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Vishinsky

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(54) **SELF-TUNED ACTIVE BANDPASS FILTERS**

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(22) Filed: **Jun. 7, 2005**

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Assistant Examiner—Kimberly E Glenn

(52) U.S. Cl. 333/17.1; 333/202; 327/552

(57) ABSTRACT

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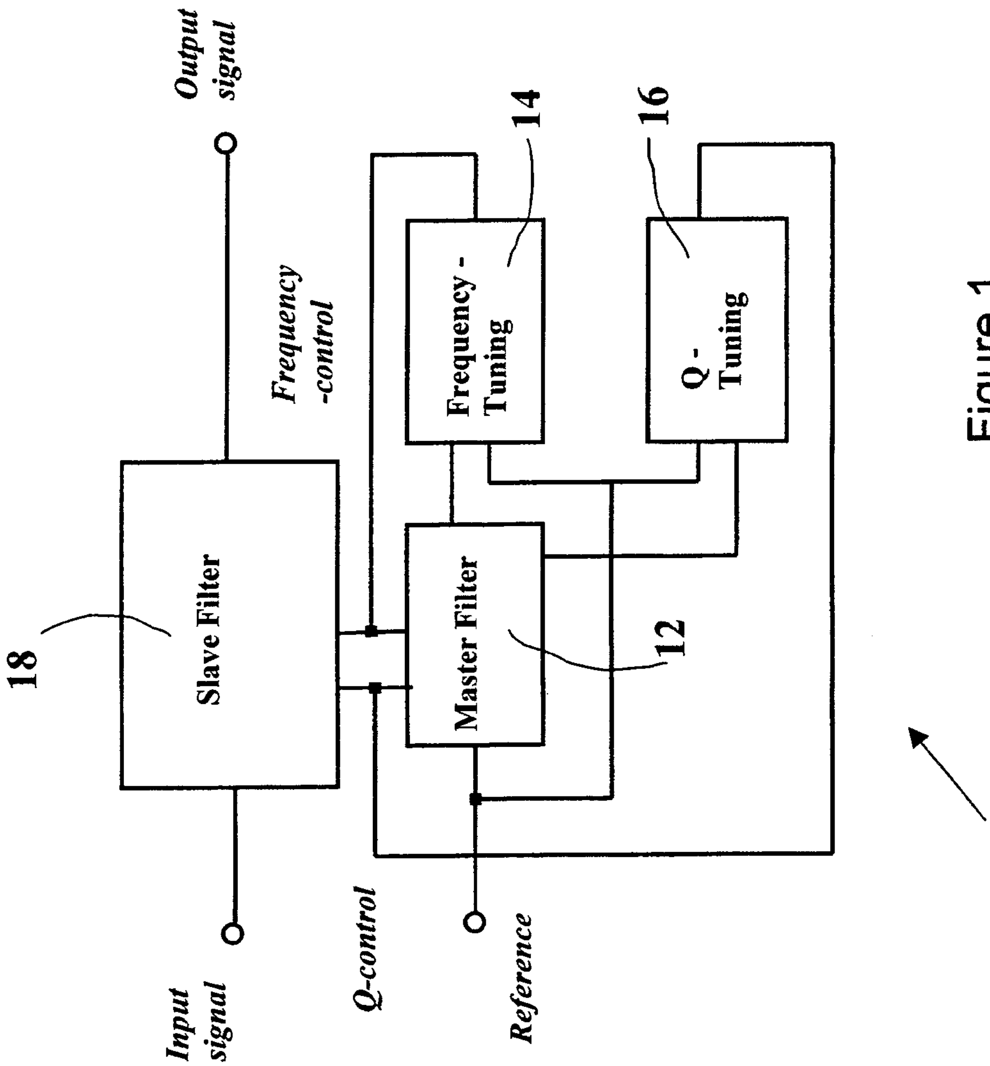


