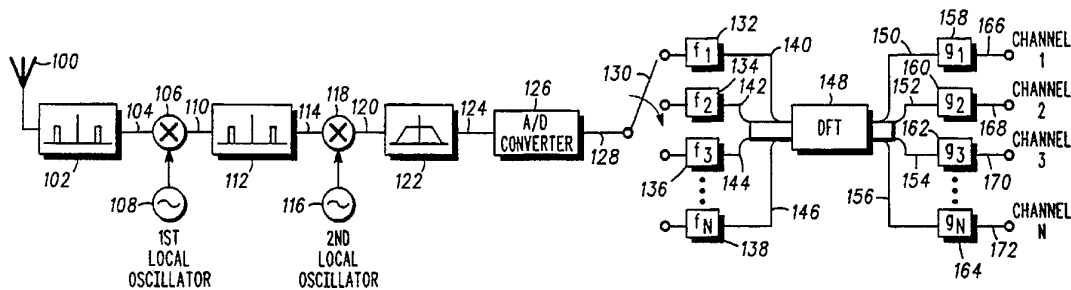




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(54) Title: A MULTI-CHANNEL DIGITAL TRANSMITTER AND RECEIVER



(57) Abstract

An apparatus is provided for use in a multi-channel digital transmitter and receiver. A filtering process is accomplished by conditioning portions of a digitized signal during digital signal processing by integrating, taking the difference of, and multiplying portions of the digitized signal. The filtering process may be used in conjunction with a receiving process to generate a digitized channel signal (166, 168, 170, 172) by conditioning (132, 134, 136, 138) the digitized signal (128) and subsequently Fourier transforming (148) the conditioned portions of the digitized signal. Alternatively, the filtering process may be used in conjunction with a transmitting process by inverse Fourier transforming (324) a plurality of digitized information signals (300, 302, 304, 306) to be transmitted, generating a composite digitized signal (352), and subsequently conditioning the composite digitized signal (358).

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A MULTI-CHANNEL DIGITAL TRANSMITTER AND RECEIVER

Cross Reference to Related Application

5 This is a continuation-in-part of U.S. Patent Application Serial No.
07/966,630 entitled "A Method and Means for Transmultiplexing Signals
Between Signal Terminals and Radio Frequency Channels" which was
filed on October 22, 1992 for the same inventor, Mark Harrison, as the
present patent application and is also assigned to Motorola, Inc. is
10 directed to a multi-channel digital transmitter and receiver which use a
polyphase filter bank to pre- and post- process digital signals,
respectively, which are used in conjunction with a Fourier transformer.

Field of the Invention

15 The present invention relates to communication units which
manipulate electromagnetic radiation within a frequency band, and
more specifically to a multi-channel receiver which detects a plurality of
information signals as well as a multi-channel transmitter which
20 prepares a plurality of information signals for subsequent transmission
over a communication channel.

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Background of the Invention

Transmitters and receivers for communication systems generally are designed such that they are tuned to transmit and receive one of a multiplicity of signals having widely varying bandwidths and which may fall within a particular frequency range. It will be appreciated by those skilled in the art that these transmitters and receivers radiate or intercept, respectively, electromagnetic radiation within a desired frequency band. The electromagnetic radiation can be output from or input to the transmitter or receiver, respectively, by several types of devices including an antenna, a wave guide, a coaxial cable, an optical fiber, and a transducer.

These communication system transmitters and receivers may be capable of transmitting and receiving a multiplicity of signals; however, such transmitters and receivers generally utilize circuitry which is duplicated for each respective signal to be transmitted or received which has a different frequency or bandwidth. This circuitry duplication is not an optimal multi-channel communication unit design architecture, because of the added cost and complexity associated with building complete independent transmitters and/or receivers for each communication channel.

An alternative transmitter and receiver architecture is possible which would be capable of transmitting and receiving signals having a desired multi-channel wide bandwidth. This alternative transmitter and receiver may utilize a digital-to-analog converter and a digitizer (analog-to-digital converter) which operates at a sufficiently high sampling rate to ensure that the signal of the desired bandwidth can be digitized in accordance with the Nyquist criterion (e.g., digitizing at a sampling rate equal to at least twice the bandwidth to be digitized). Subsequently, the digitized signal preferably is pre- or post- processed using digital signal processing techniques to differentiate between the multiple channels within the digitized desired bandwidth. Such a communication unit structure is essentially equivalent to a single radio frequency (RF) and intermediate frequency (IF) band analog signal processing portion followed by or preceded by a single digital processing portion which manipulates the digitized signal as if it represented multiple communication channels.

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It will be appreciated by those skilled in the art that another possible technique for providing this type of communication unit structure is through the use of Discrete Fourier Transforms (DFT's) in a DFT bank and Inverse Discrete Fourier Transforms (IDFT's) in an IDFT bank or similar digital filtering techniques, to synthesize a series of adjacent narrow bandwidth channels.

The disadvantage to this alternative type of communication unit is that the digital processing portion of the communication unit must have a sufficiently high sampling rate to ensure that the Nyquist criterion is met for the maximum bandwidth of the received electromagnetic radiation which is equal to the sum of the individual communication channels which form the composite received electromagnetic radiation bandwidth. If the composite bandwidth signal is sufficiently wide, the digital processing portion of the communication unit may be very costly and may consume a considerable amount of power. Additionally, the channels produced by a DFT or IDFT filtering technique must typically be adjacent to each other. Thus, the sampling rate for these filtering techniques is necessarily restricted to an integer multiplier of N times the number of possible communication channels (e.g., N , $2N$, $3N$, ... possible communication channels).

A need exists for a transmitter and a receiver, like the one which is described above, which is capable of transmitting and receiving a multiplicity of signals within corresponding channels with the same transmitter or receiver circuitry. However, this transmitter and receiver circuitry preferably should reduce communication unit design constraints associated with the digital signal processors (DSPs) which implement DFT and IDFT functions. If such a transmitter and receiver architecture could be developed, the it would be ideally suited for cellular radiotelephone communication systems. Cellular base stations typically need to transmit and receive multiple channels within a wide frequency bandwidth (e.g., 825 MegaHertz to 894 MegaHertz). In addition, commercial pressures on cellular infrastructure and subscriber manufacturers is prompting these manufacturers to find ways to reduce the cost of communication units. Similarly, such a multi-channel transmitter and receiver architecture would be well suited for personal communication systems (PCS) which will have smaller service regions (than their counterpart cellular service regions) for each base station

and as such a corresponding larger number of base stations will be required to cover a given geographic region. Operators which purchase base stations ideally would like to have a less complex and a reduced cost unit to install throughout their licensed service regions.

5 Current and future cellular and PCS communication systems have information signal coding and channelization standards (i.e., air interface standards) which share at least one common characteristic. The characteristic is that all of the systems include communication channels which are allocated substantially the same amount of
10 frequency bandwidth for each channel (i.e., each communication channel uses the same frequency bandwidth). A number of current and future planned information signal coding and channelization standards (i.e., open air interface standards) exist. These coding and
15 channelization standards which have channels allocated in equal portions of frequency bandwidth include channelization structures based on frequency division multiple access, time division multiple access, and frequency hopping code division multiple access. Some of the current and future planned coding and channelization standards have names including: Advanced Mobile Phone Service (AMPS),
20 Narrow Advanced Mobile Phone Service (NAMPS), Total Access Communication System (TACS), Japanese Total Access Communication System (JTACS), United States Digital Cellular (USDC), Japan Digital Cellular (JDC), Groupe Special Mobile (GSM), Frequency Hopping Spread Spectrum (FH-SS), Cordless Telephone 2
25 (CT2), Cordless Telephone 2 Plus (CT2 Plus), and Cordless Telephone 3 (CT3). It will be appreciated by those skilled in the art that, although digital communication open air interfaces typically have a group including more than one communication channel that occupies the same frequency bandwidth, the channels within the group are
30 separated from one another by a code (e.g., the code may be time slots within a time frame or the code may be slots within a frequency hopping pattern). Further, each group of channels utilizes the same frequency bandwidth as each other group of channels operating according to the same open air interface.

35 An additional advantage may be gained by cellular and PCS manufacturers as the result of designing multi-channel communication units which share the same analog signal portion. Traditional

communication units which are designed to operate under a single information signal coding and channelization standard. In contrast, these multi-channel communication units include a digital signal processing portion which may be reprogrammed, at will, through software during the manufacturing process or in the field after installation such that these multi-channel communication units may operate in accordance with any one of several information signal coding and channelization standards.

Summary of the Invention

A method and apparatus is provided for use in a multi-channel digital transmitter and receiver. A filtering process is accomplished by conditioning portions of a digitized signal during digital signal processing by integrating, taking the difference of, and multiplying portions of the digitized signal.

In addition, a method and apparatus is provided for utilizing the filtering process within a receiving process. The receiving process is accomplished by utilizing a plurality of filters to pre-condition a digitized signal for subsequent Fourier transformation. Each filter conditions a portion of the digitized signal by integrating, taking the difference of, and multiplying a portion of the digitized signal. Subsequently, a Fourier transformer generates a digitized channel signal from the pre-conditioned portions of the digitized signal. Alternatively, the receiving process may be described as utilizing an analog-to-digital converter to digitize a selected portion of an input frequency band. Subsequently, a commutator subdivides the digitized portion of the input frequency band into the plurality of digitized signal portions. Then, a discrete Fourier transformer generates a plurality of digitized channel signals by discrete Fourier transforming the plurality of digitized signal portions. Finally, a detector detects at least two information signals from the plurality of digitized channel signals.

Similarly, a method and apparatus is provided for utilizing the filtering process within a transmitting process. The transmitting process is accomplished by utilizing an inverse Fourier transformer to generate a composite digitized signal from a plurality of digitized information signals which are to be transmitted. Subsequently, a plurality of filters

post-conditions the composite digitized signal for subsequent transmission. Each filter post-conditions a portion of the composite digitized signal by integrating, taking the difference of, and multiplying a portion of the composite digitized signal. Alternatively, the transmitting process may be described as utilizing an inverse discrete Fourier transformer to generate a plurality of digitized signals by inverse discrete Fourier transforming a plurality of input digitized information signals. Subsequently, a commutator combines portions of the plurality of digitized signals into a composite digitized signal. Then, an digital-to-analog converter generates a composite analog transmission signal from the composite digitized signal. Finally, transmitter transmits the composite analog transmission signal over a frequency band.

