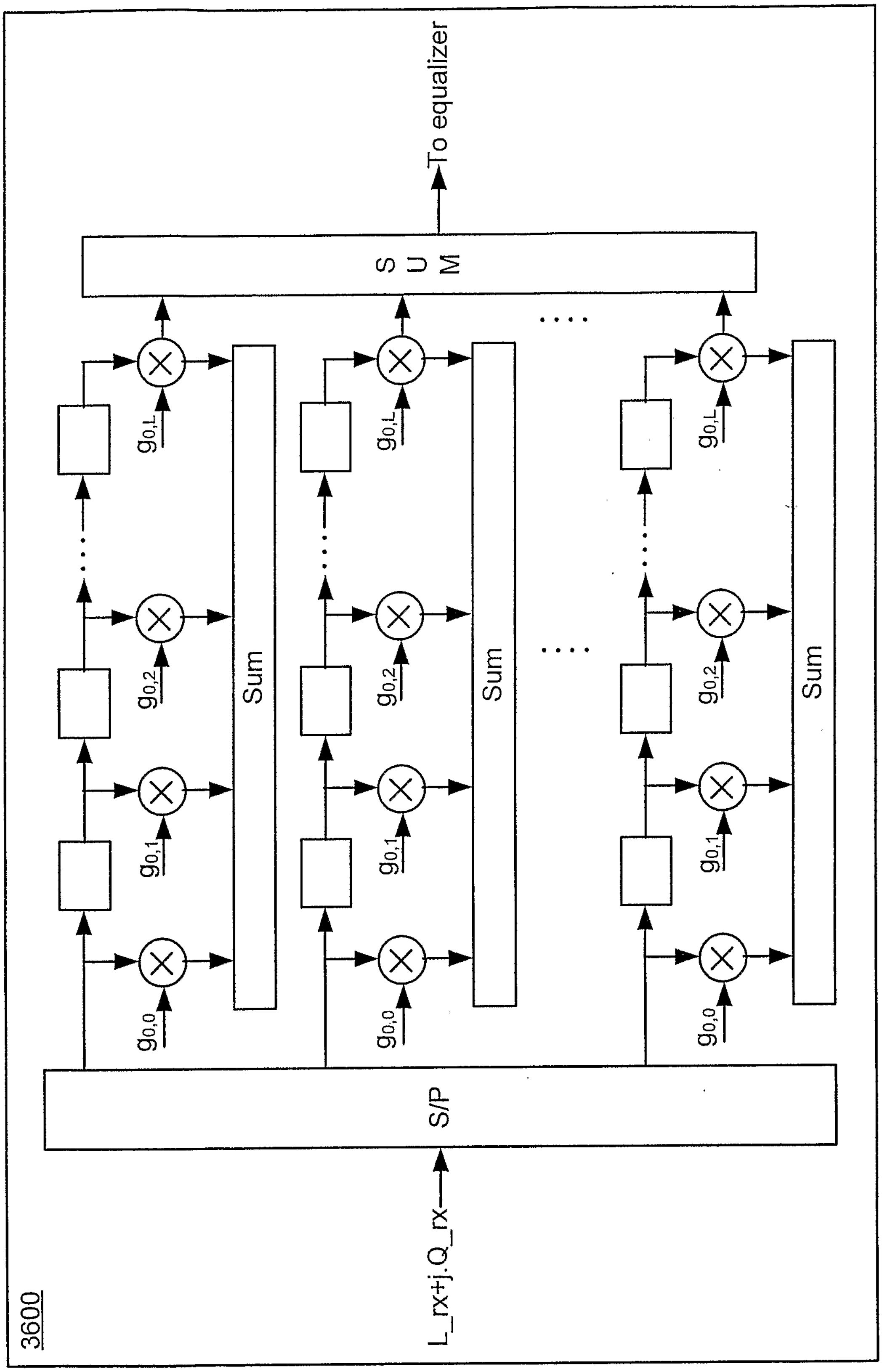


FIG. 36