Figure 1
PRIOR ART

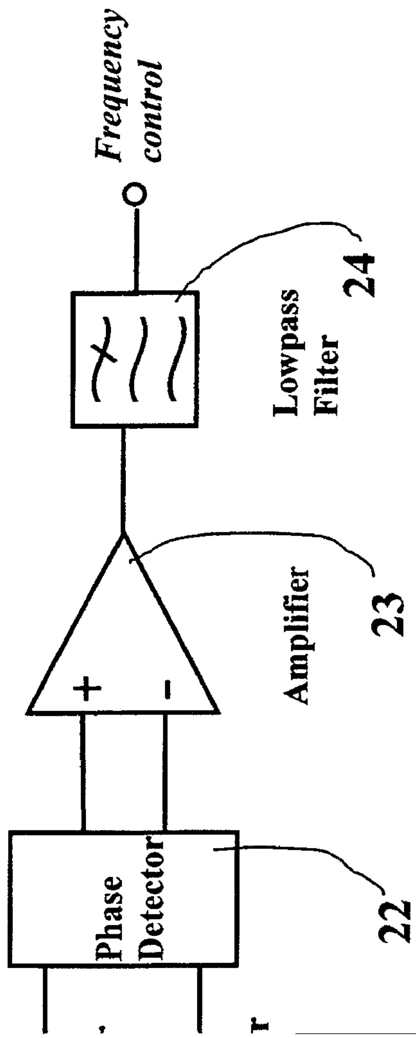


Figure 2
PRIOR ART

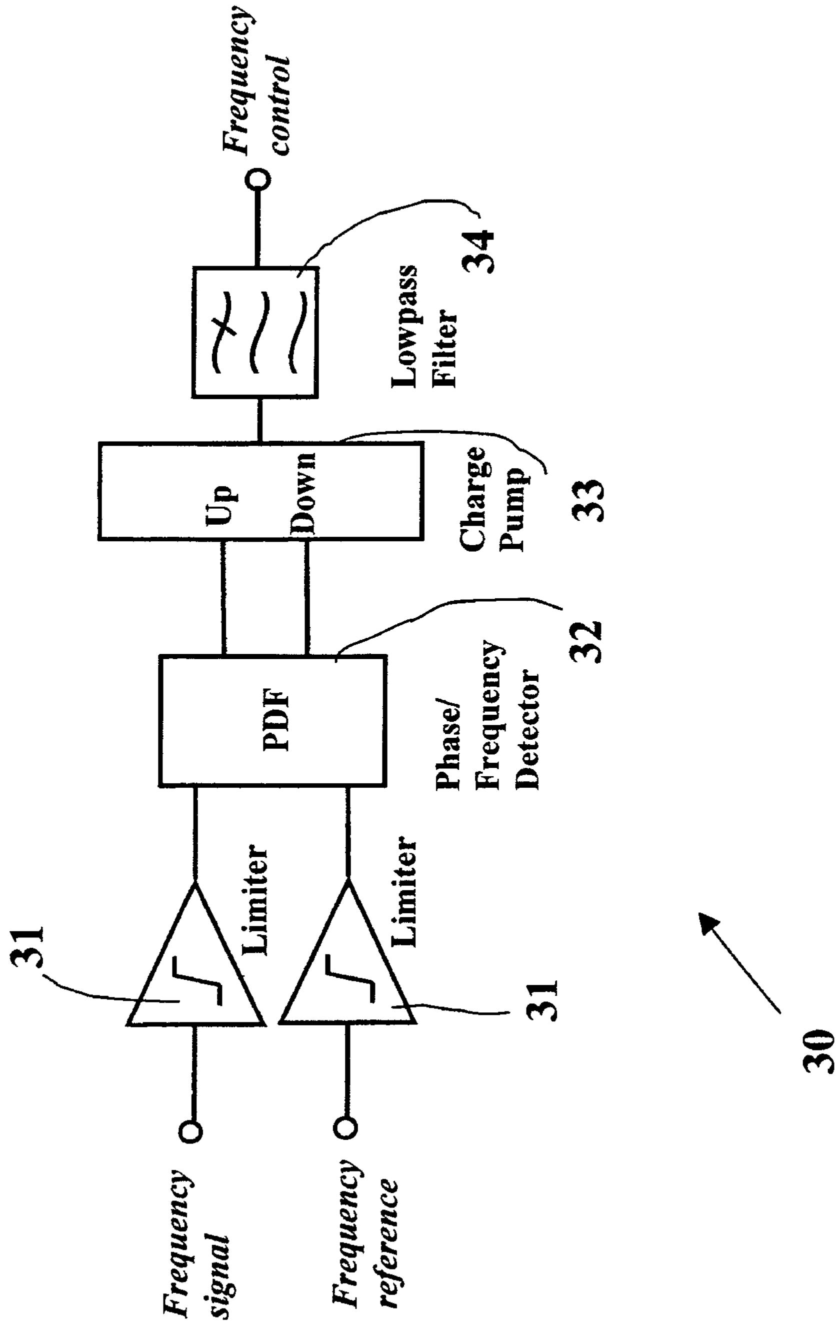


Figure 3
PRIOR ART

