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(54) SNR ESTIMATION USING FILTERS

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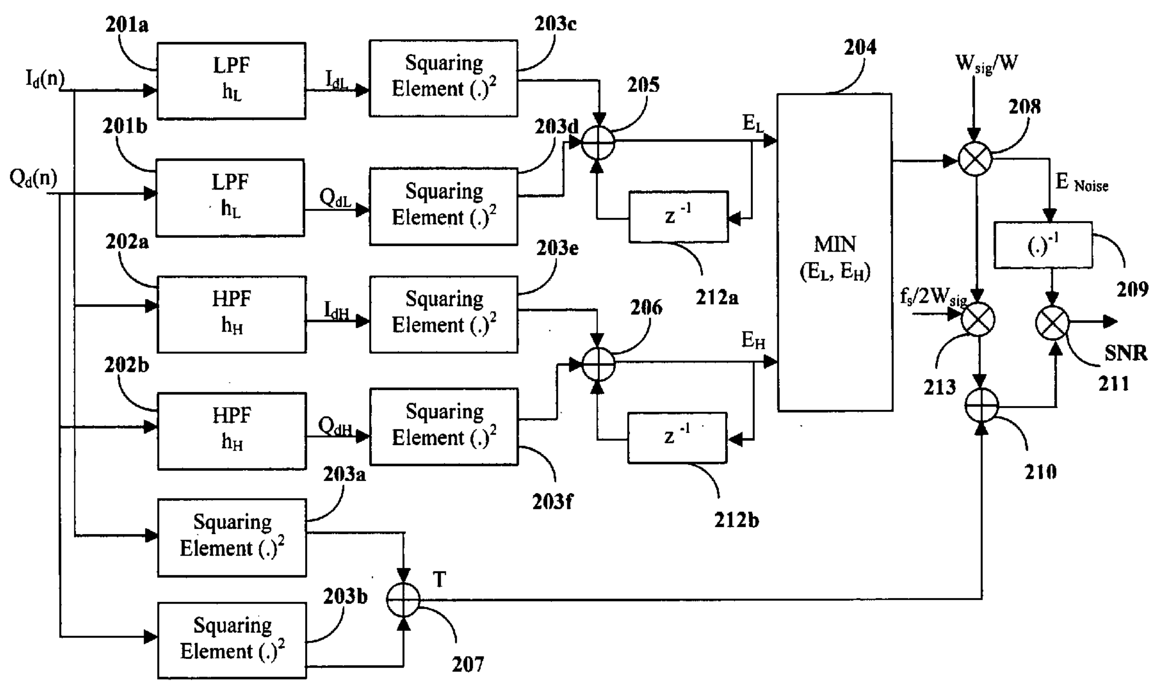
(57) ABSTRACT

The present invention provides a method of accurately determining the signal to noise ratio (SNR) using filters. Three embodiments of the invention are disclosed: the use of fixed filters, multiple filters and dynamic filters. The total noise energy is calculated by applying low pass and high pass filters. The minimum of the two noise energies estimated by the low pass and high pass filters is selected to calculate the total noise energy in the signal. The present invention provides an accurate method of SNR estimation in conditions of low SNR and high carrier offsets, without the requirement for bringing the signal to base band. The use of multiple filters provides accurate SNR measurement even in the presence of discrete interferences.