Brief Description of the Drawings

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FIG. 1 is a block diagram showing a preferred embodiment receiver in accordance with the present invention.

FIG. 2 is a block diagram showing a portion of the preferred embodiment receiver shown in FIG. 1.

20 FIG. 3 is a block diagram showing a preferred embodiment filter in accordance with the present invention for use in the preferred embodiment receiver shown in FIGs. 1 and 2.

FIG. 4 is a block diagram showing a preferred embodiment transmitter in accordance with the present invention.

25 FIG. 5 is a block diagram showing a portion of the preferred embodiment transmitter shown in FIG. 4.

FIG. 6 is a block diagram showing a preferred embodiment filter in accordance with the present invention for use in the preferred embodiment transmitter shown in FIGs. 4 and 5.

30 FIG. 7 is a block diagram showing a preferred embodiment fifth order integrator in accordance with the present invention for use in the preferred embodiment receiver and transmitter shown in FIGs. 1 through 6

35 FIG. 8 is a block diagram showing a preferred embodiment fifth order differencer in accordance with the present invention for use in the preferred embodiment receiver and transmitter shown in FIGs. 1 through 6.

Detailed Description

5 It will be appreciated by those skilled in the art that although the following discussion is directed to a particular preferred embodiment receiver and transmitter communication unit for use in a cellular or PCS communication system, the principles discussed herein can be readily applied to other wireline, radio, or satellite telecommunication systems as well as data communication systems without departing from the scope and spirit of the present invention.

10 The preferred embodiment communication unit in accordance with the present invention as described below permits the manipulation of one or more information signals within a multiplicity of communication channels. The number of communication channels which can be received is determined by the frequency bandwidth of the signals to be received. Any combination of coded and channelized signals in between these bounds may be obtained and the channelization is not restricted to immediately adjacent channels within a frequency band. The receiver's channel bandwidths may also be modified by changing the signal processing algorithms. This allows the reuse of existing hardware and minimizes the need to replace hardware when the air interface standards of the communication system (e.g., cellular radiotelephone, paging, or any other wireline or wireless communication system) change or are expanded.

25 Referring now to FIG. 1, a block diagram of a preferred embodiment multi-channel receiver in accordance with the present invention is shown. It will be appreciated that communication units also typically include transmitter circuitry (such as the one shown and described later in reference to FIGs. 4-6). The preferred embodiment receiver is preferably included within a communication unit (e.g., a cellular base station, a cellular portable phone, a patio phone, a pager, or a trunked radio) for receiving electromagnetic radiation within a frequency band containing multiple communication channels. For the purposes of the following discussion, the preferred embodiment communication unit is described for use in an AMPS or NAMPS analog cellular radiotelephone communication system (as described in Electronic Industrial Association standard EIA-553 and interim

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standards IS-88, IS-89, and IS-90). The communication channels used by each cell or service region (served by a base site) are typically assigned according to a channel reuse separation pattern (e.g., a 21-cell or 24-cell reuse pattern). This type of channel assignment based on a reuse separation pattern reduces co-channel and adjacent channel interference. Therefore, each base site is designed to receive only a subset of all available communication channels (i.e., 1/21 or 1/24 of all available channels). It will be appreciated by those skilled in the art that other reuse separation patterns may be used by an AMPS or NAMPS communication system and as such each base site will need to be designed to receive the number of channels available for serving the cell or service region served by the base site.

The receiver preferably includes an input such as an antenna **100** which intercepts electromagnetic radiation within the frequency band of interest and transducer (not shown) for converting the intercepted electromagnetic radiation into an electrical signal. It will be appreciated by those skilled in the art that other types of input devices may be used for intercepting or capturing electromagnetic radiation. For example a wave guide, a coaxial cable, an optical fiber, or an infrared frequency transducer may be used to intercept electromagnetic radiation for subsequent input into the preferred embodiment receiver.

The receiver preferably consists of a front end (analog portion) consisting of elements known to those skilled in the art that provide a low noise figure and ample protection to undesired out-of-band signals. These elements consist of a preselect band pass filter **102**. The output electrical signal **104** of the preselect band pass filter **102** is input to a down converter. The down converter consists of a mixer **106**, a local oscillator **108**, and a band pass filter **112**. The oscillator **108** preferably is a fixed frequency oscillator; however, as will be appreciated by those skilled in the art a frequency agile local oscillator may be used in place of the fixed frequency oscillator without departing from the scope and spirit of the present invention. The electrical signal **104** is operatively coupled to one input of the mixer **106** and the output of the oscillator **108** is provided to another input of mixer **106**. Mixer **106** mixes the electrical signal **104** down to an intermediate frequency band electrical signal **110**. Subsequently, the intermediate frequency (IF) band electrical signal **110** is filtered by a band pass filter **112** to

eliminate undesired frequency components into a conditioned IF band electrical signal **114**.

Subsequently, the IF band electrical signal **114** is input to another down converter. This down converter consists of a mixer **118**, a
5 frequency agile local oscillator **116**, and a low pass filter **122**. The electrical signal **114** is operatively coupled to one input of the mixer **118** and the output of the oscillator **116** is provided to another input of mixer **118**. Mixer **118** mixes the IF band electrical signal **114** down to a baseband electrical signal **120**. Subsequently, the baseband
10 electrical signal **120** is filtered by a low pass filter **122** to eliminate undesired frequency components into a conditioned baseband electrical signal **124**.

The conditioned baseband electrical signal **124** preferably is operatively coupled to a digitizing device **126** (i.e., analog-to-digital
15 (A/D) converter) which digitizes a portion of the baseband electrical signal **124** into a digitized signal **128**. The digitized signal **128** represents the electromagnetic radiation within the intercepted frequency band. FIG. 2 shows a graphical representation **128'** of this digitized signal **128** which contains a plurality of channels (i.e., N
20 channels) contained in frequency divided passbands. The sampling rate of the digitizing device **126** preferably is set to be within the Nyquist criterion of at least twice the highest desired frequency with the baseband electrical signal **124**.

Commutator **130** receives the digitized signal **128** and
25 operatively couples portions of the digitized signal **128** to a discrete Fourier transform pre-filter bank including filters **132**, **134**, **136**, and **138** (i.e., f_1 , f_2 , f_3 , and f_N). For the purposes of the following description, four filters **132**, **134**, **136**, and **138** are shown in FIG. 1; however, it will be appreciated by those skilled in the art that the principles described
30 herein may be applied to more or less filters in the filter bank. The commutator **130** distributes portions of the digitized signal **128** to four filters **132**, **134**, **136**, and **138** such that if commutator **130** is operating at a first sampling rate (F_s), then each of the N filters receives portions of the digitized signal **128** at a second sampling rate (F_s/N)
35 where N is the number of filters to which the commutator **130** operatively couples the digitized signal **128** (as previously noted N also corresponds to the number of channels in the digitized signal **128**).

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Filters **132**, **134**, **136**, and **138** attenuate undesired frequency components in the digitized signals (i.e., condition the digitized signals) so that the computational complexity of the subsequent discrete Fourier transform (DFT) operation may be reduced. These filters **132**, **134**,
5 **136**, and **138** operate similar to polyphase filters in known DFT pre-filter banks in that they receive their inputs from a commutator, and run in parallel to form the input to a discrete Fourier transformer. One such polyphase filter bank is shown and described in the previously cited related U.S. Patent Application Serial No. 07/966,630 entitled "A
10 Method and Means for Transmultiplexing Signals Between Signal Terminals and Radio Frequency Channels" which was filed on October 22, 1992. However, as will be described in reference to FIGs. 2 and 3, the preferred embodiment receiver structure of the present patent application utilizes modified Cascaded Integrator Comb (CIC) filters
15 **132**, **134**, **136**, and **138** rather than polyphase filters. The outputs **140**, **142**, **144**, and **146** of the respective filters **132**, **134**, **136**, and **138** form the inputs to a discrete Fourier transformer **148**.

Discrete Fourier transformer **148** preferably operates at the same or higher sampling rate of the filters **132**, **134**, **136**, and **138**. The
20 discrete Fourier transformer **148** performs one transform (e.g., fast Fourier transform) per input block of samples **140**, **142**, **144**, and **146** from the pre-filters **132**, **134**, **136**, and **138**. Each of the N outputs **150**, **152**, **154**, and **156** of the discrete Fourier transformer **148** contains samples for one of the N different down converted channels.
25 These N outputs **150**, **152**, **154**, and **156** are generated by performing an N point Fourier transform which produces one output sample for each of the N channels. In the preferred embodiment receiver structure the fast Fourier transformer (FFT), operating at a rate higher than the sampling rate of the filters **132**, **134**, **136**, and **138**, (e.g., the FFT may
30 operate at a rate which is three times greater than the filter sampling rate) are used to perform the operations of the discrete Fourier transformer **148**. This high rate of operation for the FFT is required to avoid aliasing, because the modified CIC pre-filters used in the preferred embodiment receiver have a rolloff rate which is not as high as
35 typical polyphase filter banks.

The use of FFTs allows the receiver structure to be implemented in a computationally efficient manner while providing for a means to

reduce the effects of errors associated with discrete Fourier transforms. Some of these potential error sources in discrete Fourier transformers include: errors due to aliasing (i.e., errors resulting from overlapping of the decimated signal in the frequency domain), spectral distortion from leakage (i.e., spectral energy leaking from one frequency to another), and the picket-fence effect which results in important spectral components being missed (i.e., the discrete sampling omits particular spectral components of a signal which occurs between sample points). These sources of errors in the discrete Fourier transformers may be reduced through a variety of techniques in FFTs. For example, the errors due to aliasing may be reduced by increasing the sampling rate of the discrete Fourier transformer **148** and/or pre-filtering to minimize high-frequency spectral content of the inputs to the discrete Fourier transformer **148**. Further, the spectral distortion due to leakage may be reduced by increasing the number of DFT points through operating at a higher sampling rate or increasing the window width using window functions that have Fourier transforms with low sidelobes, and/or eliminating large periodic components by filtering before performing window functions. Furthermore, the picket-fence effect may be reduced by increasing the number of DFT points while holding the sampling rate fixed such that the "pickets" are moved closer together and/or filling in additional DFT points with zeros (i.e., zero padding), if the data length is limited.