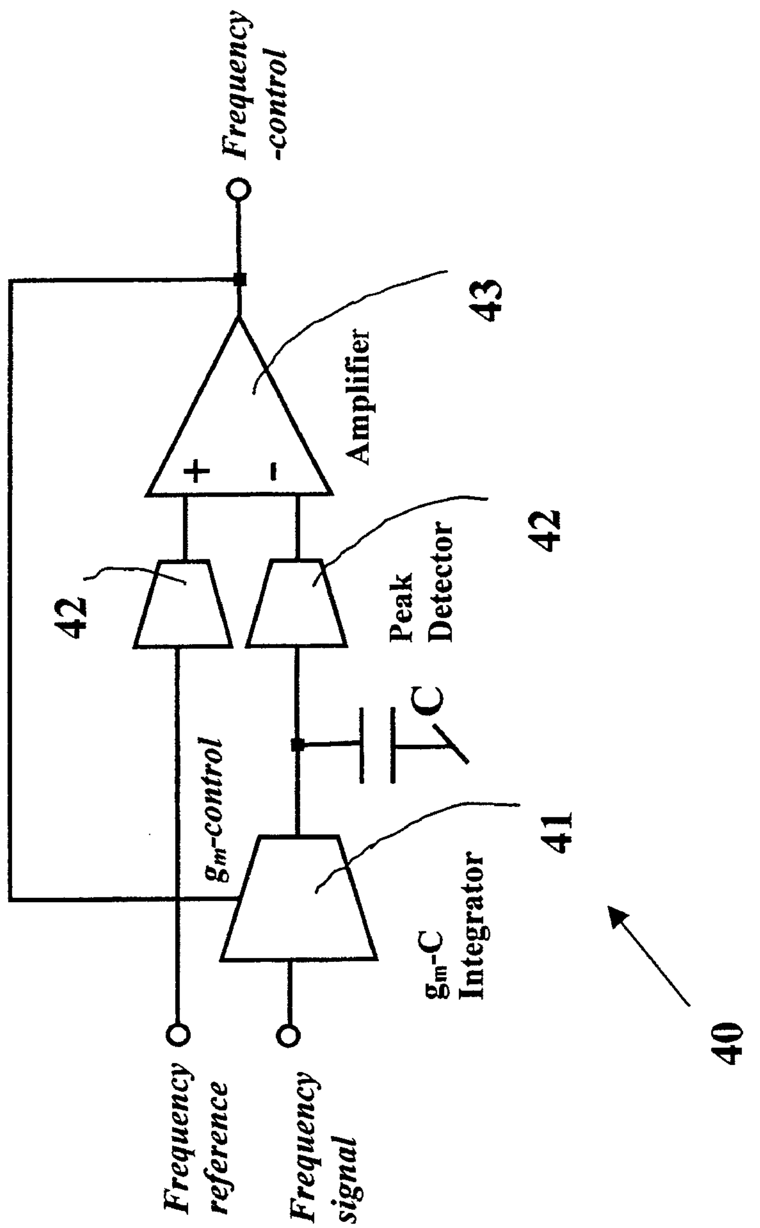


Figure 4
PRIOR ART

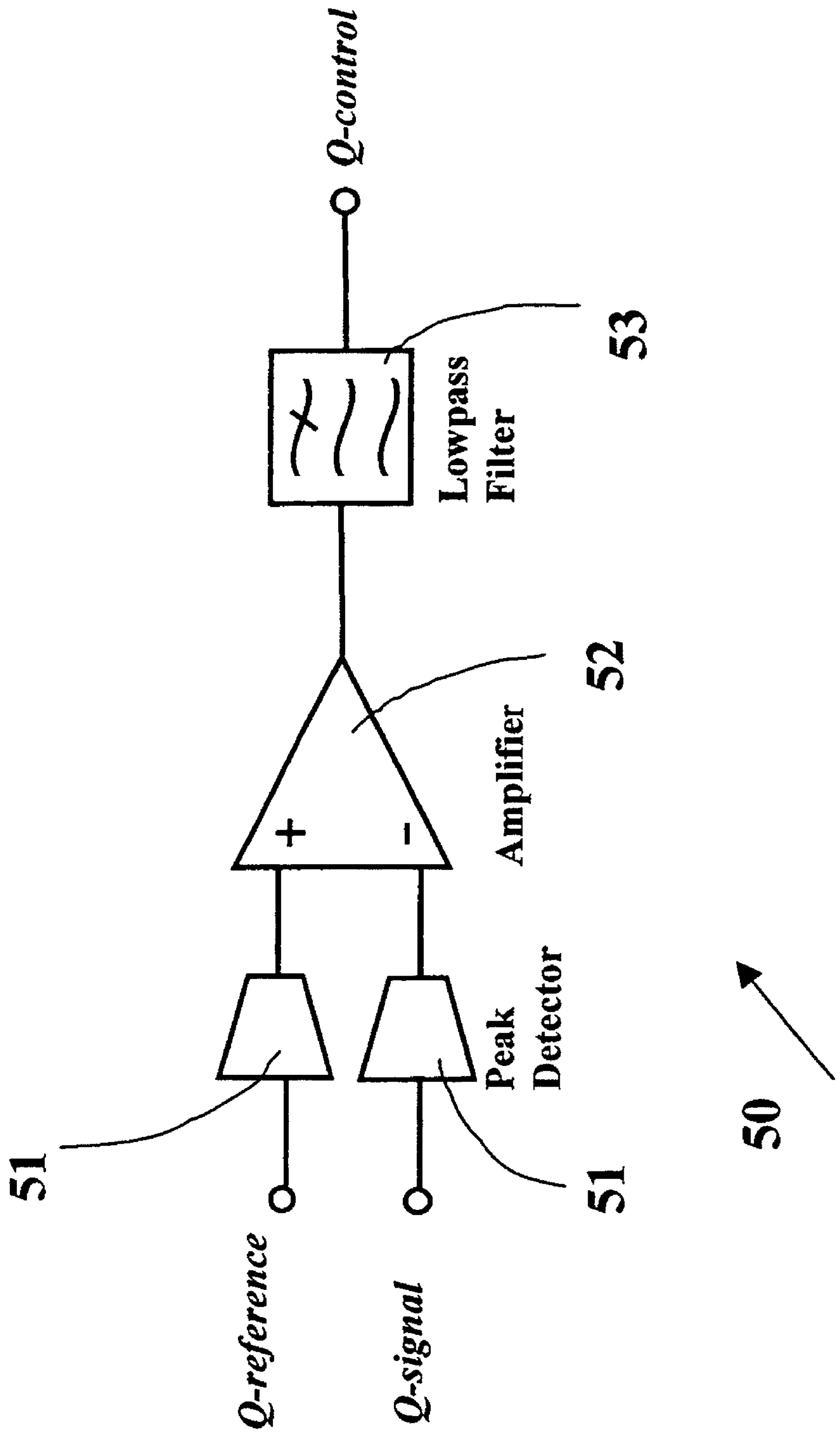
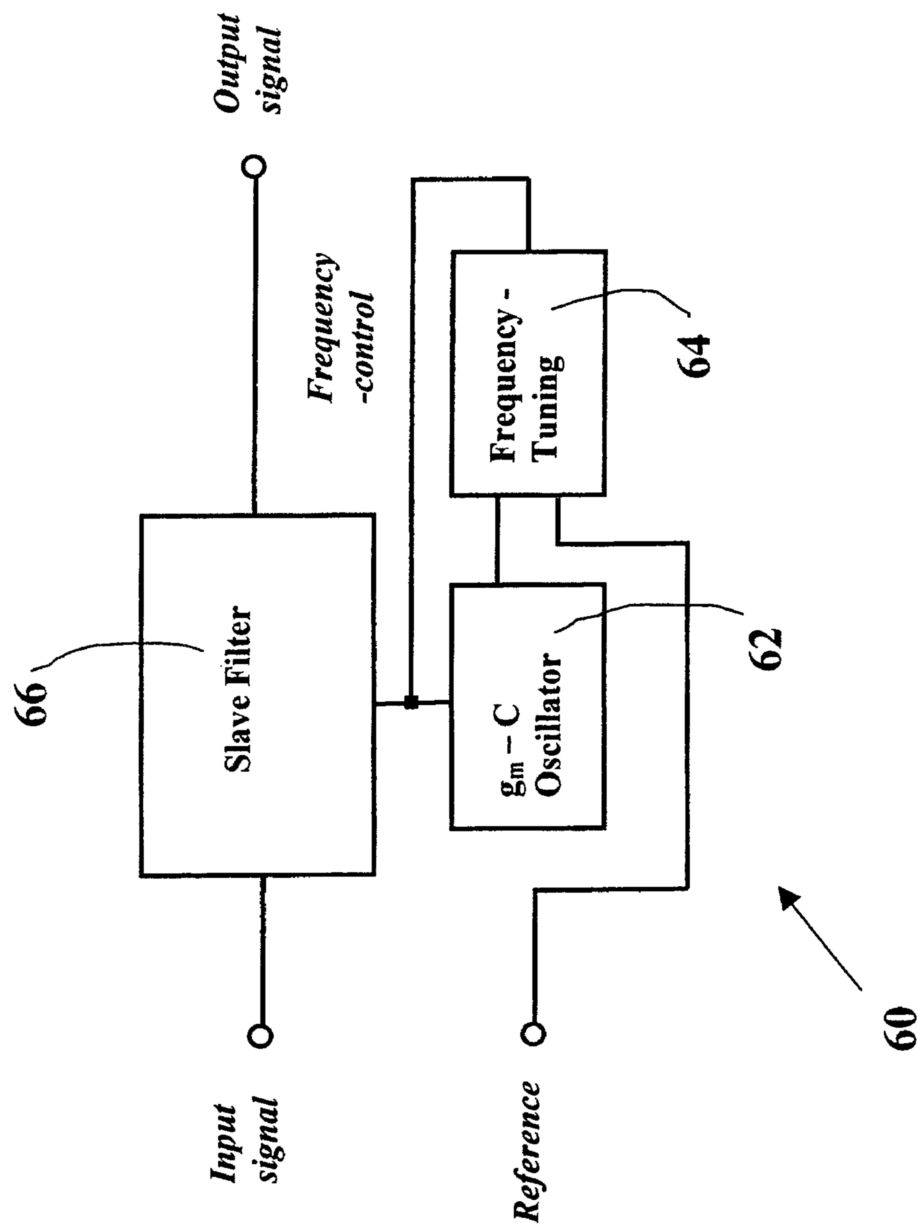