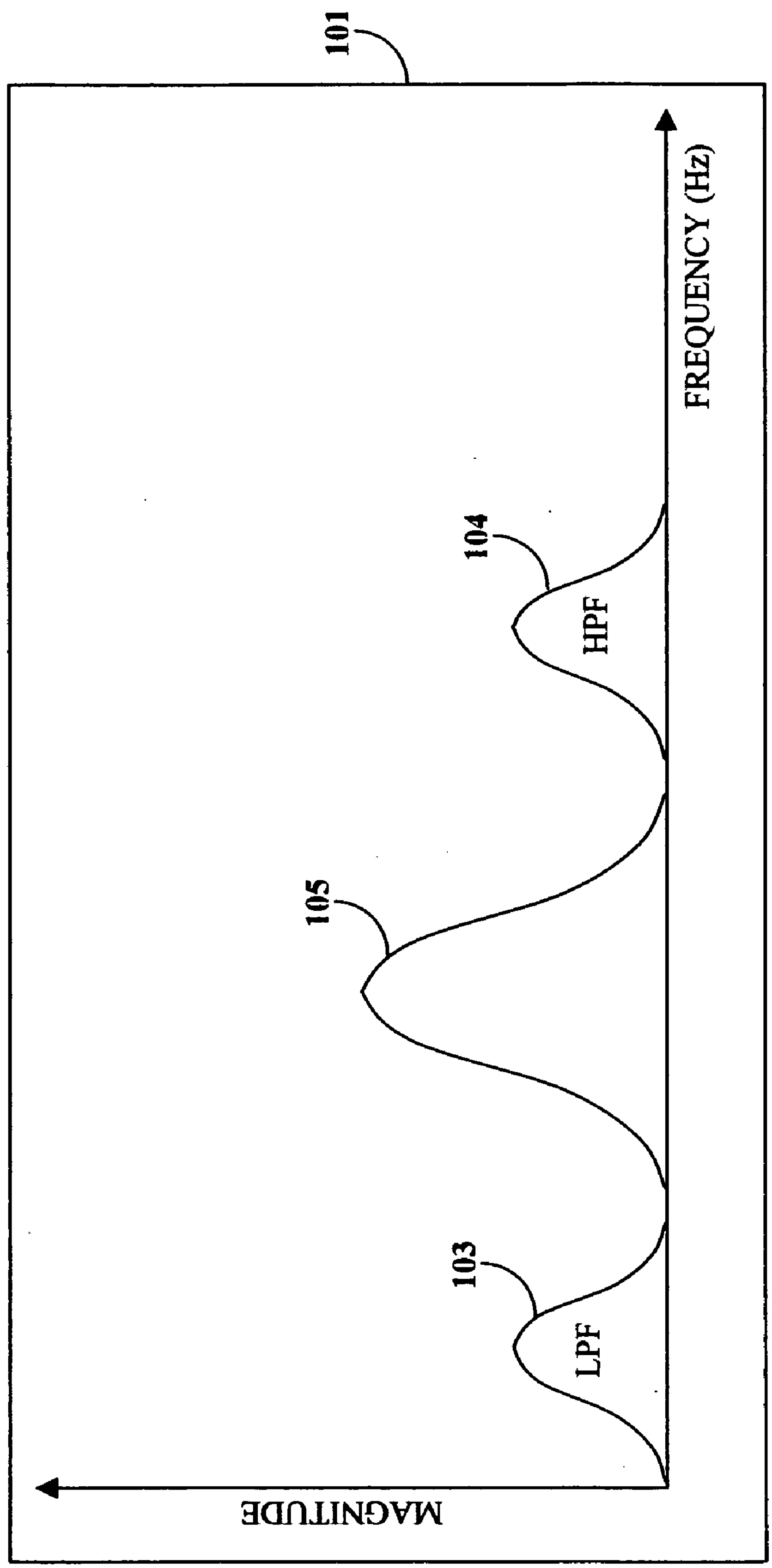



FIGURE 1A

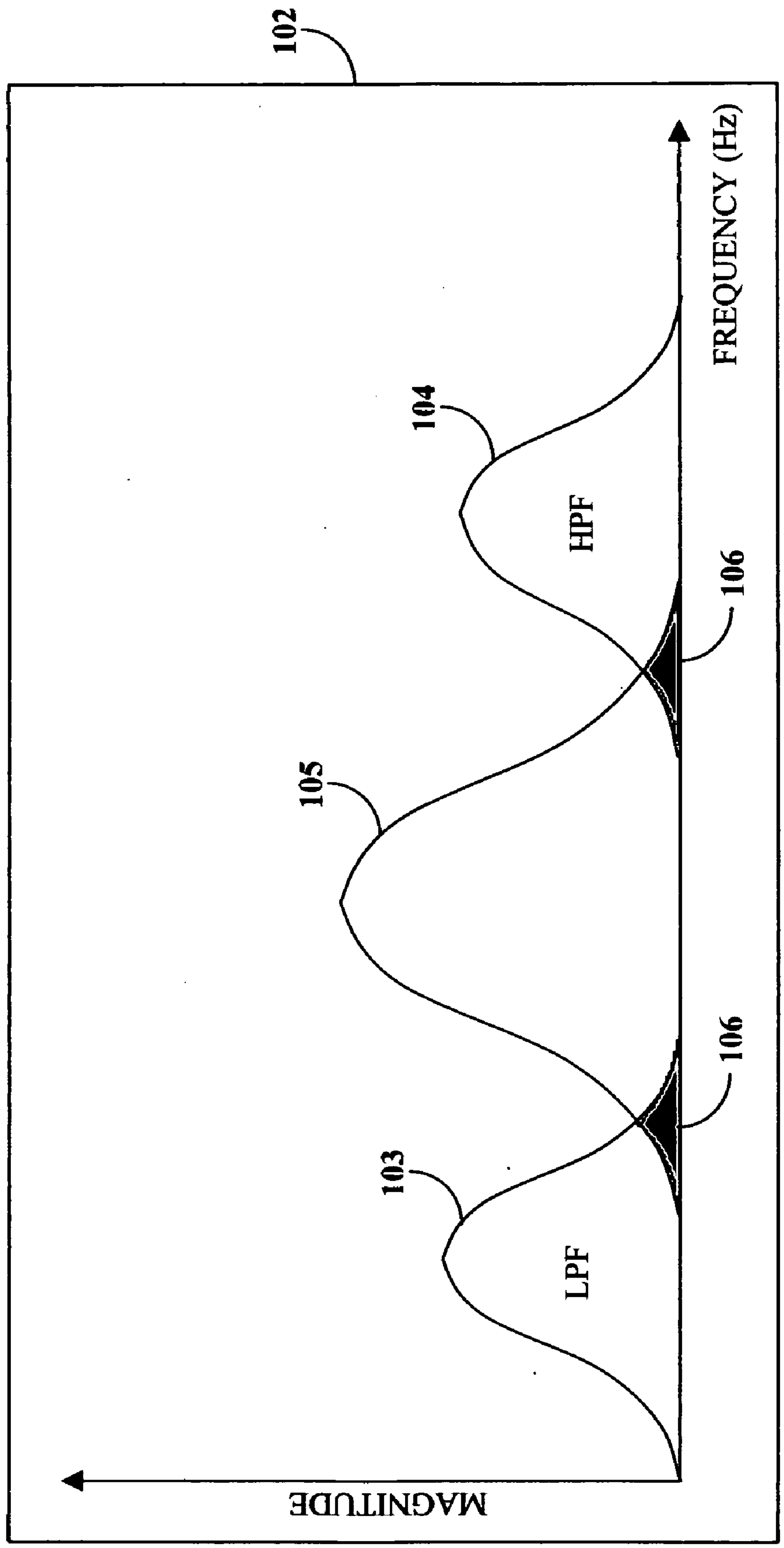



FIGURE 1B

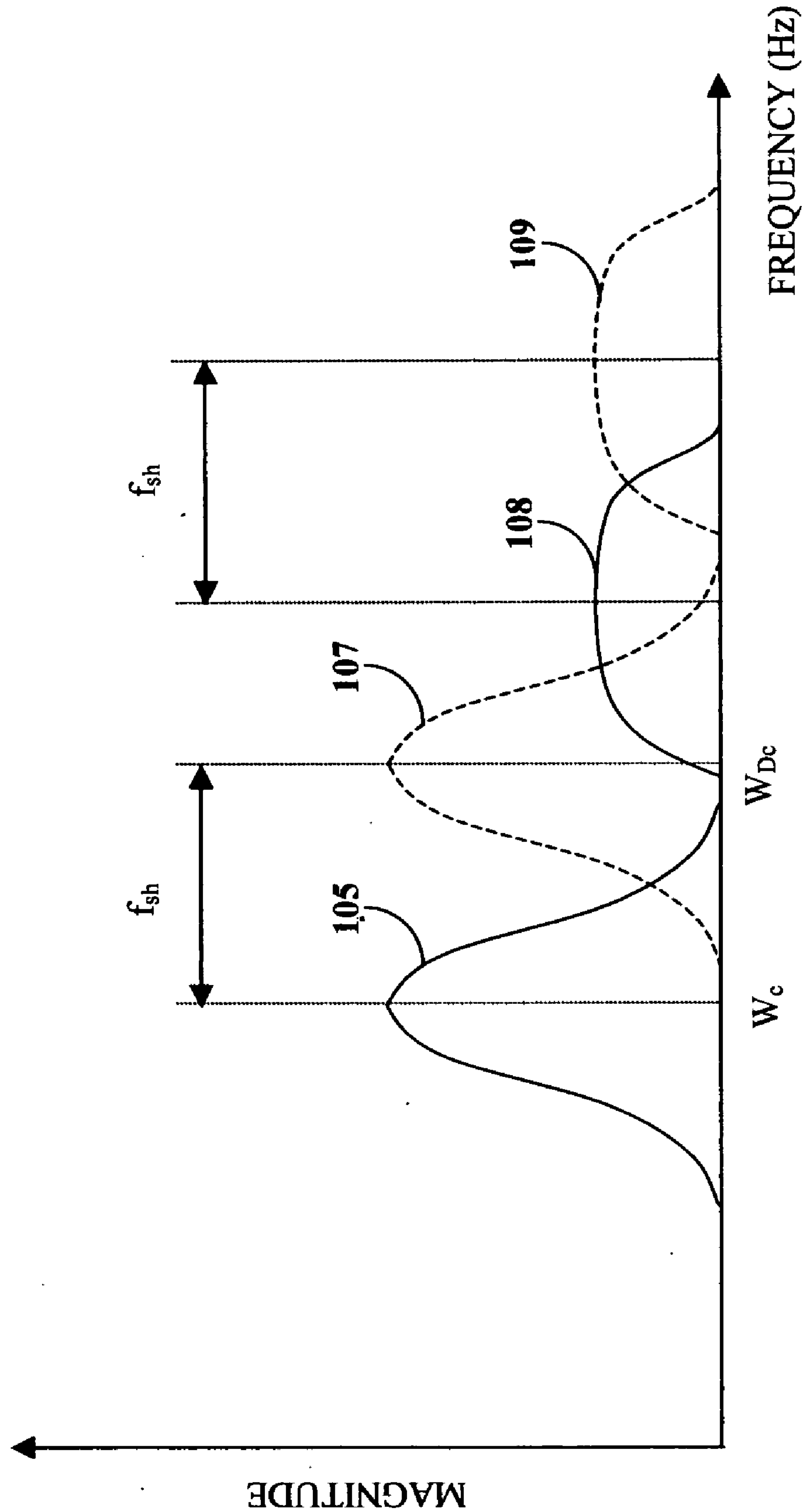



FIGURE 1C

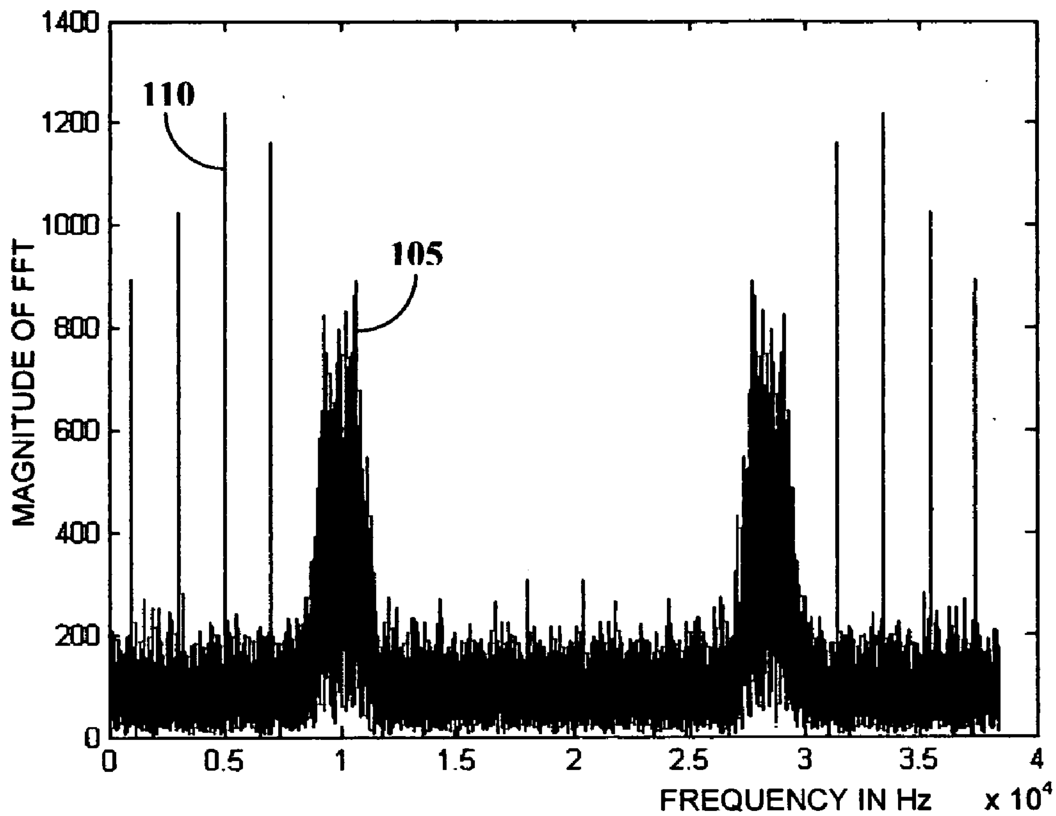

1400
1200
1000
800
600
400
200
0
MAGNITUDE OF FFT
110
105
0
0.5
1
1.5
2
2.5
3
3.5
4
FREQUENCY IN Hz
x 10^4

FIGURE 1D

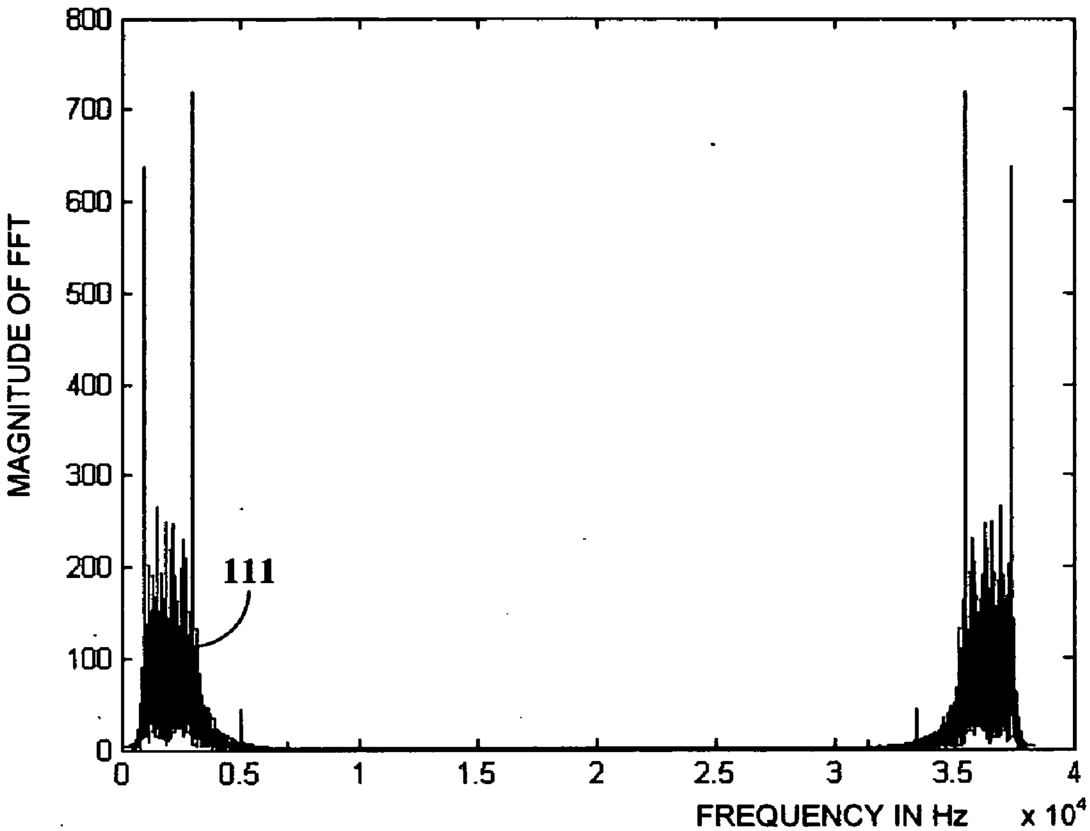

800
700
600
500
400
300
200