Equalizer

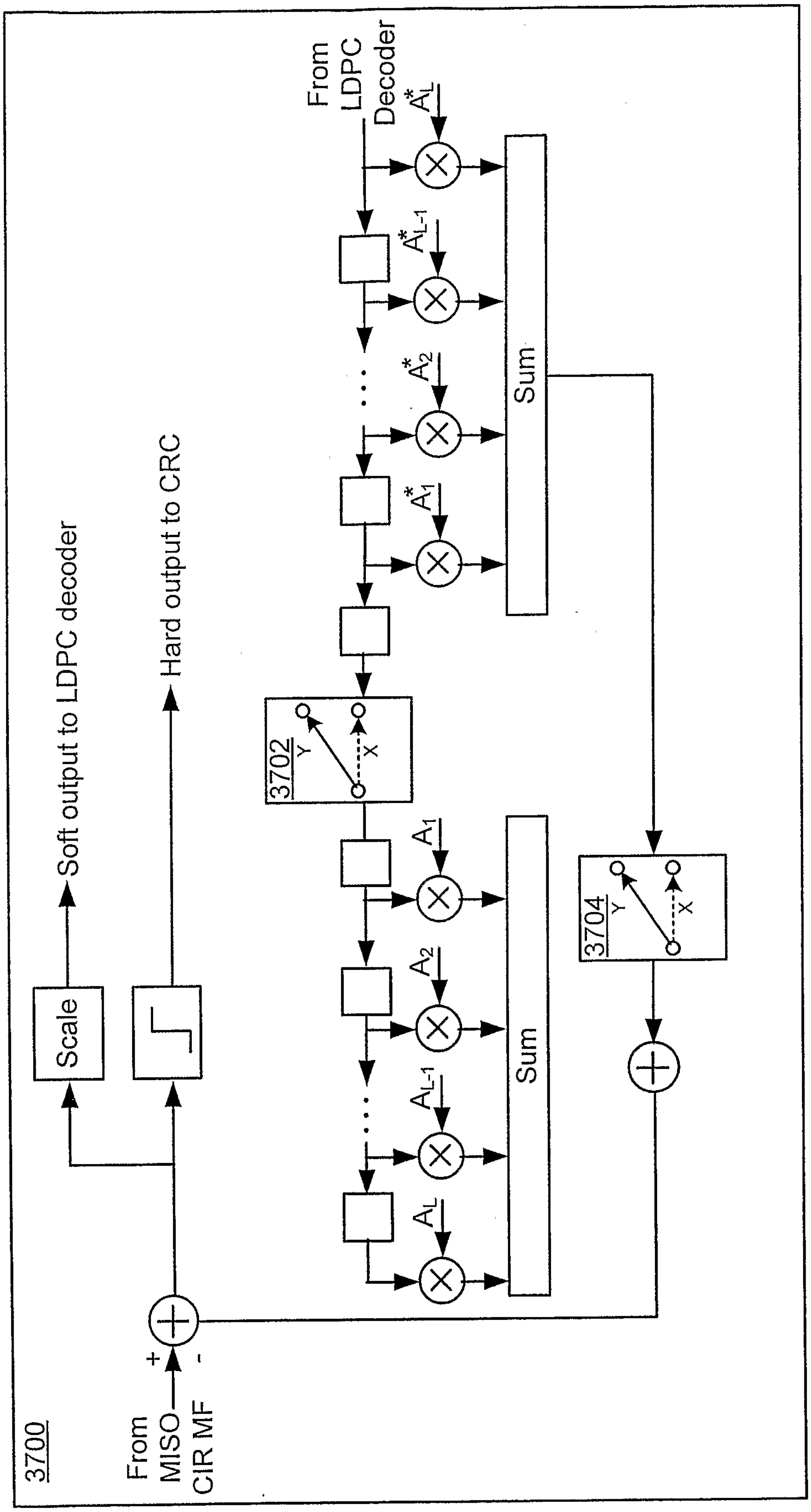


FIG. 37