Because the discrete Fourier transformer **148** in the preferred embodiment receiver structure operates at a rate higher than the decimated sampling rate of the filters **132**, **134**, **136**, and **138**, N decimating post-filters **158**, **160**, **162**, and **164** must be used to bring down the corresponding baseband response outputs **150**, **152**, **154**, and **156** sampling rate to a desired sampling rate of N conditioned outputs **166**, **168**, **170**, and **172**, respectively (e.g., the post-filters bring down the sampling rate to typical voice or data sampling rates of less than 9600 bits/second). These N post-filters **158**, **160**, **162**, and **164** (denoted g_1 , g_2 , g_3 , and g_n) may be computationally simple (e.g., the filters may be Infinite Impulse Response (IIR) filters), if phase distortion is not critical. In addition, if recovery of only a subset of the N channels is desired (e.g., in lightly loaded communication systems), then these post-filters need only be implemented for that subset of

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desired channels. In the preferred embodiment receiver structure, the N decimating post-filters **158**, **160**, **162**, and **164** preferably are designed with respect to at least one signal characteristic including: optimum selectivity, controlled phase response, and controlled
5 amplitude response.

Finally, an information signal (e.g., voice and or data) may be detected within each of the N conditioned outputs **166**, **168**, **170**, and **172**. FIG. 2 shows a graphical representation **166'**, **168'**, **170'**, and **172'** of these conditioned outputs **166**, **168**, **170**, and **172**. Each
10 graphical representation shows one of the N channels as received from the plurality of channels contained in frequency divided passbands of the digitized signal **128**. It will be appreciated by those skilled in the art that the information signal may be detected through the use of several decoding steps depending upon the particular signal coding and
15 channelization standard to be received. For example, each conditioned output **166**, **168**, **170**, and **172** may need to be convolutionally decoded, maximum likelihood sequence estimated, or vocoded to retrieve the voice or data present in the particular conditioned output.

Referring now more particularly to FIG. 2 and the pre-filters **132**,
20 **134**, **136**, and **138** (f_1 , f_2 , f_3 , and f_N), the pre-filters **132**, **134**, **136**, and **138** receive their inputs from a commutator **130** and operate in parallel to form the inputs **140**, **142**, **144**, and **146**, respectively, for the DFT filter bank **148**. However, the preferred embodiment pre-filters **132**,
25 **134**, **136**, and **138** incorporate a bank of Hogenauer's Cascaded Integrator Comb (CIC) filters. This type of filter is generally described in an article by E.B. Hogenauer, "An Economical Class of Digital Filters for Decimation and Interpolation", IEEE Trans. Acoust. Speech. Signal Processing, vol. ASSP-20, No. 2, April, 1981, pp. 155-162. The CIC
30 filters have advantages over typical polyphase filters in that CIC filters consist of cascaded first order integrators and combs (i.e., differencers), which allows these filters to be implemented without multipliers. In addition, since CIC filters include decimators, the multiply-accumulators (MACs) which are also in the pre-filters can be operated at a slower rate
35 to further reduce the required computations. Thus, the pre-filters are more computationally efficient and, when implemented in a device, require less computational power, size, and execution time, than a typical polyphase filter.

5 However, since these principles, discussed by Hogenauer, related to CIC filters were not specifically designed to be used in the preferred embodiment multi-channel communication unit receiver structure, they must be modified. For simplicity, the following discussion is directed to a modified second order CIC filter. However, it will be appreciated by those skilled in the art that for the preferred embodiment receiver structure a modified fifth order CIC filter would provide a more optimal filter characteristic for this particular receiver structure.

10 If a mixer is used in conjunction with a modified CIC filter bank, then a down converter is formed. The second order down converter $y_2(MNT)$ can be described by the following equation:

$$y_2(MNT) = \sum_{j=0}^{M-1} \sum_{l=0}^{M-1} \sum_{k=0}^{N-1} \sum_{m=0}^{N-1} x(M[NT - k - m] - j - l) \Phi(M[NT - k - m] - j - l) \quad (\text{eq. 1})$$

15 where:

$\Phi(M[NT - k - m] - j - l)$ = the mixing sequence

$x(M[NT - k - m] - j - l)$ = the input to the down converter

M = the number of channel outputs to be input into the FFT

N = the decimation rate of the modified CiC filter

20 T = the time as referenced to the FFT's output

k, m, j, and l = indices.

If the mixing sequence is periodic in M, then the sum may be separated from (eq. 1) as follows:

25

$$y_2(MNT) = \sum_{j=0}^{M-1} \sum_{l=0}^{M-1} \Phi(-j-l) \sum_{k=0}^{N-1} \sum_{m=0}^{N-1} x(M[NT - k - m] - j - l). \quad (\text{eq. 2})$$

The expression in (eq. 2) can be simplified by rewriting it as follows:

30

$$y_2(MNT) = \sum_{j=0}^{M-1} \sum_{l=0}^{M-1} \Phi(-j-l) Z(MNT - j - l) \quad (\text{eq. 3})$$

where:

$$Z(MNT - j - l) = \sum_{k=0}^{N-1} \sum_{m=0}^{N-1} x(M[NT - k - m] - j - l).$$

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The expression in (eq. 3) now has two distinct operations of computing Z and computing y_2 . Upon further examination of the Z operation, it should be noted that each Z operation is a modified second order CIC filter function operated at a particular phase (j+1) of a decimated version of the input x. These CIC filters preferably decimate by N. As a result, the computation of y_2 only involves computations based on the modified CIC filter outputs (Z). For further simplification, the expression in (eq. 3) can be rewritten to emphasize the dependence of the product term on (j+1) as follows:

10

$$y_2(MNT) = \sum_{j=0}^{M-1} \sum_{\Delta=j}^{M-1+j} \Phi(-\Delta) Z(MNT - \Delta). \quad (\text{eq. 4})$$

The expression in (eq. 4) can be rewritten to eliminate one of the summation terms as follows:

15

$$\begin{aligned} y_2(MNT) &= \sum_{j=0}^{M-1} \sum_{\Delta=j}^{M-1+j} \Phi(-\Delta) Z(MNT - \Delta) \\ &= \sum_{j=0}^{M-1} \left[\sum_{\Delta=0}^{M-1} \Phi(-\Delta) Z(MNT - \Delta) - \sum_{\Delta=0}^{j-1} \Phi(-\Delta) Z(MNT - \Delta) \right. \\ &\quad \left. + \sum_{\Delta=M}^{M-1+j} \Phi(-\Delta) Z(MNT - \Delta) \right] \\ &= M \sum_{\Delta=0}^{M-1} \Phi(-\Delta) Z(MNT - \Delta) \\ &\quad + \sum_{j=0}^{M-1} \left[\sum_{\Delta=0}^{j-1} \Phi(-\Delta) Z(MNT - \Delta) + \sum_{\Delta=0}^{j-1} \Phi(-\Delta) Z(M[NT - 1] - \Delta) \right] \\ &= M \sum_{\Delta=0}^{M-1} \Phi(-\Delta) Z(MNT - \Delta) \\ &\quad + \sum_{j=0}^{M-1} \left[\sum_{\Delta=0}^{j-1} \Phi(-\Delta) \{ Z(M[NT - 1] - \Delta) - Z(MNT - \Delta) \} \right] \\ &= M \sum_{\Delta=0}^{M-1} \Phi(-\Delta) Z(MNT - \Delta) + Y(0) + \sum_{\Delta=0}^1 Y(\Delta) + \sum_{\Delta=0}^2 Y(\Delta) + \dots + \sum_{\Delta=0}^{M-2} Y(\Delta) \\ &= M \sum_{\Delta=0}^{M-1} \Phi(-\Delta) Z(MNT - \Delta) + (M-1)Y(0) + (M-2)Y(1) + \dots + Y(M-2) \end{aligned}$$

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$$\begin{aligned}
&= M \sum_{\Delta=0}^{M-1} \Phi(-\Delta) Z(MNT - \Delta) + \sum_{i=0}^{M-1} (M-i-1) Y(i) \\
&= M \sum_{\Delta=0}^{M-1} \Phi(-\Delta) Z(MNT - \Delta) \\
&\quad + \sum_{\Delta=0}^{M-1} (M-\Delta-1) \Phi(-\Delta) \{Z(M[NT-1]-\Delta) - Z(MNT - \Delta)\} \\
&= \sum_{\Delta=0}^{M-1} \Phi(-\Delta) \{(\Delta+1)Z(MNT - \Delta) + (M-\Delta-1)Z(M[NT-1]-\Delta)\} \quad .
\end{aligned}$$

(eq. 5)

If the mixing sequence is set to be equal to $\Phi(\Delta) = \exp(-j2\pi\Delta f/M)$, then
 5 (eq. 5) can be solved as an inverse discrete Fourier transform of the bracketed sum as follows:

$$y_2(MNT) = \sum_{\Delta=0}^{M-1} \exp(j2\pi\Delta f/M) w(MNT - \Delta) \quad (\text{eq. 6})$$

where:

10

$$\begin{aligned}
w(MNT - \Delta) &= (\Delta+1)Z(MNT - \Delta) + (M-\Delta-1)Z(M[NT-1]-\Delta) \\
j &= \sqrt{-1}.
\end{aligned}$$

As previously noted, the modified second order CIC filter
 15 expression derived above which resulted in (eq. 6) does not have an optimal filter characteristic. Stated differently, the modified second order CIC filter will not provide enough selectivity for the preferred embodiment 32 channel NAMPS receiver as shown in FIGs. 1 and 2. For this preferred embodiment an eighty decibel (80 dB) stop band is
 20 required in the modified CIC filter. This can be accomplished with a modified fifth order CIC filter which is substantially similar to the modified second order CIC filter described above. A fth channel output of the modified fifth order CIC filter $y_5^f(MNT)$ can be described as follows:

$$25 \quad y_5^f(MNT) = \sum_{\Delta=0}^{M-1} \exp(-j2\pi\Delta f/M) w(MNT + \Delta)$$

where:

$$w(MNT + \Delta) = \sum_{k=0}^4 a_k(\Delta) Z(M[NT-k] + \Delta)$$

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and where:

$$Z(M[NT - k] + \Delta) = \sum_{k_0=0}^{N-1} \sum_{k_1=0}^{N-1} \sum_{k_2=0}^{N-1} \sum_{k_3=0}^{N-1} \sum_{k_4=0}^{N-1} [x(M[NT - k_0 - k_1 - k_2 - k_3 - k_4] + \Delta)]$$

$$a_k(\Delta) = A(kM - \Delta)$$

M = the number of channel outputs to be input into the FFT

5 N = the decimation rate of the modified CiC filter

T = the time as referenced to the FFT's output.