100
0
MAGNITUDE OF FFT
111
0
0.5
1
1.5
2
2.5
3
3.5
4
FREQUENCY IN Hz
x 10^4

FIGURE 1E

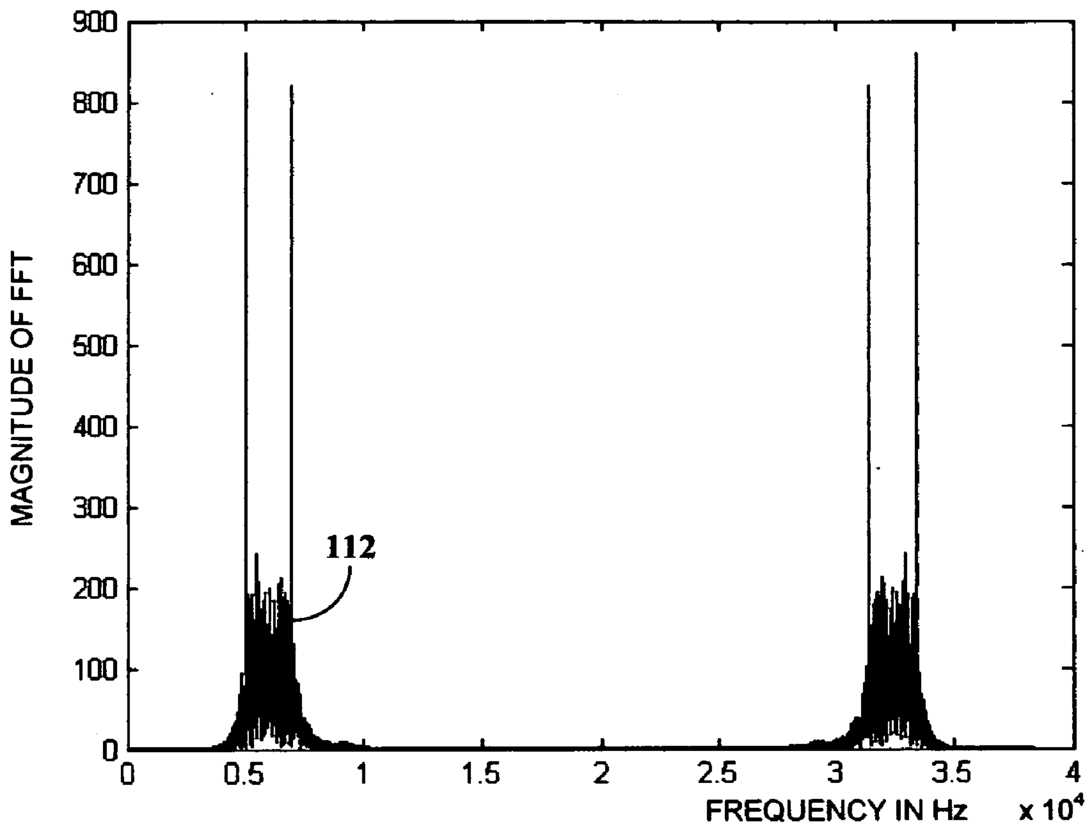

900
800
700
600
500
400
300
200
100
0
MAGNITUDE OF FFT
112
0
0.5
1
1.5
2
2.5
3
3.5
4
FREQUENCY IN Hz
x 10^4

FIGURE 1F

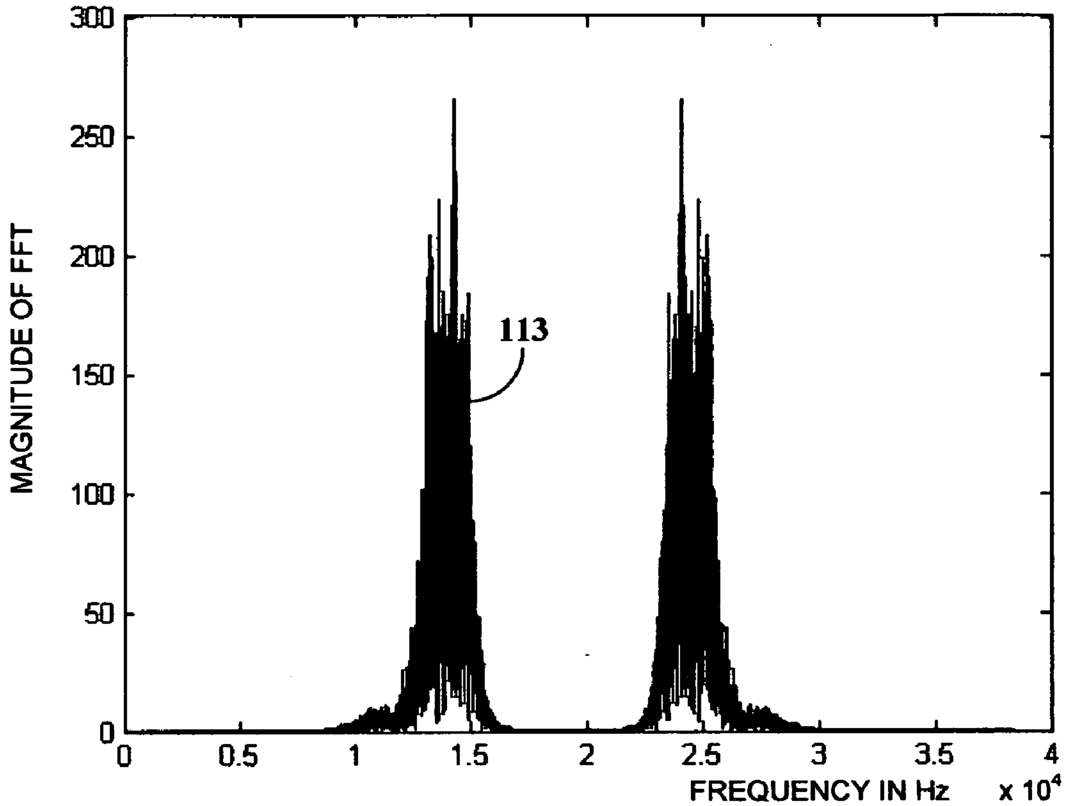

MAGNITUDE OF FFT
300
250
200
150
100
50
0
113
0
0.5
1
1.5
2
2.5
3
3.5
4
FREQUENCY IN Hz
x 10^4

FIGURE 1G