Figure 5
PRIOR ART



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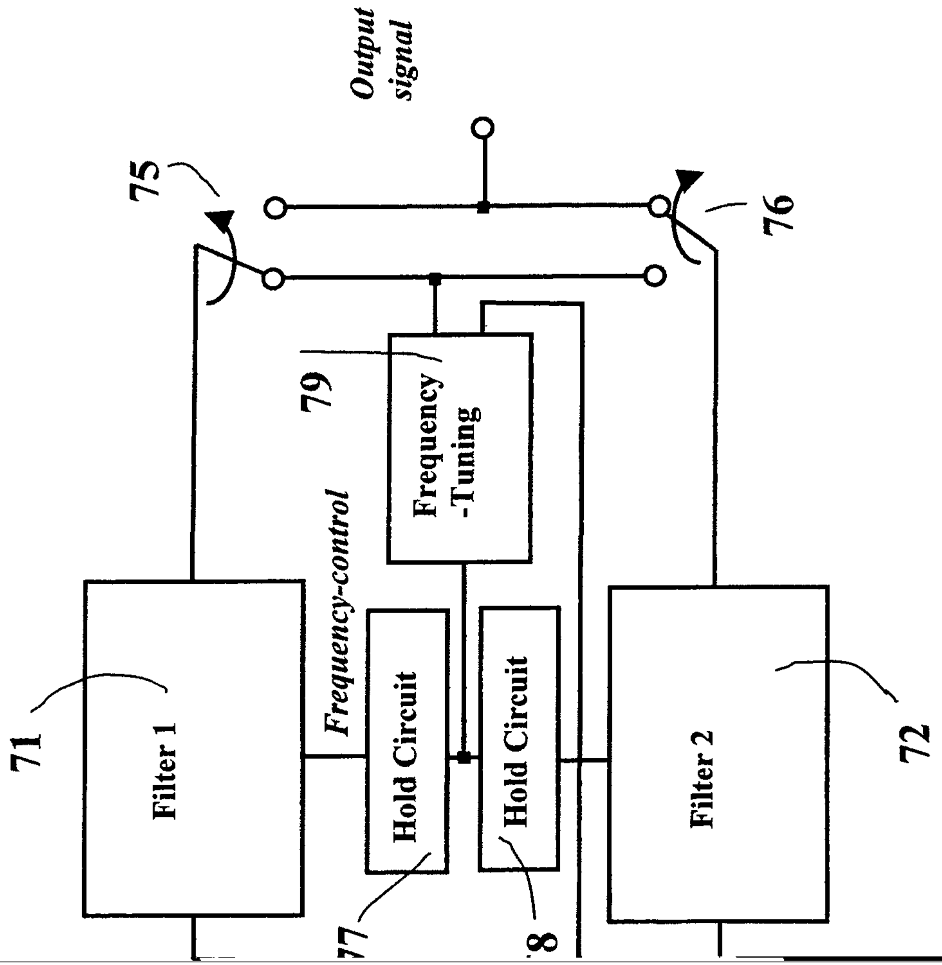
PRIOR ART