(eq. 7)

10 It will be appreciated by those skilled in the art that the sign of the Δ has been changed in (eq. 7) from the sign in (eq. 6). This allows a DFT (instead of a IDFT) to be used in the preferred embodiment receiver structure. In addition, $A(k)$ preferably is an impulse response of a 5th order CIC filter (not modified CIC) with a decimation rate M. This
 15 impulse response may be found in a variety of ways. One way is to use FFTs. Since the impulse response of a 1st order CIC filter with a decimation rate M is a unit step of length M, the response of the 5th order CIC filter is just the convolution of five of these unit steps together. This convolution may be computed using the inverse FFT of the FFT of a
 20 unit step of length M padded by $4(M-1)+1$ zeroes, raised to the 5th power. It should be noted by those skilled in the art that this $A(k)$ is only one of many $A(k)$ sets of coefficients which may serve as combiner coefficients for the CIC outputs. Other coefficients may be generated which produce other desirable frequency responses.

25 The modified fifth order CIC filter expression of (eq. 7) may be implemented in a variety of different means. However, a preferred embodiment fifth order modified CIC filter design is shown in FIG. 3. The fifth order filter design is used to implement each of the parallel modified CIC filters **132**, **134**, **136**, and **138** (i.e., f_1 , f_2 , f_3 , and f_N) as
 30 shown in FIG. 2 and previously described. By way of example, the modified fifth order CIC filter shown in FIG. 3 will be described as modified CIC filter **138** (f_N); however, each of the modified CIC filters **132**, **134**, **136**, and **138** (i.e., f_1 , f_2 , f_3 , and f_N) in the preferred

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embodiment communication unit receiver preferably implements a substantially similar modified CIC filter design.

The preferred embodiment modified CIC filter **138** receives a portion of the digitized signal **128** at an input. The portion of the digitized signal received by modified CIC filter **138** is determined by commutator **130** which operatively couples portions of the digitized signal **128** to each of the modified CIC filters **132**, **134**, **136**, and **138** (i.e., f_1 , f_2 , f_3 , and f_N). This digitized signal portion **128/130** is input to a fifth order integrator **174** which integrates bits within received in the digitized signal portion **128/130** over an infinite time period to produce part of $Z(M[NT] - \Delta)$ of (eq. 7).

A preferred embodiment fifth order integrator **174** is shown in FIG. 7 which consists of five cascaded 1st order integrators. Signal portion **128/130** is provided to a summer **701** which adds the output of delay element **702** to signal portion **128/130**. The summer output **703** is then provided to the next integrator stage, consisting of summer **704** and delay element **705**, which adds the output of delay element **705** to with summer output **703** to produce summer output **706**. This integration operation is repeated three more times utilizing summers **707**, **710**, and **713** as well as delay element **708**, **711**, and **714**. The result of this fifth order integration process is integrator **174** result **176**.

The fifth order integrator **174** result **176** is output to fifth order differencers **198**, **200**, **202**, **204**, and **206** (Δ_0 , Δ_1 , Δ_2 , Δ_3 , and Δ_4) through delay elements **178**, **180**, **182**, and **184**. More precisely, the integrator **174** result **176** (i.e., integrator output) is input to differencer **198** after being decimated to a lower sample rate (e.g., one seventh of the sample of the integrator **174** result **176**) by a decimating device **186**. It will be appreciated by those skilled in the art that this decimation of result **176** allows the subsequent parts of (eq. 7) to be implemented within pre-filter **138** in a less computationally complex manner, than would otherwise be required if the result **176** was not decimated. The differencer **198** implements part of the function within (eq. 7) and computes $Z(M[NT - k] - \Delta)$ when $k=0$).

By way of example differencer **198** is shown in FIG. 8 consisting of five cascaded 1st order differencers. The output of decimator **186** is provided to summer **801** which subtracts the output of delay element

802 from the output of decimator **186**. The summer output **803** is then provided to the next differencer stage, consisting of summer **804** and delay element **805**, which subtracts the output of delay **805** from summer output **803** to produce summer output **806**. This differencing operation is repeated three more time utilizing summers **807**, **810**, and **813** as well as delay elements **808**, **811**, and **814** to produce the fifth order differencer **198** output **208**. It will be appreciated by those skilled in the art that each of the other fifth order differencers **200**, **202**, **204**, and **206** may be implemented in a manner similar to that which has been described in reference to FIG. **8** for differencer **198**.

Subsequently, the output **208** of differencer **198** (i.e., the function $Z(M[NT]-\Delta)$) is scaled by a factor a_0 with multiplier **218**. Similarly, the integrator **174** result **176** is input to differencer **200** after being delayed for one sample time by delay element **178** and subsequently decimated to a lower sample rate by a decimating device **188**. The output **210** of differencer **200**, which is the result of the function $Z(M[NT-k]-\Delta)$ when $k=1$, is scaled by a factor a_1 with multiplier **220**. Further, a delayed and decimated version of result **176** is input to differencer **202** after being delayed for two sample times by delay elements **178** and **180** and decimated by a decimating device **190**. The output **212** of differencer **202** (i.e., the result of $Z(M[NT-k]-\Delta)$ when $k=2$) is scaled by a factor a_2 with multiplier **222**. Furthermore, a delayed and decimated version of result **176** is input to differencer **204** after being delayed for three sample times by delay elements **178**, **180**, and **182** and decimated by a decimating device **192**. The output **214** of differencer **204** (i.e., the result of $Z(M[NT-k]-\Delta)$ when $k=3$) is scaled by a factor a_3 with multiplier **224**. Finally, a delayed and decimated version of result **176** is input to differencer **206** after being delayed for four sample times by delay elements **178**, **180**, **182**, and **184** and decimated by a decimating device **194**. The output **216** of differencer **206** (i.e., the result of $Z(M[NT-k]-\Delta)$ when $k=4$) is scaled by a factor a_4 with multiplier **226**. It will be appreciated by those skilled in the art that the decimating and delay functions performed by elements **178**, **180**, **182**, **184**, **186**, **188**, **190**, **192**, and **194** alternatively may be implemented as a single decimating device coupled to integrator **174** result **176** followed by four

delay elements without departing from the scope and spirit of the present invention.

The scaled differencer outputs from multipliers **218, 220, 222, 224, and 226** preferably are then summed together by summer **228** such that the scaled differencer outputs based on the present sample plus four previous samples are summed together (i.e., a fifth order summation). Finally, the result from summer **228** is provided as the output **146** of the fifth order pre-filter **138**.

The principles described above in reference to a preferred embodiment receiver, as shown in FIGs. **1-3**, may also be applied to a preferred embodiment transmitter in accordance with the present invention, as shown in FIGs. **4-6**. By way of example, like the preceding description of the preferred embodiment receiver, the following description of the preferred embodiment transmitter refers to four channels (i.e., channels **1, 2, 3, and N**). However, as will be appreciated by those skilled in the art the principles described herein may be readily applied to more or less transmitter channels without departing from the scope and spirit of the present invention.

As shown in FIG. **4** **N** inputs **300, 302, 304, and 306** (i.e., **N** digitized channels or **N** digitized information signals **300', 302', 304', and 306'** as shown in graphical form in FIG. **5**) may be input to pre-filters **308, 310, 312, and 314**, respectively (i.e., pre-filters $g_1, g_2, g_3,$ and g_N). These pre-filters **308, 310, 312, and 314** perform baseband filtering (i.e., condition) **N** inputs **300, 302, 304, and 306** with respect to at least one signal characteristic including: optimum selectivity, controlled phase response, and controlled amplitude response. In addition, the pre-filters **308, 310, 312, and 314** interpolate the **N** inputs **300, 302, 304, and 306** to increase the sampling rate such that an subsequent inverse discrete Fourier transformer **324** can operate at a rate higher than that of the **N** inputs **300, 302, 304, and 306** (e.g., the FFTs may operate at a rate which is three times greater than the filter sampling rate). This interpolation may be accomplished by zero-padding the **N** inputs **300, 302, 304, and 306** (e.g. inserting two zero value samples between each sample of the inputs). The pre-filters **308, 310, 312, and 314**, like the receiver post-filters **158, 160, 162, and 164**, preferably are computationally simple to implement. As such, IIR filters may be used, if phase distortion is not critical. In the preferred

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embodiment transmitter structure, the N interpolating pre-filters **308**, **310**, **312**, and **314** preferably are designed with respect to at least one signal characteristic including: optimum selectivity, controlled phase response, and controlled amplitude response. The N outputs **316**, **318**, **320**, and **322** of the pre-filters **308**, **310**, **312**, and **314** preferably are provided to an inverse discrete Fourier transformer **324** which performs an N point inverse Fourier transform (e.g., inverse fast Fourier transform) per input block of pre-filtered samples **316**, **318**, **320**, and **322**. Each of the N outputs **326**, **328**, **330**, and **332** of the inverse discrete Fourier transformer **324** contains samples for one of the N different up converted frequency components.

These N outputs **326**, **328**, **330**, and **332** are post-filtered by filters **334**, **336**, **338**, and **340** (i.e., post-filters f_1 , f_2 , f_3 , and f_N). These post filters operate similar to polyphase filters in known inverse discrete Fourier transform post-filter banks in that they receive their inputs from an inverse discrete Fourier transformer, and run in parallel to form the input to a commutator **350**. One such polyphase filter bank is shown and described in the previously cited related U.S. Patent Application Serial No. 07/966,630 entitled "A Method and Means for Transmultiplexing Signals Between Signal Terminals and Radio Frequency Channels" which was filed on October 22, 1992. However, as will be described in reference to FIGs. **5** and **6**, the preferred embodiment transmitter of the present patent application utilizes modified Cascaded Integrator Comb (CIC) filters **334**, **336**, **338**, and **340** rather than polyphase filters.