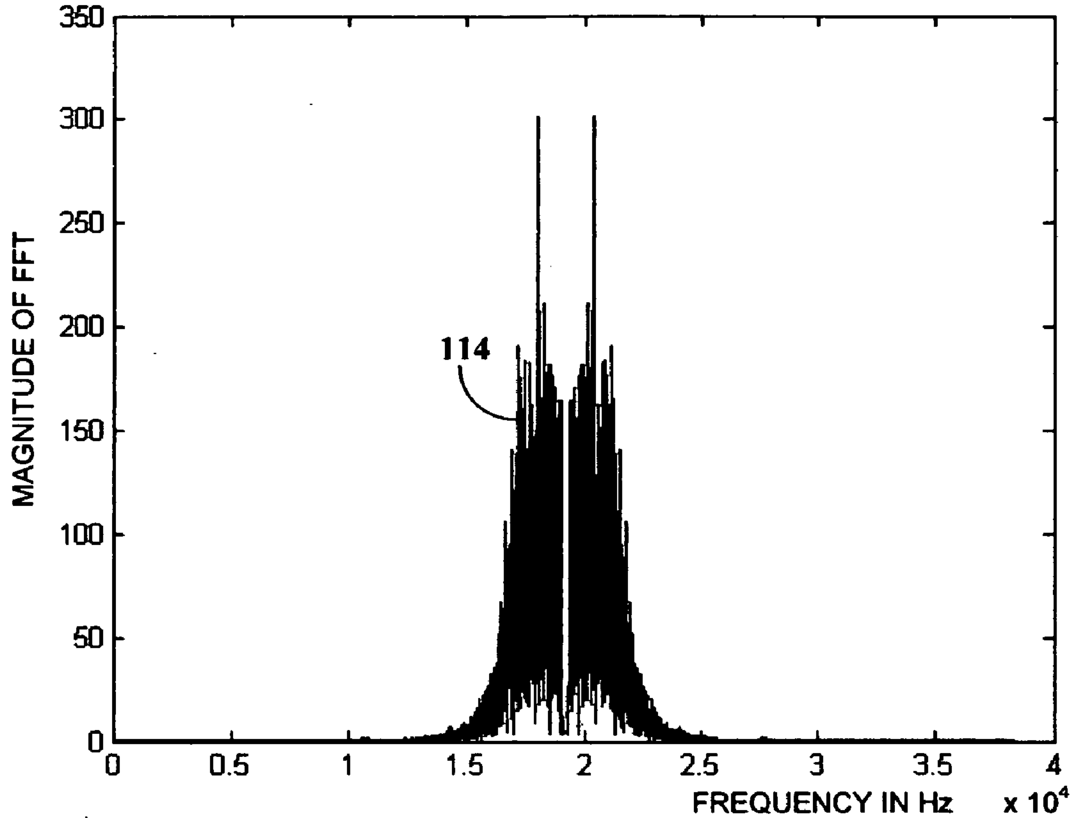

350
300
250
200
150
100
50
0
MAGNITUDE OF FFT
114
0
0.5
1
1.5
2
2.5
3
3.5
4
FREQUENCY IN Hz
x 10^4

FIGURE 1H

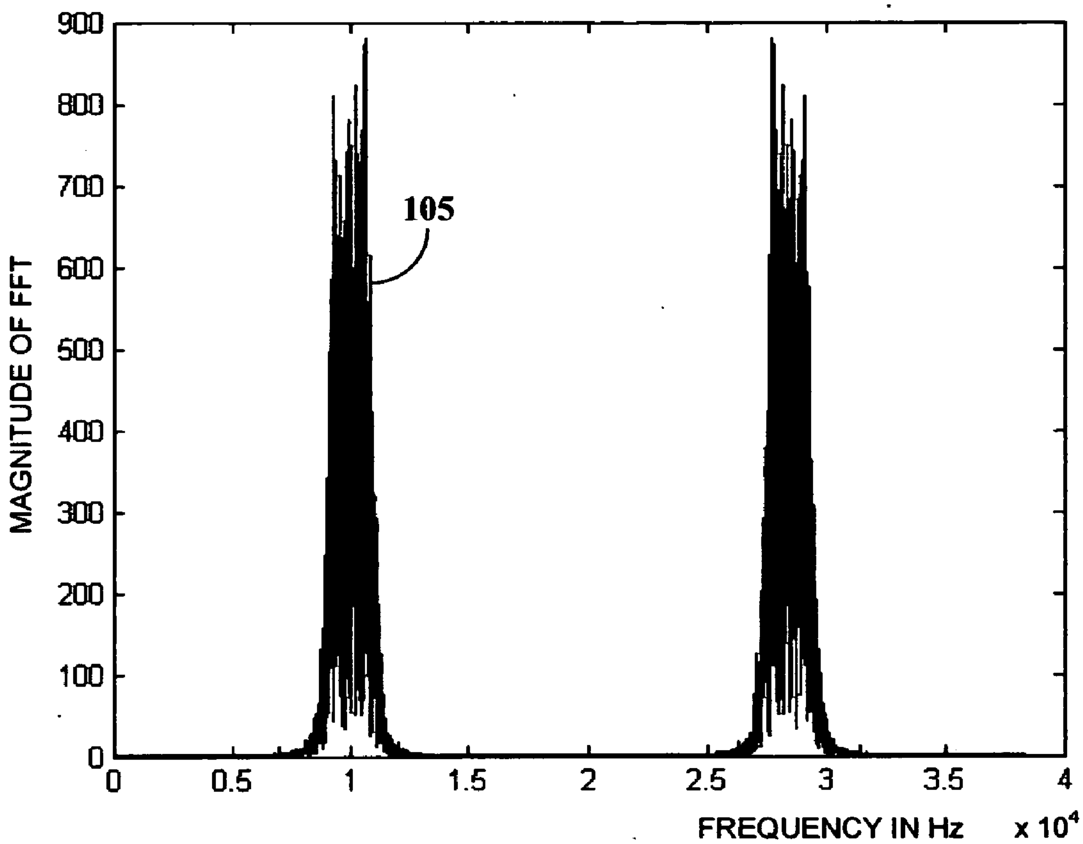

900
800
700
600
500
400
300
200
100
0
MAGNITUDE OF FFT
105
0
0.5
1
1.5
2
2.5
3
3.5
4
FREQUENCY IN Hz
x 10^4

FIGURE 1 I

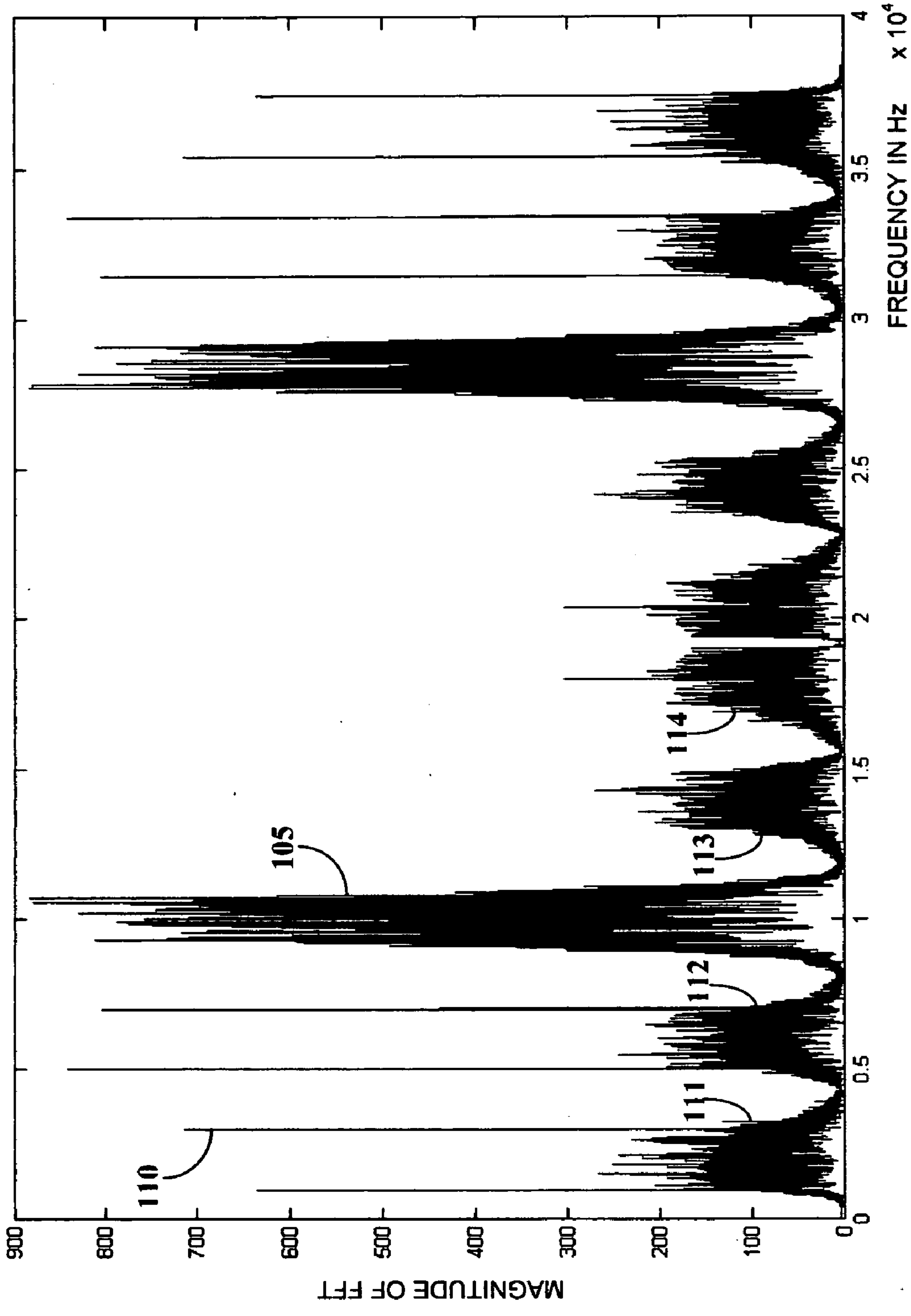
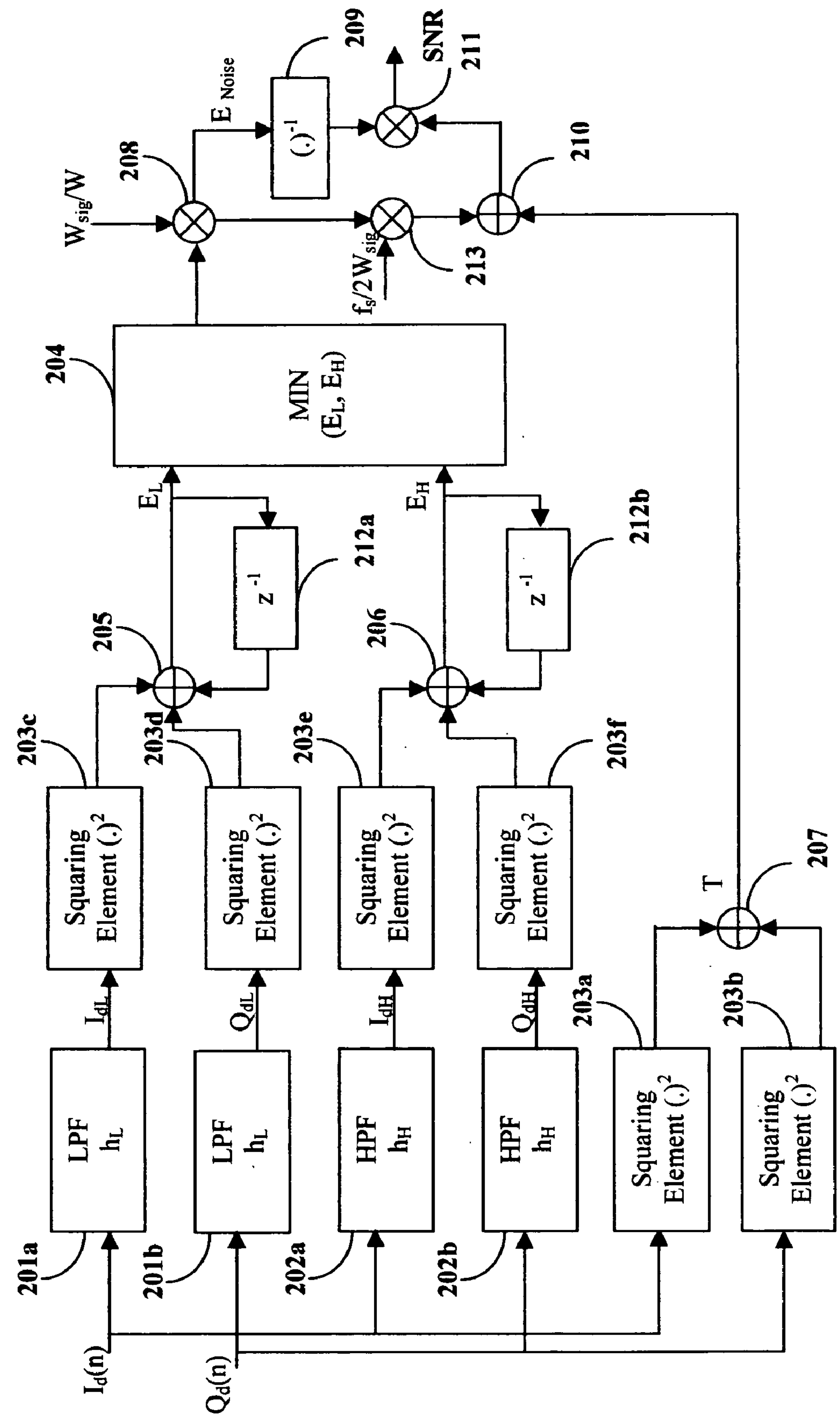