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PRIOR ART

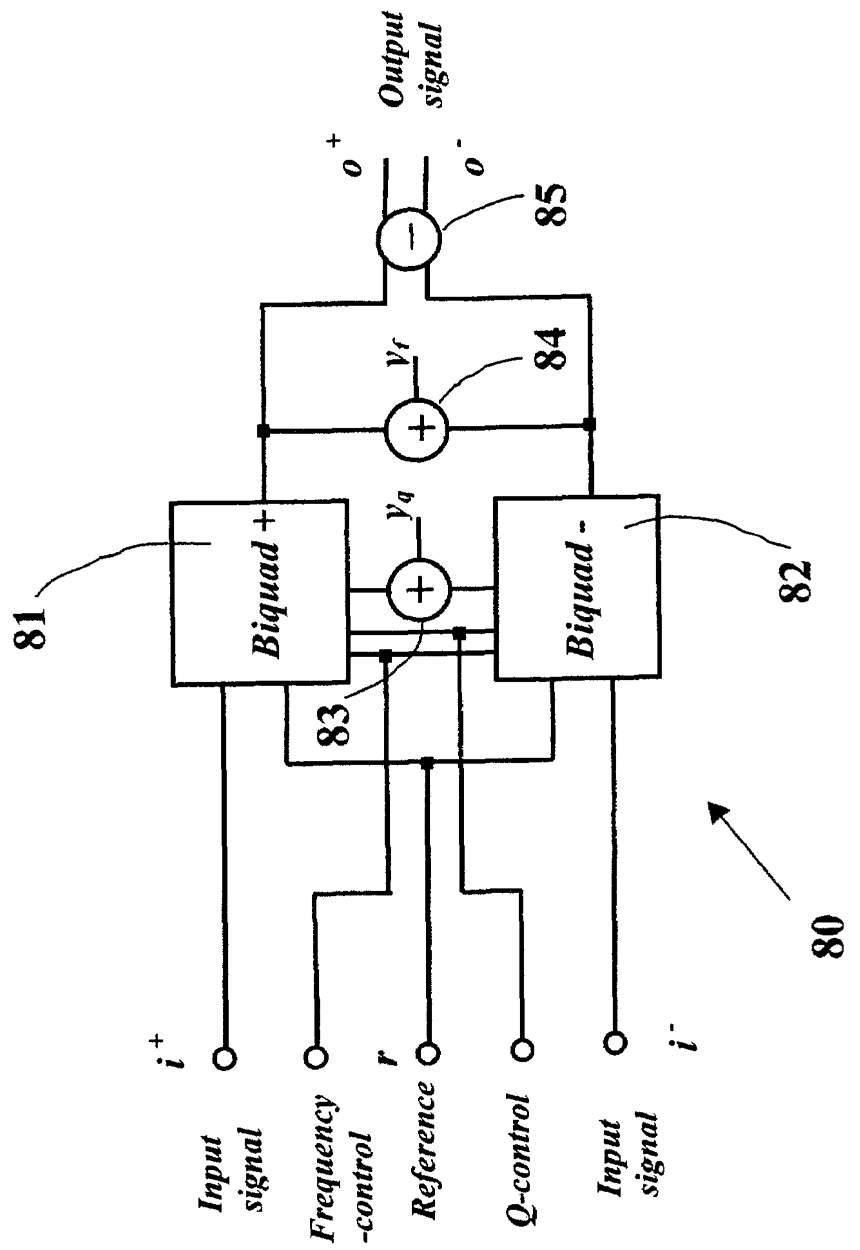
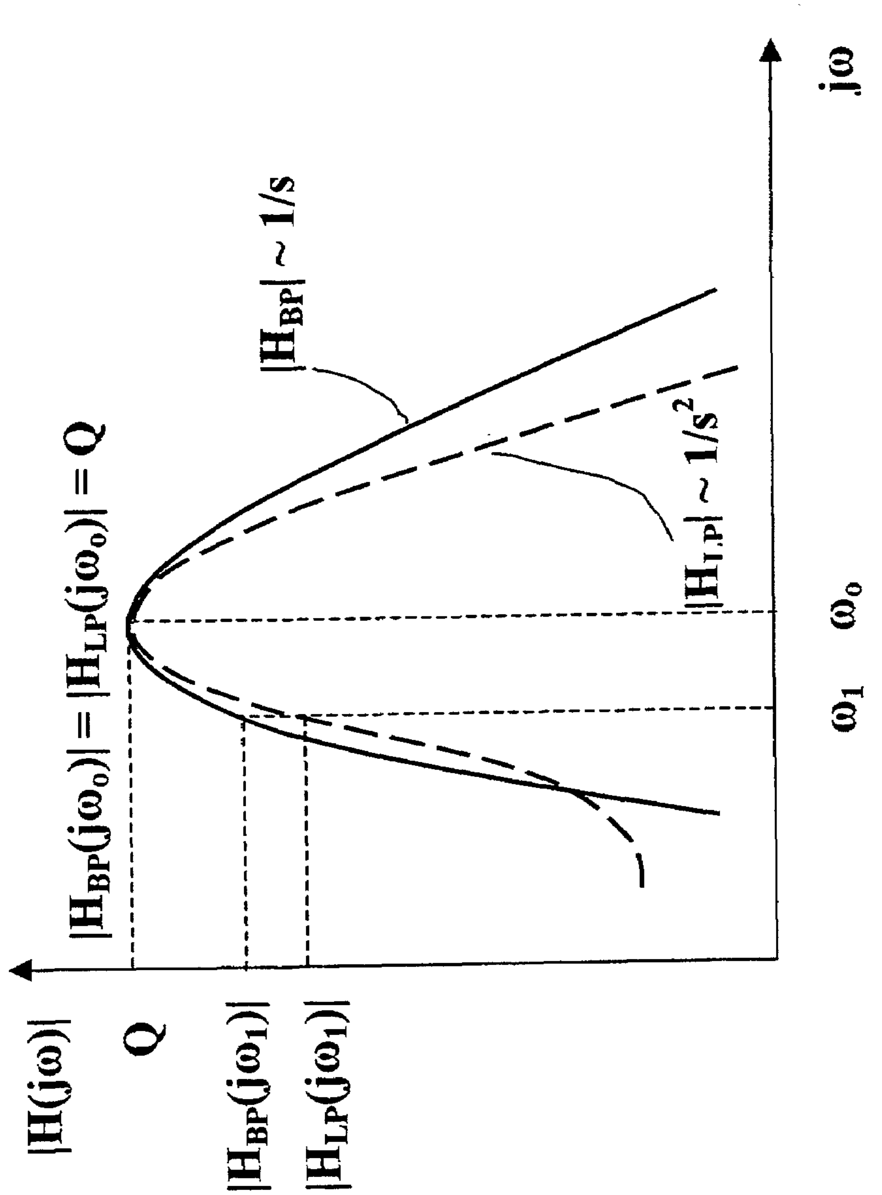


Figure 8
PRIOR ART



$|H(j\omega)|$
Q
$|H_{BP}(j\omega_1)|$
$|H_{LP}(j\omega_1)|$
$|H_{BP}(j\omega_0)| = |H_{LP}(j\omega_0)| = Q$
$|H_{BP}| \sim 1/s$
$|H_{LP}| \sim 1/s^2$
ω_1
ω_0
$j\omega$