These modified CIC post-filters **334**, **336**, **338**, and **340** implement a fifth order function $X(M[NT - k] + \Delta)$ similar to (eq. 7) as described below:

$$X(M[NT - k] + \Delta) = \sum_{k=0}^4 a_k(\Delta) Z^{\Delta}(NT - k)$$

where:

M = the number of IFFT outputs

$$a_k(\Delta) = A(kM - \Delta)$$

$$Z^{\Delta}(NT - k) = \sum_{k_0=0}^{N-1} \sum_{k_1=0}^{N-1} \sum_{k_2=0}^{N-1} \sum_{k_3=0}^{N-1}$$

$$\sum_{k_4=0}^{N-1} [\tilde{w}^{\Delta} X(NT - k - k_0 - k_1 - k_2 - k_3 - k_4)]$$

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$$\tilde{w}^{\Delta}(NT - k) = \begin{cases} w^{\Delta}(T); & k = 0 \\ 0; & k \neq 0 \end{cases} = \text{the upsampled } w^{\Delta}(T)$$

and where $w^{\Delta}(T)$ is the Δ th output of the IDFT at time T such that:

$$w^{\Delta}(T) = \sum_{f=0}^{M-1} \exp(j2\pi\Delta f/M) y_5^f(T) \quad (\text{eq. 8})$$

5 In (eq. 8), $X(M[NT - k] + \Delta)$ is the output of the transmitter, where the Δ 's are an offset in time. Thus, each branch of the inverse FFT corresponds to a different time offset, which means that a commutator can recover the output sequence from the processed inverse FFT branches. In addition, $A(k)$ preferably is the same as was previously
10 described for the preferred embodiment receiver. However, as will be appreciated by those skilled in the art that this $A(k)$ may be modified to change the transmitter's filter response to meet design specifications for the transmitter which differ from the receiver design specifications.

15 The modified fifth order CIC filter expression of (eq. 8) may be implemented in a variety of different means. However, a preferred embodiment fifth order modified CIC filter design is shown in FIG. 6. The fifth order filter design is used to implement each of the parallel modified CIC post-filters **334**, **336**, **338**, and **340** (i.e., f_1 , f_2 , f_3 , and f_N) as shown in FIG. 5 and previously described. By way of example, the
20 modified fifth order CIC filter shown in FIG. 6 will be described as modified CIC filter **340** (f_N); however, each of the modified CIC filters **334**, **336**, **338**, and **340** (i.e., f_1 , f_2 , f_3 , and f_N) in the preferred embodiment communication unit receiver preferably implements a substantially similar modified CIC filter design.

25 The preferred embodiment modified CIC filter **340** receives the N th output **332** of inverse discrete Fourier transformer **324** at an input. This N th output **332** is input to a fifth order differencer **354** (Δ). The differencer **198** implements part of the function within (eq. 8). The preferred embodiment fifth order differencer **354** was previously
30 described in reference to FIG. 8.

Subsequently, the differencer **354** result **356** (i.e., differencer output) preferably is input to a fifth order integrator **362** after being upconverted to a higher sample rate (e.g., seven times the sample rate of the differencer **354** result **356**) by an interpolating device **358**. It will

be appreciated by those skilled in the art that this interpolation of result **356** allows the subsequent parts of (eq. 8) to be implemented within post-filter **340** at a higher sampling rate such that the individual transmit channels can be further separated (i.e., separated by one or more other
5 transmit channels).

The fifth order integrator **362** which integrates bits within the interpolated differencer result **360** over an infinite time period to produce $Z^A(NT - k)$ of (eq. 8). The preferred embodiment fifth order integrator **362** was previously described in reference to FIG. 7.

10 Subsequently, the output **364** of integrator **362** is scaled by a factor a_0 with multiplier **374**. Similarly, the integrator **362** result **364** is scaled by a factor a_1 with multiplier **376** after being delayed for one sample time by delay element **366**. Further, a delayed version of result **364** is scaled by a factor a_2 with multiplier **378** after being delayed for
15 two sample times by delay elements **366** and **368**. Furthermore, a delayed version of result **364** is scaled by a factor a_3 with multiplier **380** after being delayed for three sample times by delay elements **366**, **368**, and **370**. Finally, a delayed version of result **364** is scaled by a factor a_4 with multiplier **382** after being delayed for four sample times by delay
20 elements **366**, **368**, **370**, and **372**.

The scaled integrator outputs from multipliers **374**, **376**, **378**, **380**, and **382** preferably are then summed together by summer **384** such that the scaled integrator outputs based on the present sample plus four previous samples are summed together (i.e., a fifth order
25 summation). Finally, the result from summer **384** is provided as the output of the post-filter **340**.

It will be appreciated by those skilled in the art that the integrating, delay, scaling, and summing functions performed by elements **362**, **366**, **368**, **370**, **372**, **374**, **376**, **378**, **380**, **382**, and
30 **384** alternatively may be implemented by moving the integrator **362** to follow the summer **384** without departing from the scope and spirit of the present invention.

The outputs **342**, **344**, **346**, and **348** (i.e., post-conditioned digitized information signals **342**, **344**, **346**, and **348**) of the respective
35 modified CIC post-filters **334**, **336**, **338**, and **340** form the inputs to a commutator **350**. Commutator **350** operatively couples the modified CIC post-filter outputs **342**, **344**, **346**, and **348** to a digital-to-analog

(D/A) converter **354** through commutator **350** output **352**. The commutator **350** commutates portions of each of the post-conditioned digitized information signals **342**, **344**, **346**, and **348** together to form a composite digitized information signal **352'** (shown in graphical form in FIG. 5) which is output on commutator **350** output **350**.

The composite digitized information signal **352'** is converted by the digital-to-analog (D/A) converter **354** into a composite analog transmission signal **356** (i.e., a baseband electrical signal **356**). Subsequently, the baseband electrical signal **356** is filtered by a low pass filter **358** to eliminate undesired frequency components into a conditioned baseband electrical signal **360**.

The conditioned baseband electrical signal **360** is input to an up converter. This up converter consists of a mixer **364**, a frequency agile local oscillator **362**, and a band pass filter **368**. The conditioned baseband electrical signal **360** is operatively coupled to one input of the mixer **364** and the output of the oscillator **362** is provided to another input of mixer **364**. Mixer **364** mixes the baseband electrical signal **360** up to an IF band electrical signal **366**. Subsequently, the IF band electrical signal **366** is filtered by a band pass filter **368** to eliminate undesired frequency components into a conditioned IF band electrical signal **370**.

The conditioned IF band electrical signal **370** is input to another up converter. The up converter consists of a mixer **374**, a frequency agile local oscillator **372**, and a band pass filter **378**. The electrical signal **370** is operatively coupled to one input of the mixer **374** and the output of the oscillator **372** is provided to another input of mixer **374**. Mixer **374** mixes the IF band electrical signal **370** up to a radio frequency (RF) band electrical signal **376**. Subsequently, the RF band electrical signal **376** is filtered by a band pass filter **378** to eliminate undesired frequency components into a conditioned RF band electrical signal.

Finally, the conditioned RF band electrical signal preferably is output by a device such as antenna **380** which amplifies and radiates the conditioned RF band electrical signal as electromagnetic radiation within the frequency band of interest. It will be appreciated by those skilled in the art that other types of output devices may be used for transmitting or radiating electromagnetic radiation. For example a wave

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guide, a coaxial cable, an optical fiber, or an infrared frequency transducer may be used to transmit electromagnetic radiation.

The present invention may alternatively be described in reference to FIGs 1-6 as follows. Referring now to FIG. 1, a communication receiving unit is shown. The communication receiving unit preferably includes a frequency selector mechanism **102**, **106**, **108**, **112**, **116**, **118**, and **122** for intercepting and selecting a portion of an input frequency band. The frequency selector preferably includes a transducer (not shown) which converts an input frequency band (e.g. received by an antenna **100**) to an electrical signal. This electrical signal is subsequently provided to an input of a filter **102** that selects a portion of an electrical signal **104** which represents intercepted electromagnetic radiation within a particular portion of the input frequency band. In addition, this electrical signal **104** (i.e., which represents the intercepted electromagnetic radiation within a particular portion of the frequency band) may be frequency translated to a different frequency band (e.g., an intermediate frequency (IF) band **110** or baseband frequency **120**) to facilitate signal processing of the electrical signal. This frequency translation may be accomplished through a one, two, or more stage process. For example, the two stage process consists of a first stage in which the selected portion **104** of the electrical signal is input to one input of a mixer **106** and the output of a local oscillator **108** (set at a predetermined IF frequency) is provided to another input of the mixer **106**. The output of the mixer **106** is a down converted electrical signal **110** having frequency components predominately near the predetermined frequency of the local oscillator **108**. This electrical signal **110** may be conditioned to remove undesired frequency components by a bandpass filter **112**. The filter **112** output (i.e., the conditioned IF band electrical signal **112**) may then be input to a second stage.

In the second stage, the electrical signal **114** is down converted by a mixer **118** in conjunction with a local oscillator **116** and subsequently conditioned by a lowpass filter **122** to produce a baseband electrical signal **124**. This baseband electrical signal **124** is then operatively coupled to an analog-to-digital converter **126** which digitizes the baseband electrical signal **124** (that represents the selected portion of an input frequency band). Next, a commutator **130**,

operatively coupled to the analog-to-digital converter **126**, subdivides the digitized portion **128** of the input frequency band into the plurality of digitized signal portions.

5 A discrete Fourier transformer **148**, operatively coupled to the commutator **130**, generates a plurality of digitized channel signals **166**, **168**, **170**, and **172** by discrete Fourier transforming the plurality of digitized signal portions. The discrete Fourier transformer **148** preferably includes signal conditioners **132**, **134**, **136**, and **138** for pre-conditioning the plurality of digitized signal portions prior to the
10 plurality of digitized signal portions being discrete Fourier transformed into the plurality of digitized channel signals **166**, **168**, **170**, and **172**. The signal conditioners **132**, **134**, **136**, and **138** may be modified cascaded integrator comb (CIC) filters or polyphase filters as described above. The pre-conditioning performed by the signal conditioners **132**,
15 **134**, **136**, and **138** preferably consists of filtering each digitized signal portion with respect to a signal characteristic (e.g., optimum selectivity, controlled phase response, or controlled amplitude response).

Finally, detectors, operatively coupled to the discrete Fourier transformer **148**, detect at least two information signals from the plurality
20 of digitized channel signals **166**, **168**, **170**, and **172**. Each information signal preferably conforms with a signal coding and channelization standard such as frequency division multiple access, time division multiple access, or frequency hopping code division multiple access.