MAGNITUDE OF FFT
900
800
700
600
500
400
300
200
100
0
0
0.5
1
1.5
2
2.5
3
3.5
4
x 10^4
FREQUENCY IN Hz
110
111
112
105
113
114

FIGURE 1J



$I_d(n)$
$Q_d(n)$
201a
201b
202a
202b
LPF $h_L$
LPF $h_L$
HPF $h_H$
HPF $h_H$
$I_{dL}$
$Q_{dL}$
$I_{dH}$
$Q_{dH}$
Squaring Element $(.)^2$
203a
203b
203c
203d
203e
203f
207
T
205
206
212a
212b
$z^{-1}$
$E_L$
$E_H$
204
MIN $(E_L, E_H)$
$W_{sig}/W$
208
$E_{Noise}$
209
$(.)^{-1}$
211
SNR
$f_s/2W_{sig}$
213
210


FIGURE 2

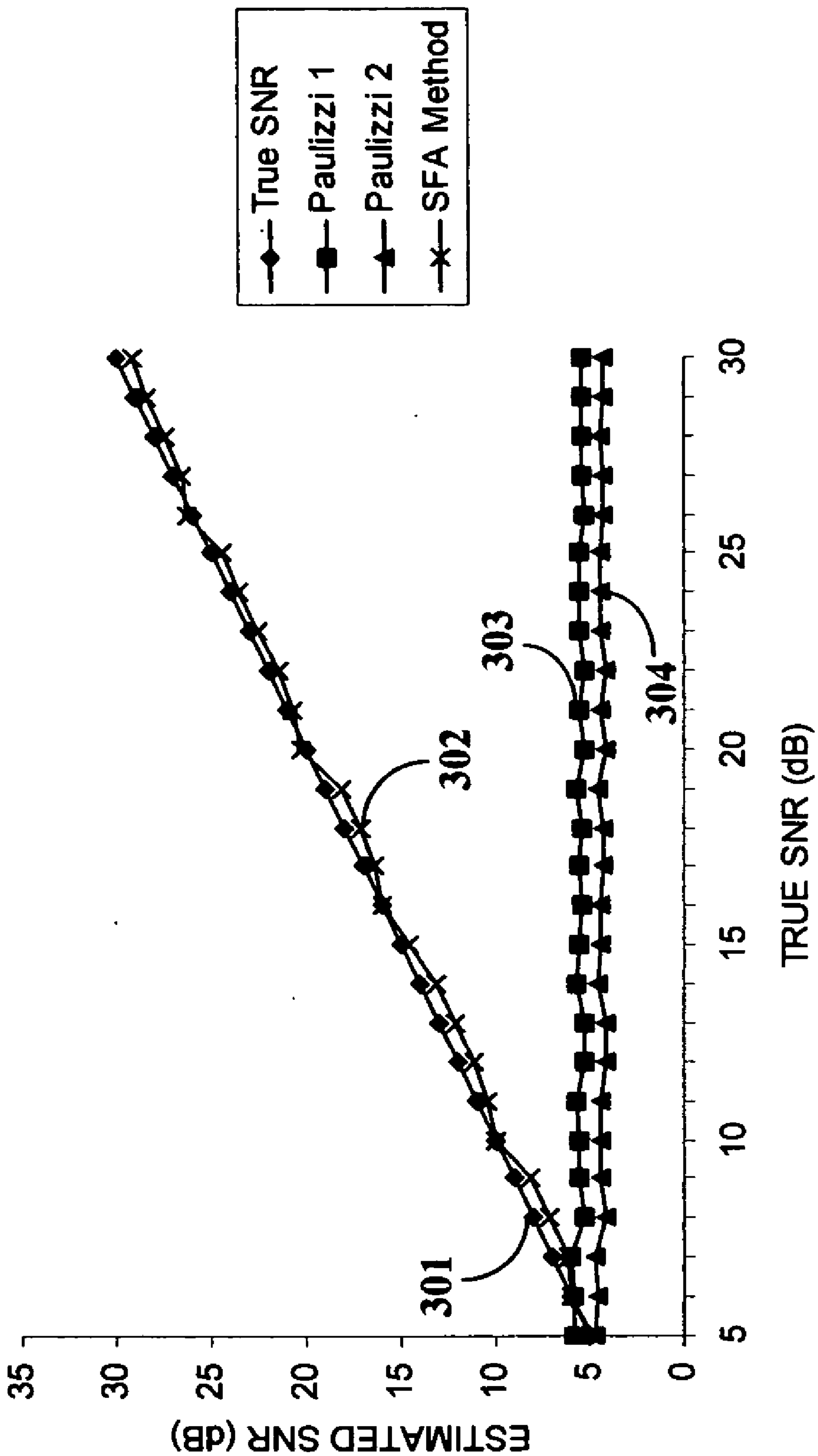

True SNR
Paulizzi 1
Paulizzi 2
SFA Method
ESTIMATED SNR (dB)
TRUE SNR (dB)
0
5
10
15
20
25
30
35
301
302
303
304