Figure 9

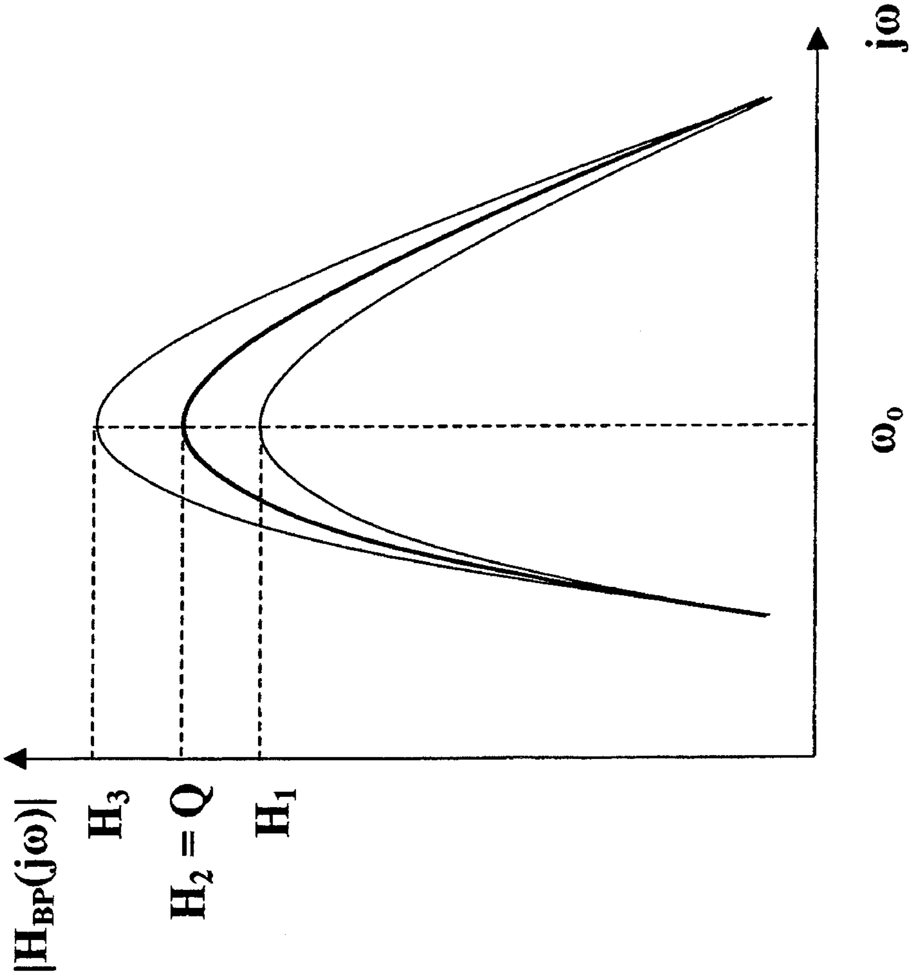


Figure 10

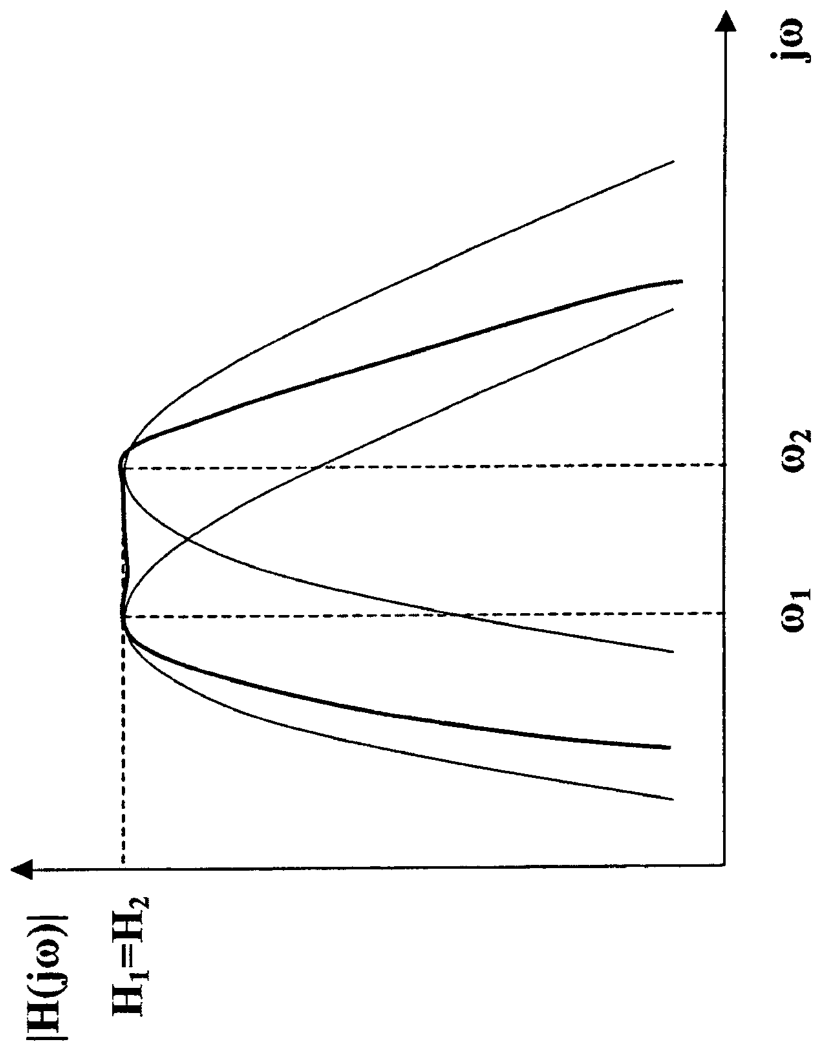
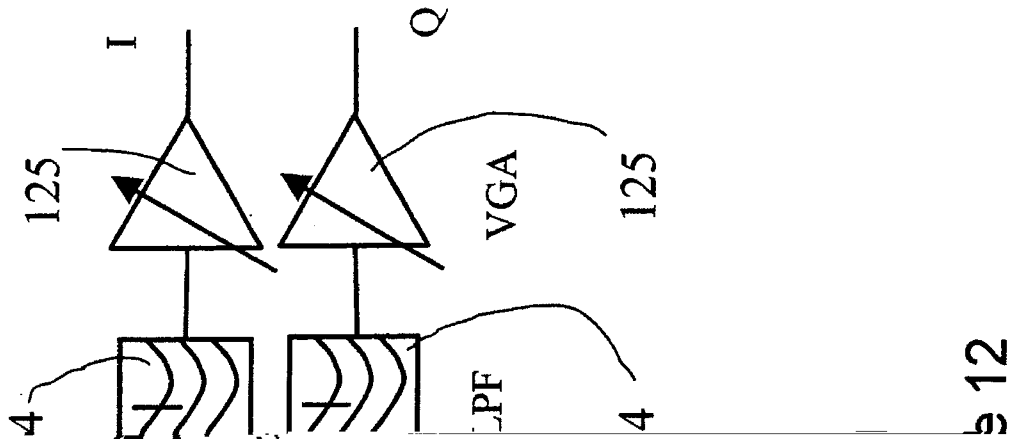