Referring now to FIG. 4, a communication transmitting unit is
25 shown. The communication transmitting unit preferably includes an inverse discrete Fourier transformer **324** for generating a plurality of digitized signals **342**, **344**, **346**, and **348** by inverse discrete Fourier transforming a plurality of input digitized information signals **300**, **302**, **304**, and **306**. Each information signal preferably conforms with a
30 signal coding and channelization standard such as frequency division multiple access, time division multiple access, or frequency hopping code division multiple access. The inverse discrete Fourier transformer **324** preferably includes signal conditioners **334**, **336**, **338**, and **340** for post-conditioning the plurality of digitized signals **342**, **344**, **346**,
35 and **348** after the plurality of input digitized information signals **300**, **302**, **304**, and **306** have been inverse discrete Fourier transformed into the plurality of digitized signals **342**, **344**, **346**, and **348**. The signal

conditioners **334**, **336**, **338**, and **340** may be modified cascaded integrator comb (CIC) filters or polyphase filters as described above. The post-conditioning performed by the signal conditioners **334**, **336**, **338**, and **340** preferably consists of filtering each digitized signal with
5 respect to a signal characteristic (e.g., optimum selectivity, controlled phase response, or controlled amplitude response).

A commutator **350**, operatively coupled to the inverse discrete Fourier transformer **324**, combines portions of the plurality of digitized signals **342**, **344**, **346**, and **348** into a composite digitized signal **352**.
10 An analog-to-digital converter **354**, operatively coupled to the commutator **350**, generates a composite analog transmission signal **356** from the composite digitized signal **352**.

The communication transmitting unit preferably includes a frequency selector mechanism **358**, **362**, **364**, **368**, **372**, **374**, and
15 **378**, operatively coupled to the analog-to-digital converter **354**, for selecting a portion of a frequency band that the composite analog transmission signal **356** is to be transmitted. The composite analog transmission signal **356** preferably is provided to an input of a filter **358** that selects the portion **360** of the composite analog transmission signal
20 **356** which is to be transmitted. This selected portion **360** of composite analog transmission signal **356** may be frequency translated to a different frequency band (e.g., an intermediate frequency (IF) band **366** or a radio frequency (RF) band **376**) to facilitate transmission. This frequency translation may be accomplished through a one, two, or more
25 stage process. For example, the two stage process consists of a first stage in which the selected portion **360** is input to one input of a mixer **364** and the output of a local oscillator **362** (set at a predetermined IF frequency) is provided to another input of the mixer **364**. The output of the mixer **364** is an up converted composite analog transmission signal
30 **366** having frequency components predominately near the predetermined frequency of the local oscillator **362**. This composite analog transmission signal **366** may be conditioned to remove undesired frequency components by a bandpass filter **368**. The filter **368** output (i.e., the conditioned IF band analog transmission signal
35 **370**) may then be input to a second stage. In the second stage, the analog transmission signal **370** is up converted by a mixer **374** in conjunction with a local oscillator **372** and subsequently conditioned by

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a bandpass filter **378** to produce an RF band analog transmission signal. Finally, a transmitting mechanism **380**, operatively coupled to the frequency selector mechanism **358, 362, 364, 368, 372, 374,** and **378**, transmits the RF band analog transmission signal over a frequency
5 band.

Although the invention has been described and illustrated with a certain degree of particularity, it is understood that the present disclosure of embodiments has been made by way of example only and that numerous changes in the arrangement and combination of parts as
10 well as steps may be resorted to by those skilled in the art without departing from the spirit and scope of the invention as claimed.

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Claims

What is claimed is:

- 5 1. A filter comprising a plurality of conditioning means for conditioning a digitized signal during digital signal processing, each conditioning means comprising an integrator, at least one differencer, and at least one multiplier operatively coupled together to condition a portion of the digitized signal.
- 10 2. The filter of claim 1 wherein:
- 15 (a) the at least one differencer of each conditioning means comprises a first and a second differencer, each operatively coupled to an output of the integrator, the second differencer being operatively coupled through a delay element;
- 20 (b) the at least one multiplier of each conditioning means comprises a first and a second multiplier, operatively coupled to the first and the second differencer, respectively, for generating scaled outputs of the first and the second differencer; and
- 25 (c) each conditioning means further comprises a summing means, operatively coupled to the first and the second multiplier, for generating a portion of the conditioned digitized signal from the scaled outputs of the first and the second multiplier.

3. The filter of claim 2 wherein:
- (a) each conditioning means further comprises:
 - (i) a third differencer operatively coupled to an output of the integrator through at least two delay elements;
5 and
 - (ii) a third multiplier, operatively coupled to the third differencer, for generating a scaled output of the third differencer; and
 - (b) the summing means is further operatively coupled to the
10 third multiplier, for generating the portion of the conditioned digitized signal from the scaled output of the third multiplier.
4. The filter of claim 1 wherein:
- (a) the integrator of each conditioning means is operatively
15 coupled to an output of the differencer of each conditioning means;
 - (b) the at least one multiplier of each conditioning means comprises a first and a second multiplier, operatively
20 coupled to the integrator, for generating scaled outputs of the integrator, the second multiplier being operatively coupled through a delay element; and
 - (c) each conditioning means further comprises a summing
25 means, operatively coupled to the first and the second multiplier, for generating a portion of the conditioned digitized signal from the scaled outputs of the first and the second multiplier.

-30-

5. The filter of claim 4 wherein:
- (a) each conditioning means further comprises a third multiplier, operatively coupled to an output of the integrator through at least two delay elements, for generating a scaled output of the integrator; and
 - (b) the summing means is further operatively coupled to the third multiplier, for generating the portion of the conditioned digitized signal from the scaled output of the third multiplier.
6. A communication receiving unit, comprising:
- (a) a plurality of filtering means for pre-conditioning a digitized signal for subsequent Fourier transformation each filtering means comprising an integrator, at least one differencer, and at least one multiplier operatively coupled together to pre-condition a portion of the digitized signal; and
 - (b) Fourier transforming means, operatively coupled to the plurality of filtering means, for generating a digitized channel signal.
7. The communication unit of claim 6 further comprising:
- (a) input means for intercepting electromagnetic radiation within a frequency band; and
 - (b) digitizing means, operatively coupled to the input means, for digitizing a portion of the intercepted electromagnetic radiation into the digitized signal.
8. The communication unit of claim 6 further comprising conditioning means, operatively coupled to the Fourier transforming means, for removing undesired signals from the digitized channel signal by filtering the digitized channel signal with respect to a signal characteristic selected from the group consisting of optimum selectivity, controlled phase response, and controlled amplitude response.

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9. The communication unit of claim 6 further comprising a detecting means, operatively coupled to the Fourier transforming means, for detecting an information signal within the digitized channel signal.
- 5
10. A communication transmitting unit, comprising:
- (a) inverse Fourier transforming means for generating a composite digitized signal from a plurality of digitized information signals to be transmitted; and
- 10 (b) a plurality of filtering means, operatively coupled to the inverse Fourier transforming means, for post-conditioning the composite digitized signal for subsequent transmission, each filtering means comprising an integrator, a differencer, and at least one multiplier
- 15 operatively coupled together to post-condition a portion of the digitized signal.
11. The communication unit of claim 10 further comprising an input means, operatively coupled to the inverse Fourier transforming means, for receiving the plurality of digitized information signals to be transmitted and subsequently providing the plurality of digitized information signals to the inverse Fourier transforming means.
- 20
12. The communication unit of claim 10 further comprising a conditioning means, operatively coupled to the inverse Fourier transforming means, for removing undesired signals from the plurality of digitized information signals prior to the generation of the composite digitized signal by filtering each digitized information signal with respect to a signal characteristic selected from the group consisting of optimum selectivity, controlled phase response, and controlled amplitude response.
- 25
- 30

13. A communication receiving unit, comprising:
- (a) an analog-to-digital conversion means for digitizing a selected portion of an input frequency band;
 - (b) commutating means, operatively coupled to the analog-to-digital conversion means, for subdividing the digitized portion of the input frequency band into the plurality of digitized signal portions;
 - (c) discrete Fourier transforming means, operatively coupled to the commutating means, for generating a plurality of digitized channel signals by discrete Fourier transforming the plurality of digitized signal portions, the discrete Fourier transforming means comprising conditioning means for pre-conditioning the plurality of digitized signal portions with a modified cascaded integrator comb filter; and
 - (d) a detecting means, operatively coupled to the discrete Fourier transforming means, for detecting at least two information signals from the plurality of digitized channel signals.
14. The communication unit of claim 13 wherein the conditioning means comprises means for filtering each digitized signal portion with respect to a signal characteristic selected from the group consisting of optimum selectivity, controlled phase response, and controlled amplitude response.
15. The communication unit of claim 13 wherein each information signal conforms with a signal coding and channelization standard selected from the group consisting of frequency division multiple access, time division multiple access, and frequency hopping code division multiple access.

16. A communication transmitting unit, comprising:
- 5 (a) inverse discrete Fourier transforming means for generating a plurality of digitized signals by inverse discrete Fourier transforming a plurality of input digitized information signals, the inverse discrete Fourier transforming means comprising conditioning means for post-conditioning the plurality of digitized signals with a modified cascaded integrator comb filter;
- 10 (b) commutating means, operatively coupled to the inverse discrete Fourier transforming means, for combining portions of the plurality of digitized signals into a composite digitized signal;
- 15 (c) digital-to-analog conversion means, operatively coupled to the commutating means, for generating a composite analog transmission signal from the composite digitized signal; and
- 20 (d) transmitting means, operatively coupled to the digital-to-analog conversion means, for transmitting the composite analog transmission signal over a frequency band.
17. The communication unit of claim 16 wherein each information signal conforms with a signal coding and channelization standard selected from the group consisting of frequency division multiple access, time division multiple access, and frequency hopping code division multiple access.
- 25
18. The communication unit of claim 16 wherein the conditioning means comprises means for filtering each digitized signal with respect to a signal characteristic selected from the group consisting of optimum selectivity, controlled phase response, and controlled amplitude response.
- 30