FIGURE 3A

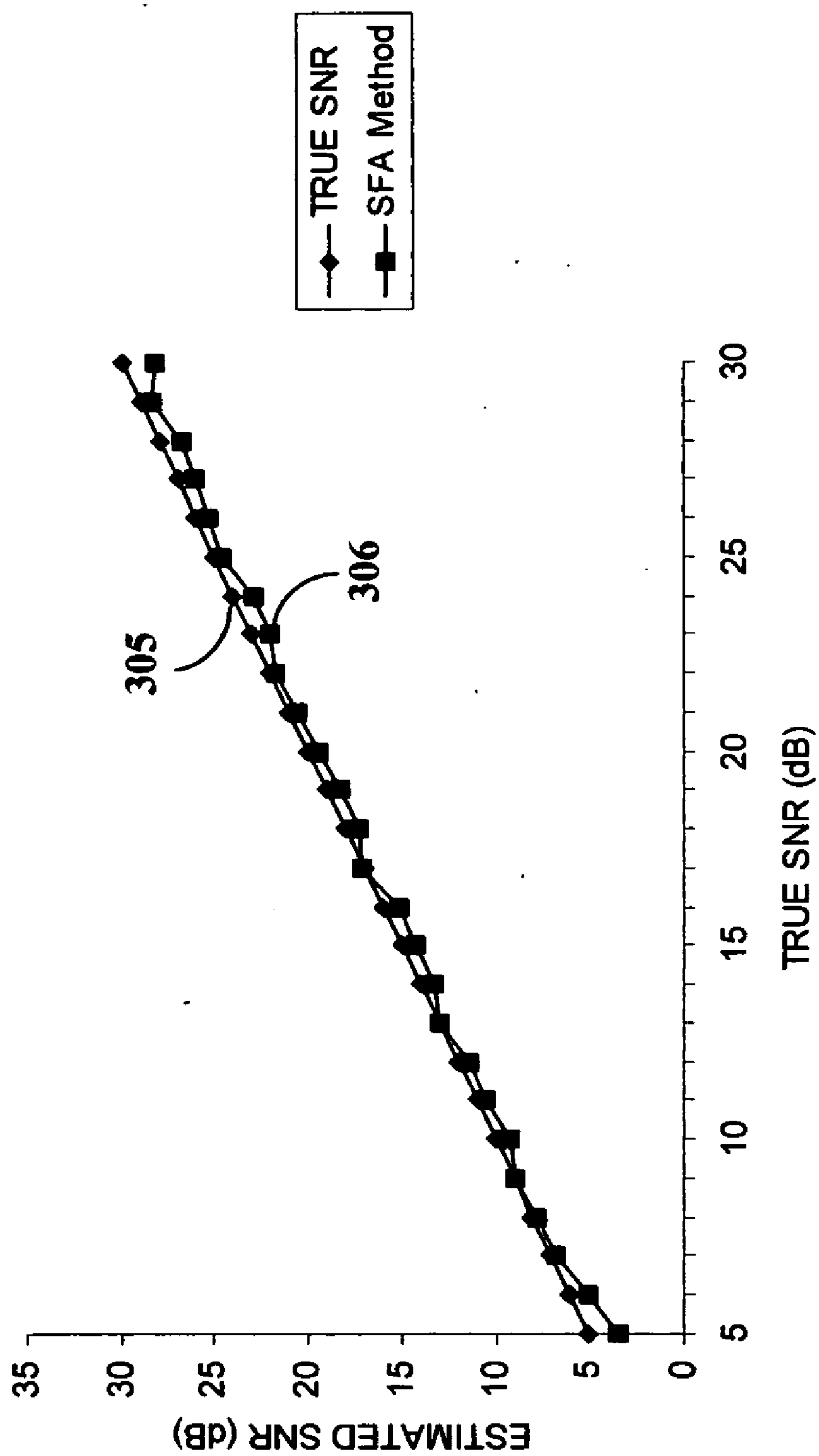

TRUE SNR
SFA Method
305
306
ESTIMATED SNR (dB)
TRUE SNR (dB)
0
5
10
15
20
25
30
35

FIGURE 3B

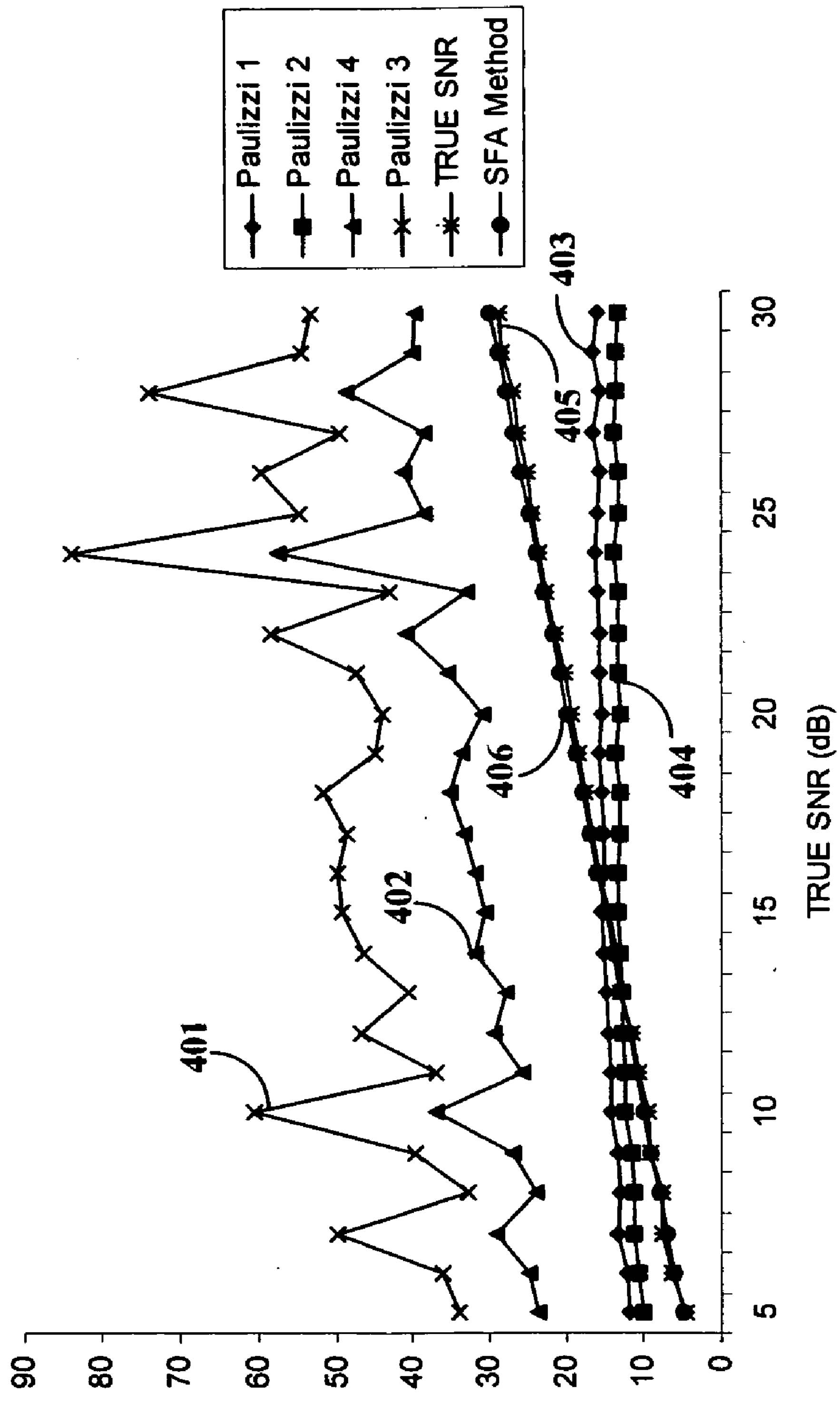

Paulizzi 1
Paulizzi 2
Paulizzi 4
Paulizzi 3
TRUE SNR
SFA Method
ESTIMATED SNR (dB)
TRUE SNR (dB)
0
10
20
30
40
50
60
70
80
90
5
10
15
20
25
30
401
402
403
404
405
406

FIGURE 4A

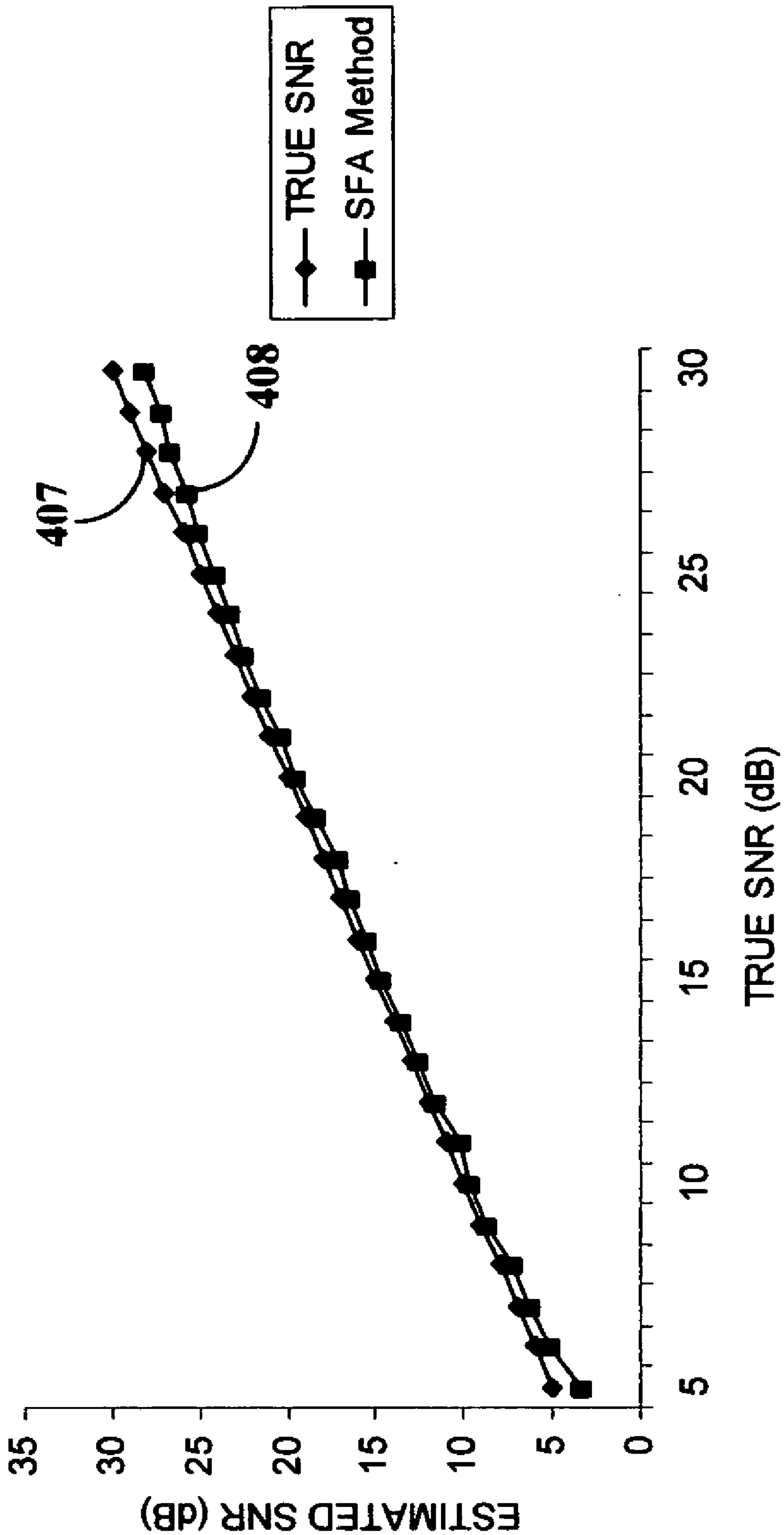

TRUE SNR
SFA Method
407
408
ESTIMATED SNR (dB)
0
5
10
15
20
25
30
35
TRUE SNR (dB)
5
10
15
20
25
30

FIGURE 4B

# SNR ESTIMATION USING FILTERS

## BACKGROUND OF THE INVENTION

[0001] The present invention, in general, relates to the estimation of signal to noise ratio (SNR) in digital communication receivers and specifically relates to SNR estimation using fixed, dynamic and multiple filters in the applications of low earth orbit (LEO) communication satellites.

[0002] The accurate estimation of SNR is a critical requirement in satellite communications applications. SNR estimation is required for the monitoring of impairments in the received signal in digital mobile radio systems and satellite modems. A typical satellite communications receiver or a digital mobile receiver is activated only at predetermined SNR thresholds. When the minimum SNR level is reached, the satellite communications receiver or a digital mobile receiver is switched on. The SNR value indicates whether the target satellite is within the permissible range of the receiver.

[0003] The methods in the art for SNR estimation are effective in medium (approximately 12 dB) to high SNR conditions. However, in the case of satellite modem applications in low SNR conditions, the relative strength of noise is high, hence the methods currently used in the art yield inaccurate results in SNR estimation.

[0004] Methods in the art for SNR estimation require the signal to be completely brought to base band. In satellite communication applications, it is difficult to bring the signal to base band, since there will be an uncertainty in the frequency estimation of the signal due to the Doppler shift induced by the high velocity of the LEO satellites.

[0005] Carrier offsets resulting from Doppler shifts are a recurring problem in LEO satellite communication systems. The methods disclosed in the art for SNR estimation typically fail in the presence of carrier offsets.

[0006] Discrete interference external radiation sources result in non-uniformity in the pattern of noise energy. The methods disclosed in the art for SNR estimation typically fail in the presence of discrete interferences.

[0007] Hence, there is an unsatisfied market need for an accurate method for SNR estimation, that can be effectively applied in conditions of low SNR and high carrier offsets, and that does not require the signal to be brought to base band. There is also an unsatisfied market need for an accurate method of SNR estimation in the presence of discrete interferences.

## SUMMARY OF THE INVENTION

[0008] The present invention provides a method of accurately determining SNR using filters. The following three embodiments of the invention are disclosed: a method for accurately determining SNR using fixed filters, multiple filters and dynamic filters. The noise energy value is calculated by applying low pass and high pass filters. The minimum amongst the energy values estimated by the low pass and high pass filters, is selected to calculate the actual noise energy in the signal. The present invention provides an accurate method for SNR estimation in conditions of low SNR and high carrier offsets, and also provides a method that does not require the signal to be brought to base band. The use of multiple filters provides an accurate SNR measurement even in the presence of discrete interferences.

[0009] This invention proposes a method of filter selection and application and a method of SNR computation. The method of this invention, including all its embodiments will be hereafter referred to as the selective filter application (SFA) method.

[0010] One advantage of the SFA method is that SNR is accurately estimated even in the conditions of extreme frequency variations. Extreme frequency variations typically exist in the signals received from LEO satellites. The Doppler shift varies significantly in the signals transmitted by LEO satellites, for example, from −3000 Hz to +3000 Hz, varying at a maximum rate of 27 Hz/sec.