Figure 11


I
Q
125
125
VGA
4
4

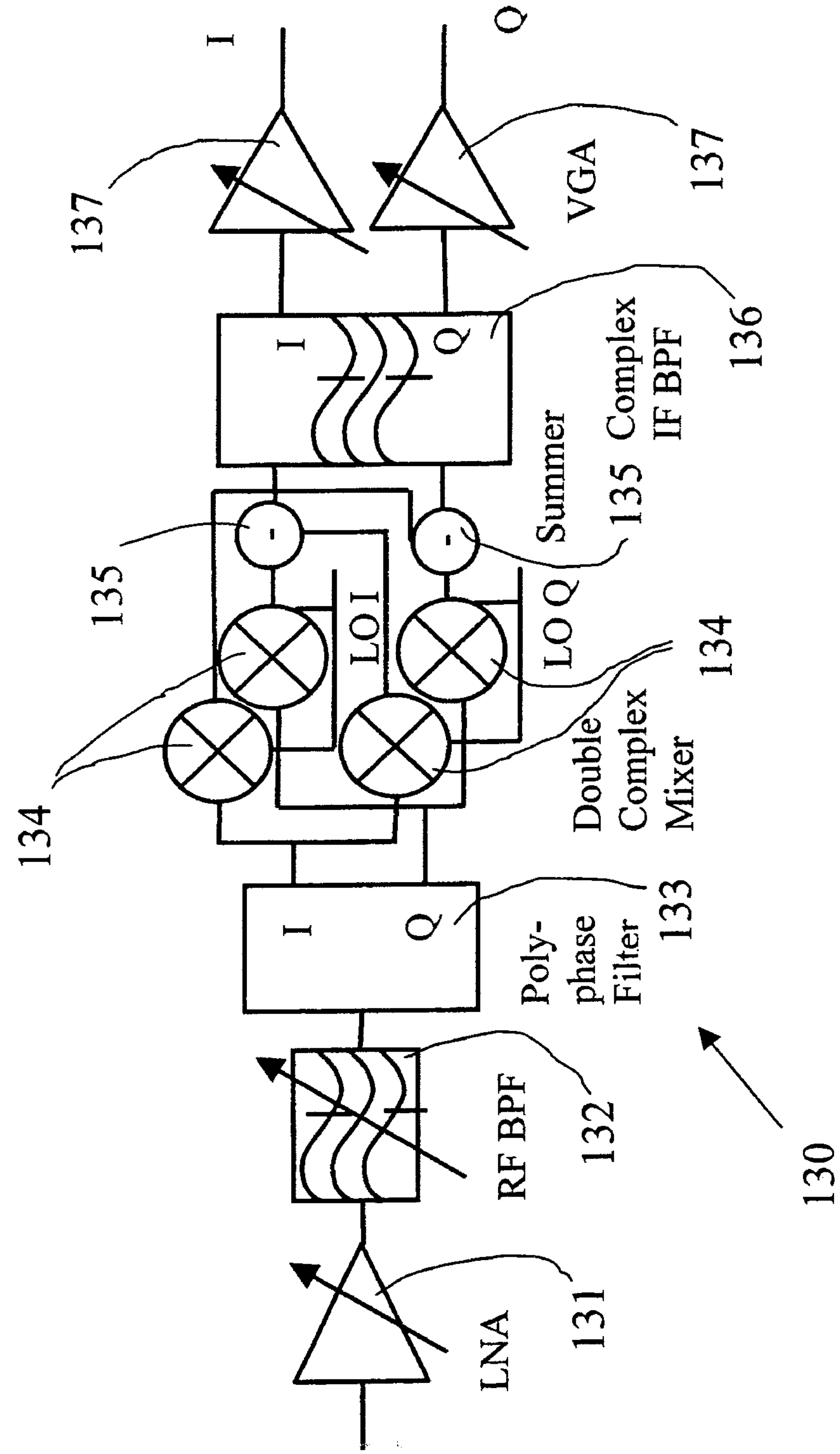

I
Q
137
137
VGA
I
Q
Complex
IF BPF
136
Summer
135
135
LO I
LO Q
134
134
Double
Complex
Mixer
I
Q
Poly-
phase
Filter
133
RF BPF
132
130
LNA
131