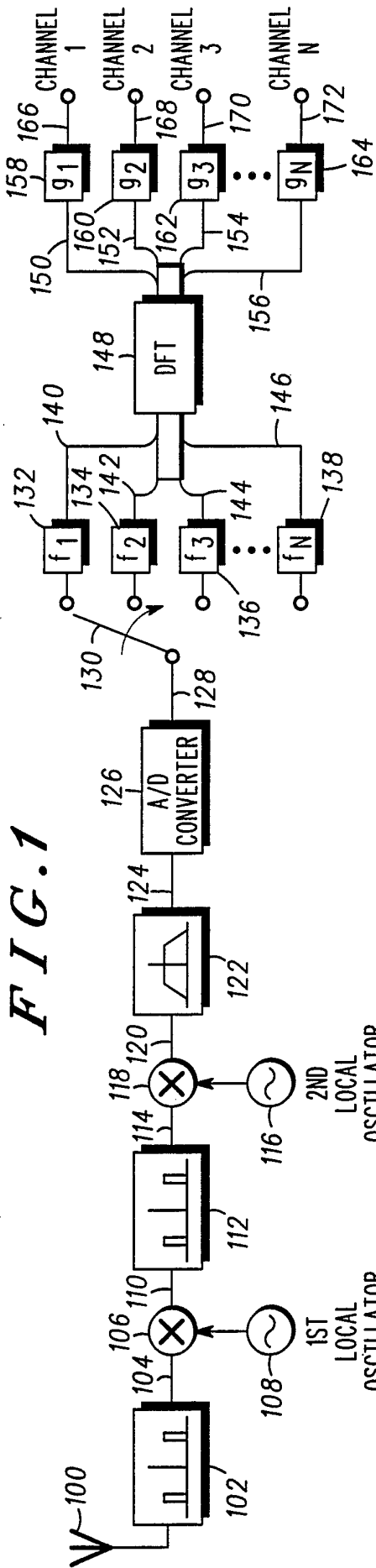


FIG. 1

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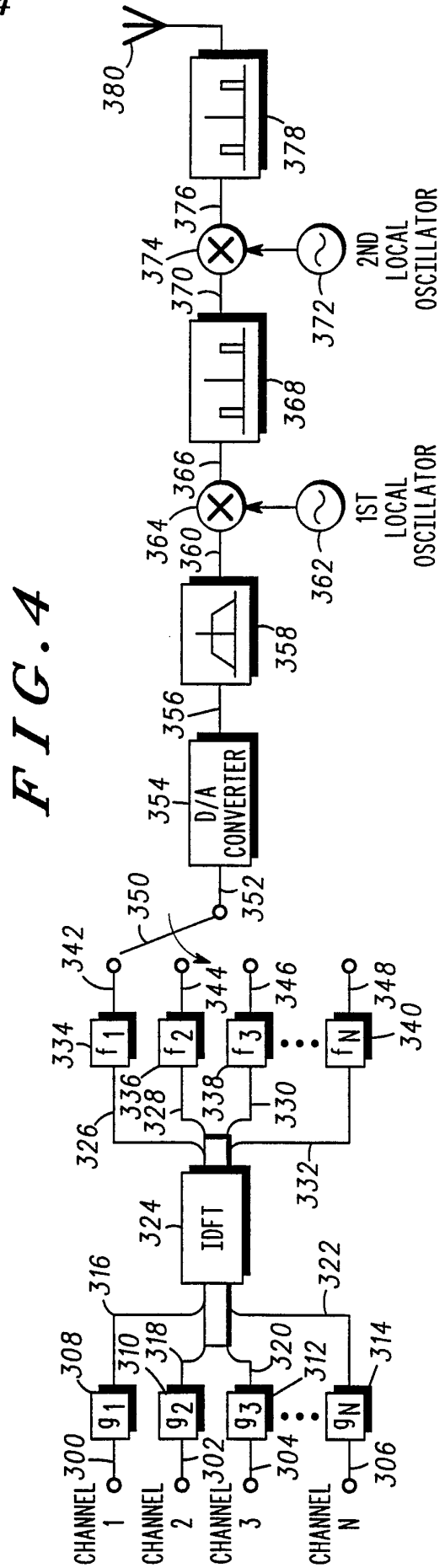
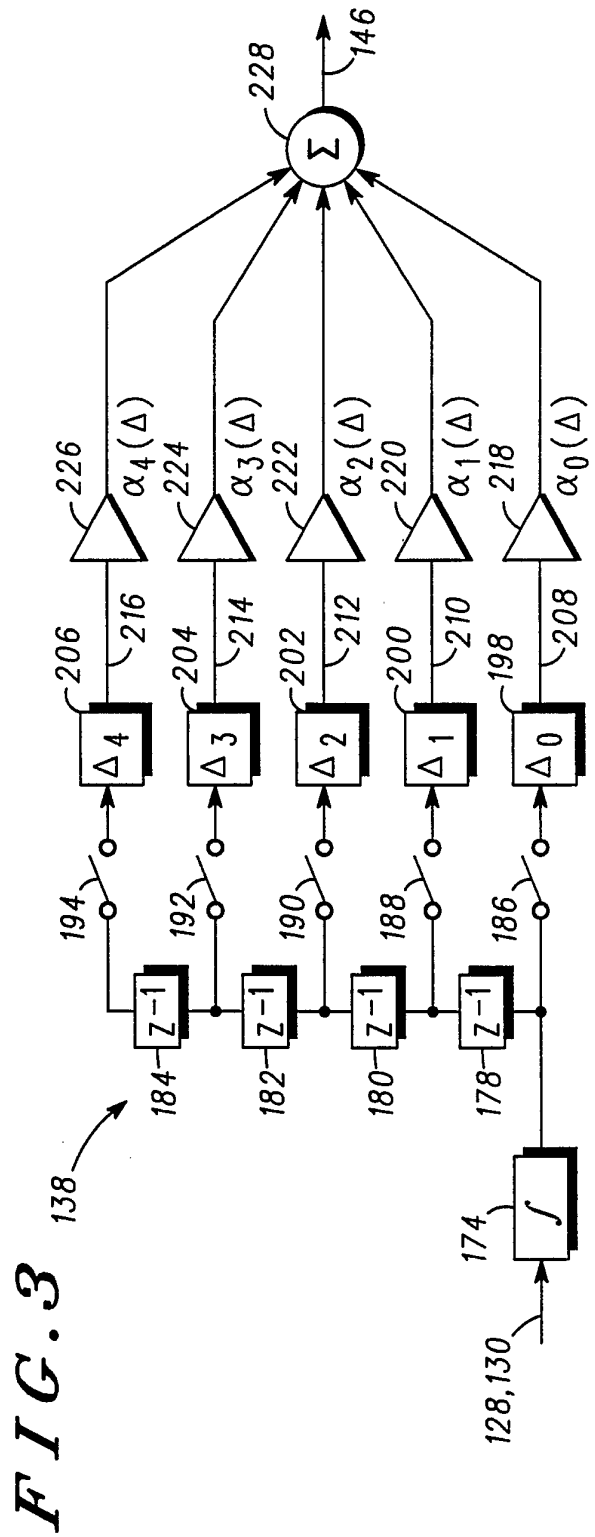
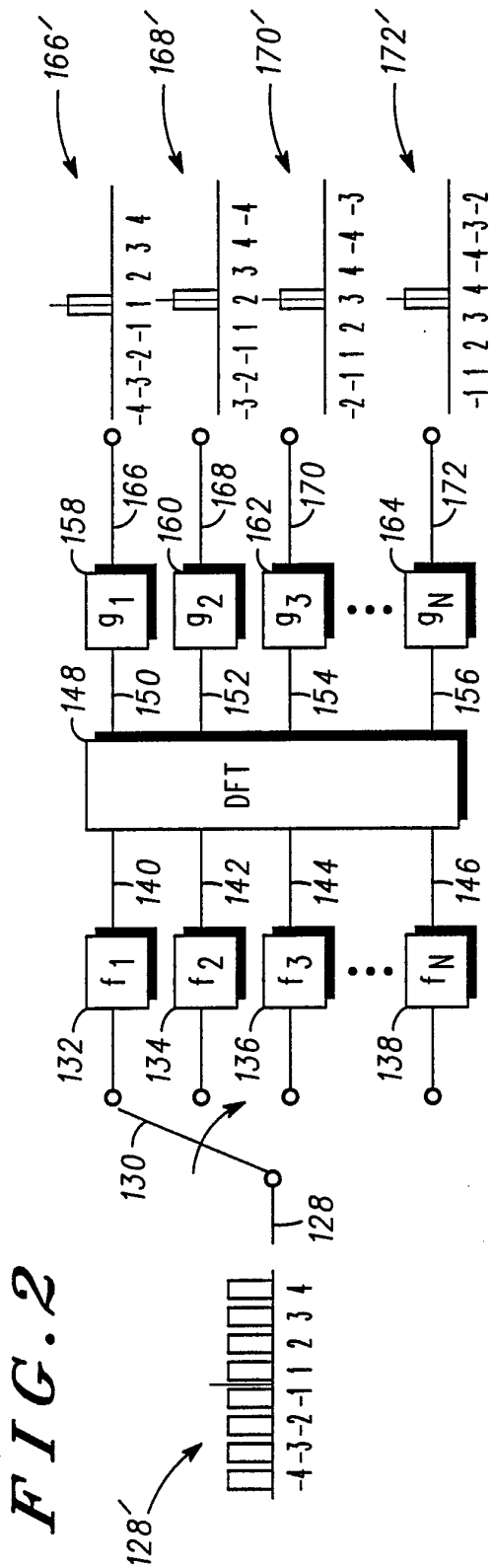