[0011] Another advantage of the SFA method is that SNR is accurately estimated even in low SNR conditions.

[0012] Another advantage of the SFA method is that SNR is accurately estimated even in the presence of high carrier offsets.

[0013] Another advantage of the SFA method is that it can be used for different types of modulation schemes, such as binary phase shift keying (BPSK), M-ary phase shift keying (mPSK), M-ary frequency shift keying (mFSK), and differential modulation schemes.

[0014] Another advantage of the SFA method is that the signal does not have to be brought to base band for SNR estimation.

## BRIEF DESCRIPTION OF THE DRAWINGS

[0015] FIG. 1A illustrates the graphical representation of the equation $2W+W_{sig}\leqq f_s/2$ for the first embodiment of the invention that uses a fixed filter low pass and high pass filter.

[0016] FIG. 1B illustrates the intersection of signal and filter spectra that occurs when the equation $2W+W_{sig}\leqq f_s/2$ is not satisfied.

[0017] FIG. 1C illustrates the signal and filter spectrum for the second embodiment of the invention that uses dynamic filters.

[0018] FIG. 1D illustrates an example of a sampled signal spectrum with discrete interferences.

[0019] FIG. 1E illustrates the application of the first multiple filter to the signal spectrum.

[0020] FIG. 1F illustrates the application of the second multiple filter to the signal spectrum.

[0021] FIG. 1G illustrates the application of the third multiple filter to the signal spectrum.

[0022] FIG. 1H illustrates the application of the fourth multiple filter to the signal spectrum.

[0023] FIG. 1I illustrates the original signal spectrum.

[0024] FIG. 1J is a representation of the outputs of the multiple filters as illustrated in FIG. 1E, FIG. 1F, FIG. 1G, FIG. 1H and the original signal spectrum shown in FIG. 1I.

[0025] FIG. 2 illustrates the method of SNR computation.

[0026] FIG. 3A illustrates a graph that compares the SNR estimated by the proposed SFA method with the methods of prior art for a quadrature phase shift keying (QPSK) modulated base band signal at 10 Hz frequency offset.

[0027] FIG. 3B illustrates a graph that compares the SNR estimated by the proposed SFA method with the methods of prior art for a QPSK modulated base band signal at 1000 Hz frequency offset.

[0028] FIG. 4A illustrates a graph that compares the SNR estimated using the SFA method with the SNR estimated using a method of prior art for a symmetrical differential phase shift keying (SDPSK) modulated base band signal at 0 Hz frequency offset.

[0029] FIG. 4B illustrates a graph that compares the true SNR with the estimated SNR using the SFA method for a SDPSK modulated base band signal at a high frequency offset of 1000 Hz.

## DETAILED DESCRIPTION OF THE INVENTION

[0030] FIG. 1A illustrates the graphical representation **101** of the equation $2W+W_{sig} \leqq f_s/2$ for the first embodiment of the invention that uses a fixed filter at both ends of the signal spectrum **105**. The sampling rate, abbreviated as $f_s$, is determined by an analog to digital converter (ADC). The bandwidth of the low pass **103** and high pass **104** filters are fixed at W Hz. The maximum signal bandwidth required is assumed to be $W_{sig}$ Hz. The bandwidth W of the low and high pass frequency bands are selected such that the spectrum of one of those filters does not intersect with that of the incoming signal **105**. This is ensured by compliance to the equation:

$$2W+W_{sig} \leqq f_s/2$$

[0031] FIG. 1B illustrates the graphical representation **102** intersection of signal spectrum **105** and filter spectra **103**, **104** that occurs when the equation $2W+W_{sig} \leqq f_s/2$ is not satisfied. Ideally at least one of the two filters, i.e., either the low pass filter **103** or high pass filter **104** must estimate only the noise energy and should not estimate the energy in any section of the signal spectrum **105**. However, when the signal spectrum **105** and filter spectrum **103**, **104** intersect as illustrated in FIG. 1B when the equation $2W+W_{sig} \leqq f_s/2$ is not satisfied, both the high pass **104** and low pass filters **103** estimate not only the noise, but also a section of the signal spectrum **105**, as depicted by the shaded areas **106**.

[0032] FIG. 1C illustrates the signal spectrum **105** and dynamic filter spectrum **108** for the second embodiment of the invention that uses dynamic filters **108**. When the characteristics of the signal **105** change to signal **107** due to frequency shift, the filter **108** dynamically shifts to a new position **109**.

[0033] The center frequency $W_c$ of the received signal is estimated by the receiver. The bandwidth of the filter (B) is also known, for example B can be equal to signal bandwidth $W_{sig}$. For a modulated signal, $W_{sig}=B+\alpha B$, where B is the baud rate and $\alpha$ is the roll off factor of the pulse shaping filter.

[0034] A single dynamic filter **108** is used at one end of the signal spectrum **105** for noise energy estimation. The same method, as illustrated under the description of FIG. **2** is applied for the estimation of SNR using dynamic filters.

[0035] When the center frequency of the signal **105** shifts to a new position **107** as illustrated in FIG. 1C, the filter response shifts correspondingly **109** and thereby, overlap of the shifted filter response **109** with the actual signal **105** is avoided.

[0036] A method for deriving a dynamic filter is given below. The filter coefficients of the dynamic filter are calculated. The dynamic filter is defined by its filter coefficients. The following values are known: bandwidth of the filter B, center frequency of the original signal $W_C$, frequency shift $f_{sh}$, shift in center frequency of the signal $W_{DC}$ and sampling duration $T_s$. The objective is to derive the filter coefficients, thereby creating the dynamic filter.

[0037] Given the center frequency $W_C$ and the frequency shift $f_{sh}$ of the original signal, the center frequency of the shifted signal is defined by the equation,

$$W_{DC}=W_C+f_{sh}$$

[0038] The design and application of a dynamic filter comprises of first creating a filter and shifting it appropriately as the original signal shifts.

[0039] The following example describes the method of creating a Butterworth filter and dynamically shifting it. Note that this invention is not restricted to the use of a particular type of filter, such as the filter provided below. The filter coefficients of a second order Butterworth band pass filter are given as $n_0$, $n_1$, $n_2$, $d_0$, $d_1$ and $d_2$. $T_s$ is the sampling duration and B is the bandwidth of the filter.

$$C=\cot(W_{DC}\times T_s/2)$$

$$n_0=B.C$$

$$n_1=0$$

$$n_2=-B.C$$

$$d_0=B.C+C^2+1$$

$$d_1=-2(C^2-1)$$

$$d_2=-B.C+C^2+1$$

[0040] The filter transform function H(z) is given as,

$$H(z)=(n_0+n_1z^{-1}+n_2z^{-2})/(d_0+d_1\,z^{-1}+d_2z^{-2})$$

[0041] When the center frequency of the original signal shifts, the frequency shift is known and accordingly the dynamic filter parameters are determined. After the dynamic filter is applied to the signal spectrum, $E_{Noise}$ is determined by computing the dynamic filter output.

[0042] FIG. 1D illustrates an example of a sampled signal spectrum **105** with discrete interferences. Discrete interferences to the signal cause non-uniform noise **110** in the frequency domain. However, the non-uniform noise **110** or discrete interferences in the frequency domain may or may not reside in the signal spectrum **105**.

[0043] FIG. 1E illustrates the application of the first multiple filter **111** to the signal spectrum **105**.

[0044] FIG. 1F illustrates the application of the second multiple filter **112** to the signal spectrum.

[0045] FIG. 1G illustrates the application of the third multiple filter **113** to the signal spectrum.

[0046] FIG. 1H illustrates the application of the fourth multiple filter **114** to the signal spectrum.

[0047] FIG. 1I illustrates the original signal spectrum **105**.

[0048] FIG. 1J is a representation of the outputs of the multiple filters as illustrated in FIG. 1E, FIG. 1F, FIG. 1G, FIG. 1H and the original signal spectrum 105 shown in FIG. 1I.

[0049] Consider the case of a discrete interference or a non uniform noise 110 illustrated in FIG. 1D. Consider the case when M multiple filters are used. The actual white noise energy $E_{Noise}$ is best represented by the filter that measures the least energy.

$$E_{Noise}=[W_{sig}/w](\min(E_1,E_2 \ldots E_M))$$

where $W_{sig}$ is the maximum signal bandwidth, M is the number of multiple filters, w is the bandwidth of each multiple filter, and $E_1$, $E_2$ . . . $E_M$ are the noise values estimated by the filters.

[0050] FIG. 2 illustrates the method of computation of SNR. The in-phase sample $I_d(n)$ of the incoming signal is passed through a low pass filter 201*a*, while the quadrature phase sample $Q_d(n)$ is passed through another low pass filter 201*b*, both the filters having a bandwidth W. $I_{dL}$ is the in-phase sample of the digital signal after the application of the low pass filter 201*a* to the in-phase sample $I_d(n)$ of the sampled signal. $Q_{dL}$ is the quadrature phase sample after the application of the low pass filter 201*b* to the quadrature phase sample $Q_d(n)$ of the sampled signal.

[0051] The application of the filters is illustrated by the following equations:

$$I_{dL}=I_d*h_L,$$

$$Q_{dL}=Q_d*h_L$$

[0052] where $h_L$ is the filter coefficient and * represents convolution.

[0053] The relation between the input $\chi_n$ and the output $y_n$ of the filter is given by the equation:

$$(y_n\times b_0)+(y_{n-1}\times b_1)+(y_{n-2}\times b_2)+ \ldots =(\chi_n\times a_0)+(\chi_{n-1}\times a_1)$$

[0054] where $a_0$, $a_1$, . . . , and $b_1$, $b_2$ . . . are filter coefficients.