FIG. 4



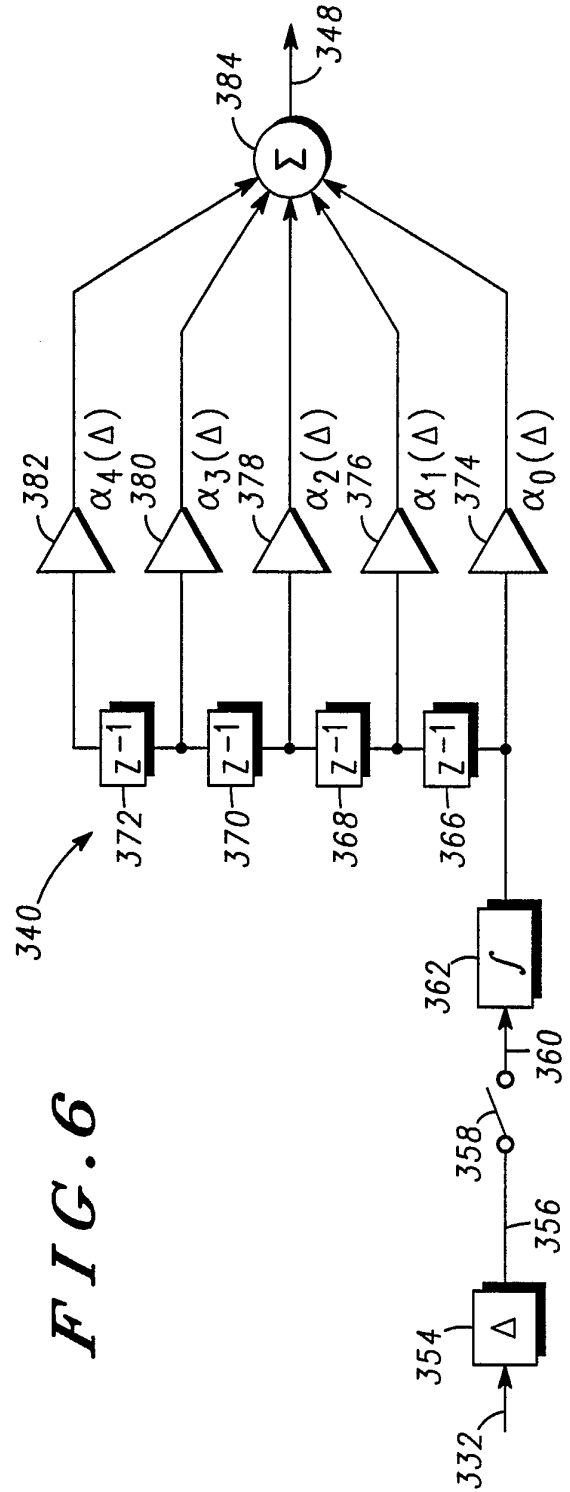
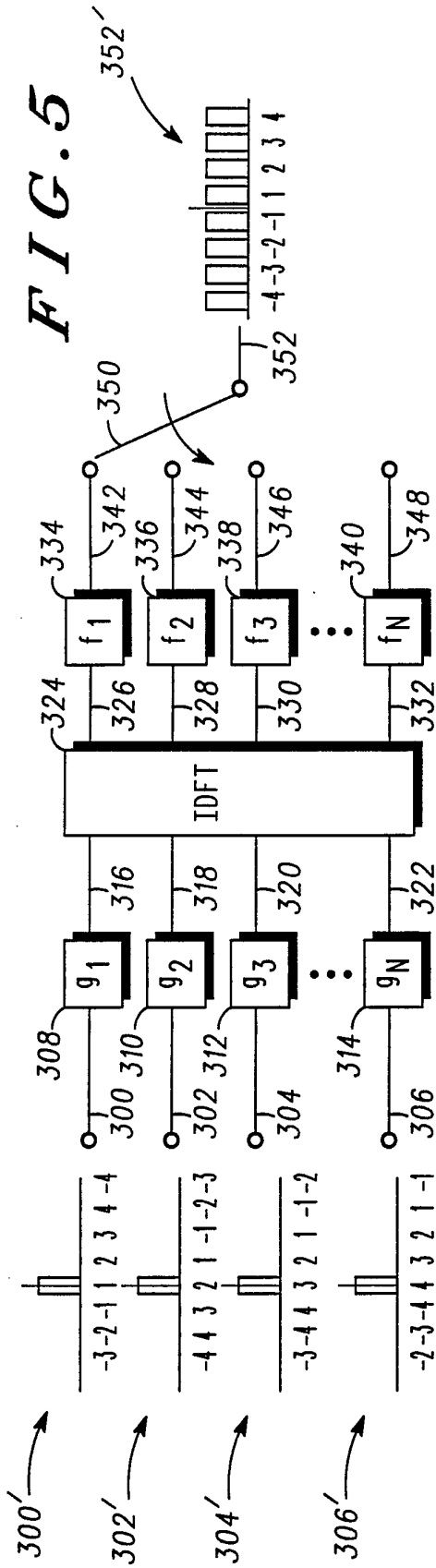


FIG. 7

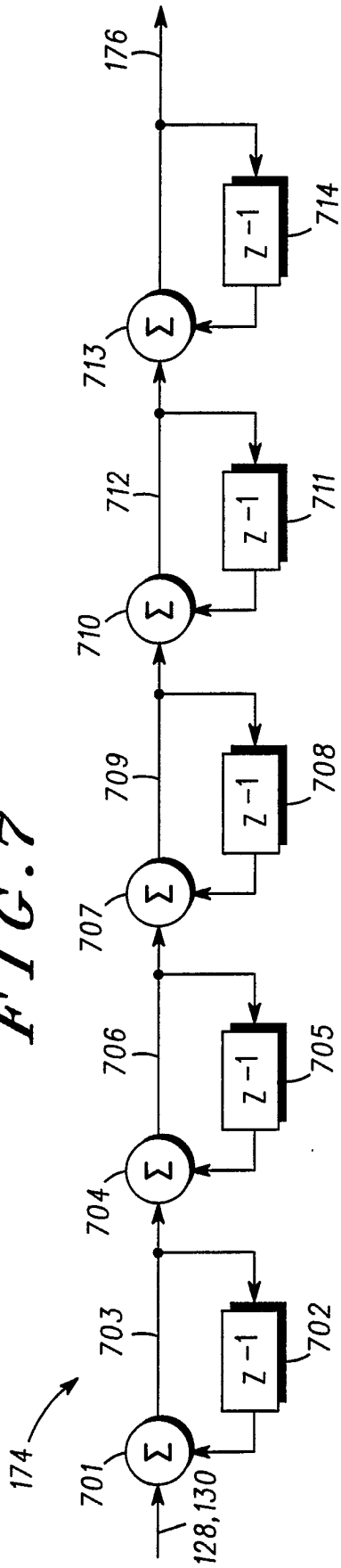
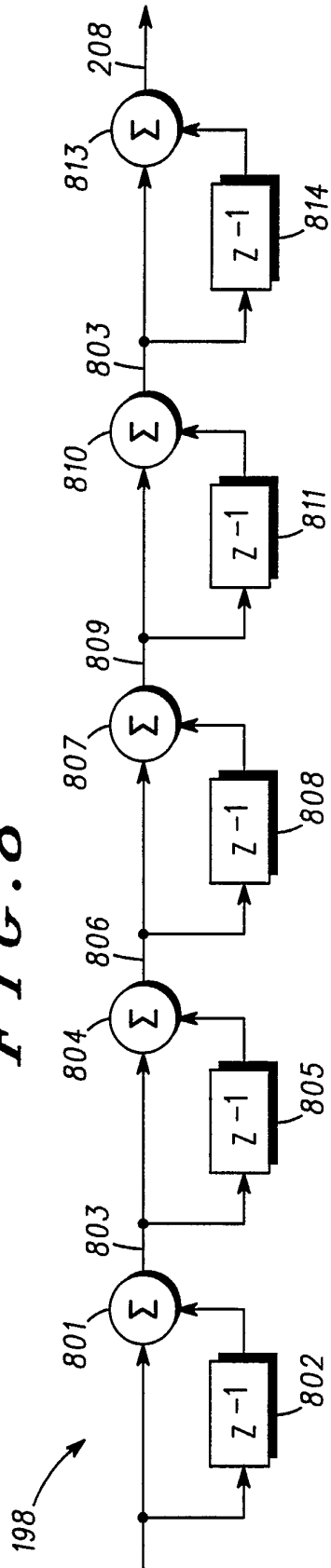


FIG. 8



INTERNATIONAL SEARCH REPORT

International application No.
PCT/US94/01338

A. CLASSIFICATION OF SUBJECT MATTER

IPC(5) :H04J 1/02

US CL :370/70; 364/572,576,724.01,724.1,724.13; 375/102,103; 455/307

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

B. FIELDS SEARCHED

Minimum documentation searched (classification system followed by classification symbols)

U.S. : 370/70; 364/572,576,724.01,724.1,724.13; 375/102,103; 455/307

375/60,99; 455/63,303,306; 364/574,577,724.16,724.17

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

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

Please See Extra Sheet.

C. DOCUMENTS CONSIDERED TO BE RELEVANT

| Category* | Citation of document, with indication, where appropriate, of the relevant passages | Relevant to claim No. |
|-----------|--|-----------------------|
| A | US, A, 4,112,374 (STEINBRECHER) 05 September 1978 | 1-18 |
| A | US, A, 4,230,956 (STEINBRECHER) 28 October 1980 | 1-18 |
| A | US, A, 4,785,447 (ICHIYOSHI) 15 November 1988 | 1-18 |
| A | US, A, 4,876,542 (VAN BAVEL et al.) 24 October 1989 | 1-18 |
| A | US, A, 5,043,933 (BOUTAUD et al.) 27 August 1991 | 1-18 |

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

| | |
|---|--|
| * Special categories of cited documents: | "T" later document published after the international filing date or priority date and not in conflict with the application but cited to understand the principle or theory underlying the invention |
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| "P" document published prior to the international filing date but later than the priority date claimed | |

Date of the actual completion of the international search

08 JUNE 1994

Date of mailing of the international search report

JUL 08 1994

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INTERNATIONAL SEARCH REPORT

International application No.
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C (Continuation). DOCUMENTS CONSIDERED TO BE RELEVANT

| Category* | Citation of document, with indication, where appropriate, of the relevant passages | Relevant to claim No. |
|-----------|---|-----------------------|
| A | Multirate Digital Signal Processing, CROCHIERE, R. et al., 1983, "Chapter Seven: Multirate Techniques in Filter Banks and Spectrum Analyzers and Synthesizers", Prentice-Hall, Englewood Cliffs, NJ, USA, pgs. 291-313. | 1-18 |
| A | IEEE Transactions on Acoustics, Speech and Signal Processing, Vol. ASSP-29, No. 2. HOGENAUER, E., April 1981, "An Economical Class of Digital Filters for Decimation and Interpolation", pgs. 155-162. | 1-18 |
| A | Harris Semiconductor Digital Signal Processing Databook, HSP45116, Harris Corporation 1992, "Numerically Controlled Oscillator/Modulator", pgs. 5-26 through 5-40. | 1-18 |

INTERNATIONAL SEARCH REPORT

International application No.

PCT/US94/01338

B. FIELDS SEARCHED

Electronic data bases consulted (Name of data base and where practicable terms used):

APS

search terms: cascaded integrated comb, polyphase filter, dft, fft, fourier, integrator, multiplier, differencer, transmitter, receiver, transceiver