[0055] The filter may be of infinite impulse response (IIR) or finite impulse response (FIR). The output in frequency domain is given by:

$$\tilde{I}_{dL}(w)=H(w)\times\tilde{I}_d(w)$$

$$\tilde{Q}_{dL}(w)=H(w)\times\tilde{Q}_d(w)$$

[0056] The outputs of the low pass filter 201*a* and 201*b* are fed to the squaring elements 203*c* and 203*d* respectively, that square both the in-phase $I_{dL}$ and quadrature phase $Q_{dL}$ low pass filtered samples. The low pass signal energy ($E_L$) is obtained by summing the squares of the quadrature phase sample of the noise energy and the in-phase sample of the noise energy in a summer 205. The summer is fed with a one step delay $z^{-1}$ 212*a*. The summing operation is illustrated by the following equation

$$E_L=\sum_{P=0}^{P-1} I_{dL}^2[p]+Q_{dL}^2[p]$$

wherein, p is the time index with a predetermined limit.

[0057] The in-phase sample $I_d(n)$ of the incoming signal is passed through a high pass filter 202*a*, while the quadrature phase samples $Q_d(n)$ is passed through another high pass filter 202*b*, both the filters having a bandwidth W. $I_{dH}$ is the in-phase sample of the digital signal after the application of the high pass filter 202*a*, while $Q_{dH}$ is the quadrature phase sample after the application of the high pass filter 202*b*.

[0058] Application of the filters is illustrated by the following equation:

$$I_{dH}=I_d*h_H$$

$$Q_{dH}=Q_d*h_H$$

where $h_H$ is the filter coefficient of the high pass filter.

[0059] Outputs of the high pass filters 202*a* and 202*b* are fed to the squaring elements 203*e* and 203*f* respectively, that square both the in-phase and quadrature phase high pass filtered output samples. The high pass signal energy ($E_H$) is obtained by summing the squares of the quadrature phase samples of the noise energy and the in-phase sample of the noise energy in a summer 206. The summer is also fed with one step delay $z^{-1}$ 212*b*.

$$E_H=\sum_{n=0}^{n-1} I_{dH}^2[n]+Q_{dH}^2[n]$$

Wherein n is the time index with a predetermined limit.

[0060] The minimum of the low pass signal energy and high pass signal energy is determined. This minimum is represented by min ($E_L$, $E_H$) 204. The minimum is chosen because if any one of the two low pass filters 201*a* or 201*b*, or the high pass filters 202*a* or 202*b*, filters a section of the actual signal along with the noise. That particular filter will provide a higher value for noise compared to the filter that filters only the noise. Hence, the minimum of $E_L$ and $E_H$ 204 is the most representative of the noise energy without discrete components. In case there is a discrete noise component, which comes in the region of the low or high pass filters, the minimum value of $E_L$ and $E_H$ ensures that the discrete noise is not measured.

[0061] The minimum of $E_L$ and $E_H$ 204 is fed to the distributor 208. The ratio of maximum signal bandwidth $W_{sig}$ to bandwidth W of the high pass filters 202*a* or 202*b* and the low pass filters 201*a* or 201*b*, $W_{sig}$/W is also fed to the distributor 208.

[0062] Assuming additive white noise, the total noise energy ($E_{Noise}$) is calculated by the formula:

$$E_{Noise}=W_{sig}/W[\min(E_L,E_H)]$$

[0063] The in-phase $I_d$ (n) and quadrature phase $Q_d$ (n) samples are fed to the squaring elements 203*a* and 203*b* respectively. The squared samples are summed using a summer 207.

The total energy of the incoming signal T is represented by

$$T=\sum_{K=0}^{N-1} I_d^2[k]+Q_d^2[k]$$

[0064] where $I_d$ [k] and $Q_d$ [k] are the in-phase and quadrature phase samples of the incoming signal respectively.

[0065] The total noise energy ($E_{Noise}$) is subtracted 210 from the energy of the received signal (T). The output of this operation provides the actual signal energy (S).

$$S=T-(fs/2)*E_{Noise}/Wsig$$

[0066] The calculated total noise energy ($E_{Noise}$) is fed to an inverter **209** in order to determine $1/E_{Noise}$, and is forwarded to a distributor **211**. The actual signal energy (S) is also fed to the distributor **211**. The signal to noise ratio (SNR) determined at the distributor **211** is the ratio of actual signal energy (S) to the total noise energy ($E_{Noise}$).

$$\mathrm{SNR}=S/E_{Noise}$$

[0067] FIG. **3**A illustrates a graph that compares the SNR estimated by the SFA method with two methods known in the art for a quadrature phase shift keying (QPSK) modulated base band signal at 10 Hz frequency offset. The two methods in the art are the SNR estimator techniques of Paulizzi's, et al., illustrated as equations (3) and (4) in the literature "A Comparison of SNR Estimation Techniques for QPSK Modulations" by David R. Pauluzzi, Andrew S. Toms and Norman C. Beaulieu, IEEE Transactions on Communications, Vol. 4, No. 2, February 2000. The plots of the two equations (3) and (4) are depicted as plot **303** and **304** in FIG. **3**A. It can be observed from FIG. **3**A that the above two methods yield SNR measurements that significantly deviate from the true SNR values. Plot **302** of the SFA method lies closest to the true SNR plot **301**, even at low SNRs.

[0068] FIG. **3**B illustrates a graph that compares the SNR estimated by the proposed SFA method with the true SNR for a QPSK modulated base band signal at a 1000 Hz frequency offset. It is observed that even at high frequency offsets, the plot **306** of the SFA method lies closest to the true SNR plot **305**. Note that the prior art is not plotted in FIG. **3**B as it significantly fails in the accuracy of SNR estimation.

[0069] FIG. **4**A illustrates a graph that compares the SNR estimated using the SFA method against the SNR estimated using a method in the art for a symmetrical differential phase shift keying (SDPSK) modulated base band signal at 0 Hz frequency offset. The four methods in the art considered are the SNR estimator techniques of Pauluzzi et al., illustrated as equations (3), (4), (5) and (6) "A Comparison of SNR Estimation Techniques for QPSK Modulations" by David R. Pauluzzi, Andrew S. Toms and Norman C. Beaulieu, IEEE Transactions on Communications, Vol. 4, No. 2, February 2000". The results of the method using equation (3) disclosed above by Paulizzi, et al. is depicted as plot **403**, the result of the method using equation (4) is depicted as **404**, the result of the method using equation (5) is depicted as plot **401** in FIG. **4**A and the result of the method using equation (6) is depicted as plot **402**. It can be observed from FIG. **4**A that the plot of the SFA method **406** of this invention lies closest to the true SNR **405**.

[0070] FIG. **4**B illustrates a graph that compares true SNR against the SNR estimated using the SFA method for a SDPSK modulated base band signal at a high frequency offset of 1000 Hz. It can be observed from FIG. **4**B that the plot **408** of the SFA method of this invention lies very close to the true value of the true SNR **407**.

[0071] In summary, it can be observed from FIG. **3**A, FIG. **3**B, FIG. **4**A and FIG. **4**B that the SFA method provides an accurate estimation of SNR, and the accuracy is not significantly affected by low SNR and high frequency offset conditions.

**1**. A method of estimating the signal to noise ratio of a sampled signal, comprising the steps of:

determining the bandwidth W of a low pass filter and the bandwidth W of a high pass filter from the relation $2W+W_{sig} \leqq f_s/2$, where $W_{sig}$ is the maximum bandwidth of the sampled signal and $f_s$ is the sampling rate at which an analog signal is sampled to derive said sampled signal;

estimating a first noise energy value by applying said low pass filter and estimating a second noise energy value by applying said high pass filter to the sampled signal;

determining the minimum noise energy value amongst said first noise energy value and said second noise energy value;

calculating the noise energy within the signal using said minimum noise energy value; and

estimating the received signal energy and computing the signal to noise ratio as a ratio of said actual signal energy to the noise energy in the signal.

**2**. A method of estimating the signal to noise ratio of a sampled signal, comprising the steps of:

estimating the noise energy by applying a dynamic filter to said sampled signal, wherein the center frequency of the filter is dynamically shifted;

calculating the noise energy value in the sampled signal;

estimating the received signal energy; and

estimating the actual signal energy from said received signal energy by subtracting the noise energy value and computing the signal to noise ratio as a ratio of said actual signal energy to the noise energy in the sampled signal.

**3**. A method of estimating the signal to noise ratio of a sampled signal, comprising the steps of:

applying multiple filters and determining the value of the noise energy measured by each of said multiple filters;

determining the lowest value of the noise among the noise energies measured by said multiple filters;

calculating the noise energy in said sampled signal using said lowest value of noise; and

estimating the received signal energy and computing the actual signal energy by subtracting the noise energy in the sampled signal from said received signal energy and computing the signal to noise ratio as a ratio of said actual signal energy to the noise energy in the signal.

**4**. The method of claim 1, wherein the step of estimating said first noise energy by applying said low pass filter of bandwidth W to the sampled signal, further comprises the steps of:

computing the quadrature phase energy value of the first noise energy by applying the low pass filter to said quadrature phase sample of the sampled signal;

computing the in-phase sample of the first noise energy by applying the low pass filter to said in-phase sample of the sampled signal; and,

computing the first noise energy by summing the squares of the said quadrature phase energy value of the first noise energy and said in-phase energy value of the first noise energy.

**5**. The method of claim 1, wherein the step of estimating said second noise energy by applying said high pass filter of bandwidth W to the sampled signal, further comprises the steps of:

computing the quadrature phase sample of the second noise energy by applying the high pass filter to said quadrature phase sample of the sampled signal;

computing the in-phase sample of the second noise energy by applying the high pass filter to said in-phase sample of the sampled signal; and

computing the second noise energy by summing the squares of the said quadrature phase sample of the second noise energy and said in-phase sample of the second noise energy.

**6**. The method of claim 1, wherein the step of determining the total noise energy applies the equation:

$$E_{Noise}=[W_{sig}/W]\min(E_L,E_H)$$

where $W_{sig}$ is the maximum signal bandwidth, W is the bandwidth of the low pass and high pass filters, $E_L$ is the first noise energy, $E_H$ is the second noise energy, and min ($E_L$, $E_H$) is the minimum value of said first noise energy and said second noise energy.

**7**. The method of claim 3, wherein the step of determining the total noise energy comprises:

applying the equation $E_{Noise}=[W_{sig}/W]\ (\min(E_1, E_2 \ldots E_M))$

where $W_{sig}$ is the maximum signal bandwidth, M is the number of multiple filters, W is the bandwidth of each of the multiple filters and $E_1$, $E_2 \ldots E_M$ are the noise values estimated by the filters.

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