Figure 13

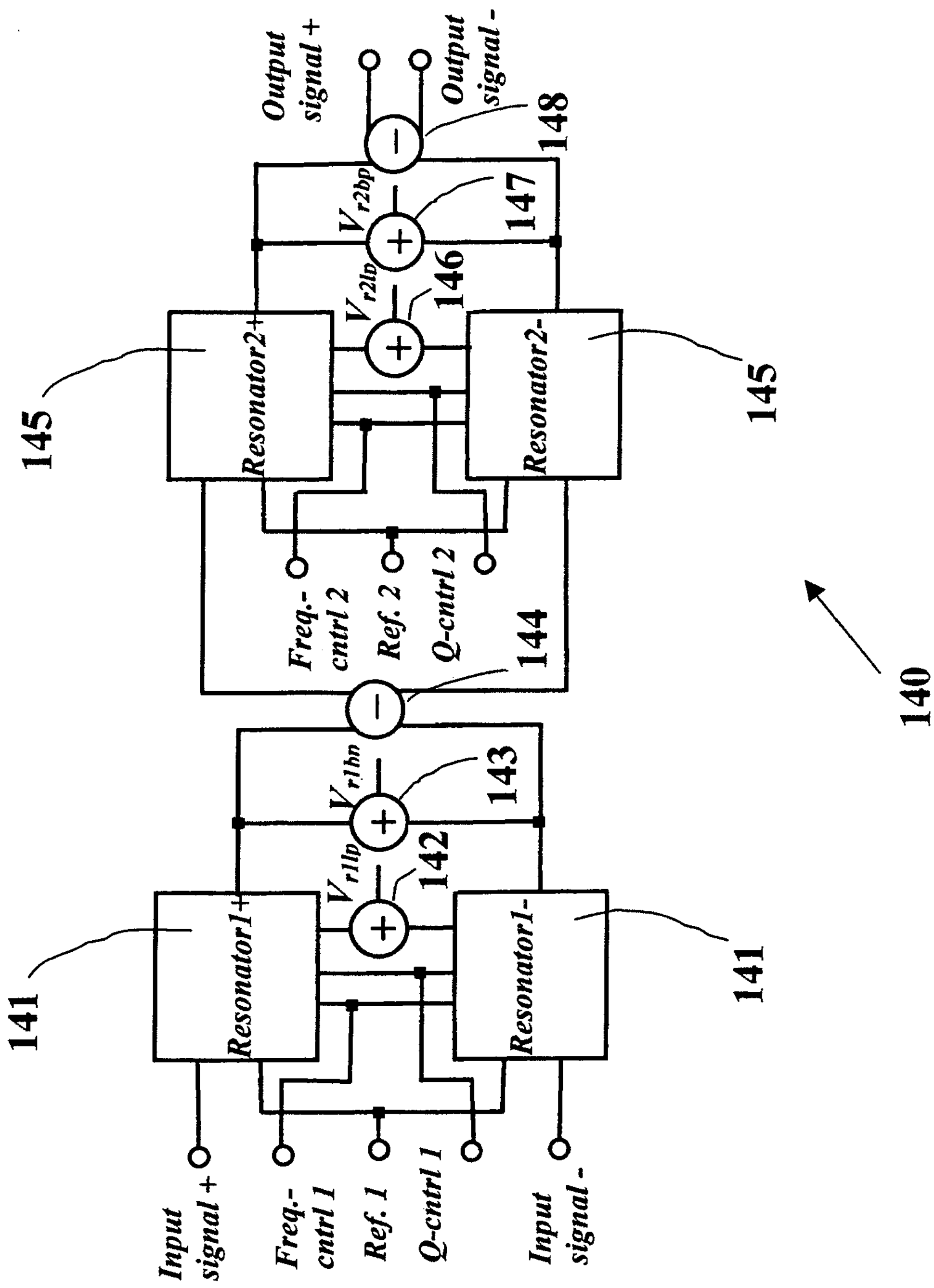

Input signal +
Freq.-cntrl 1
Ref. 1
Q-cntrl 1
Input signal -
Resonator1+
Resonator1-
141
141
V_{r1lp}
142
V_{r1hn}
143
144
Freq.-cntrl 2
Ref. 2
Q-cntrl 2
Resonator2+
Resonator2-
145
145
V_{r2lp}
146
V_{r2bp}
147
148
Output signal +
Output signal -
140

Figure 14

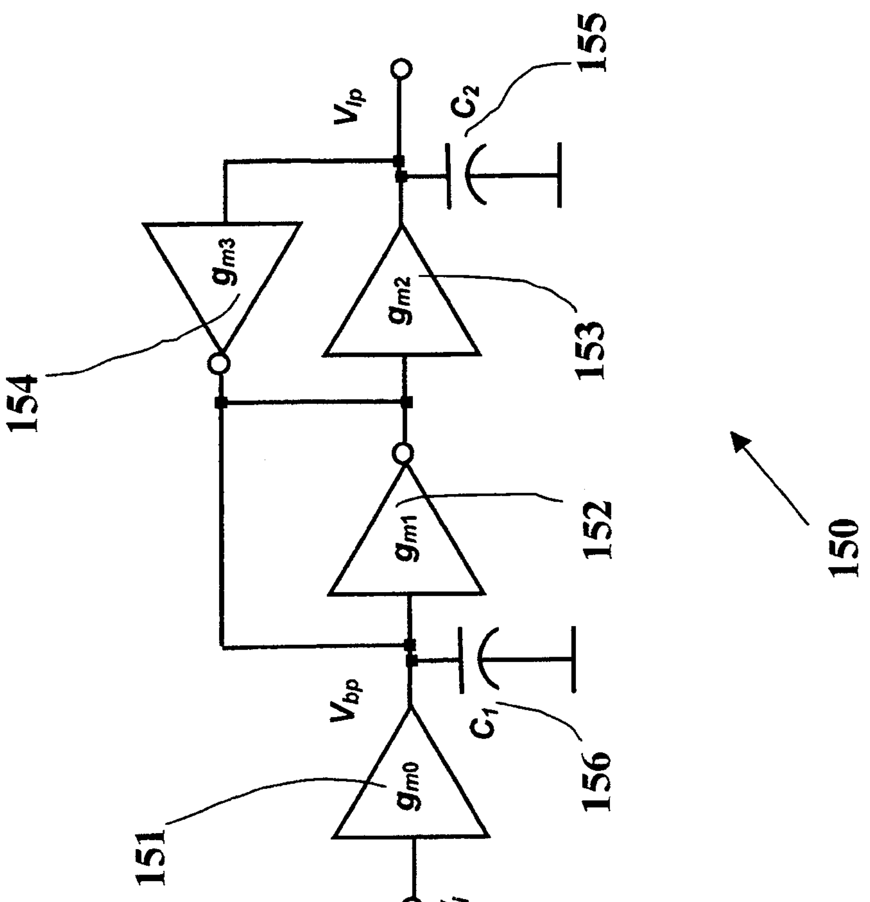


Figure 15

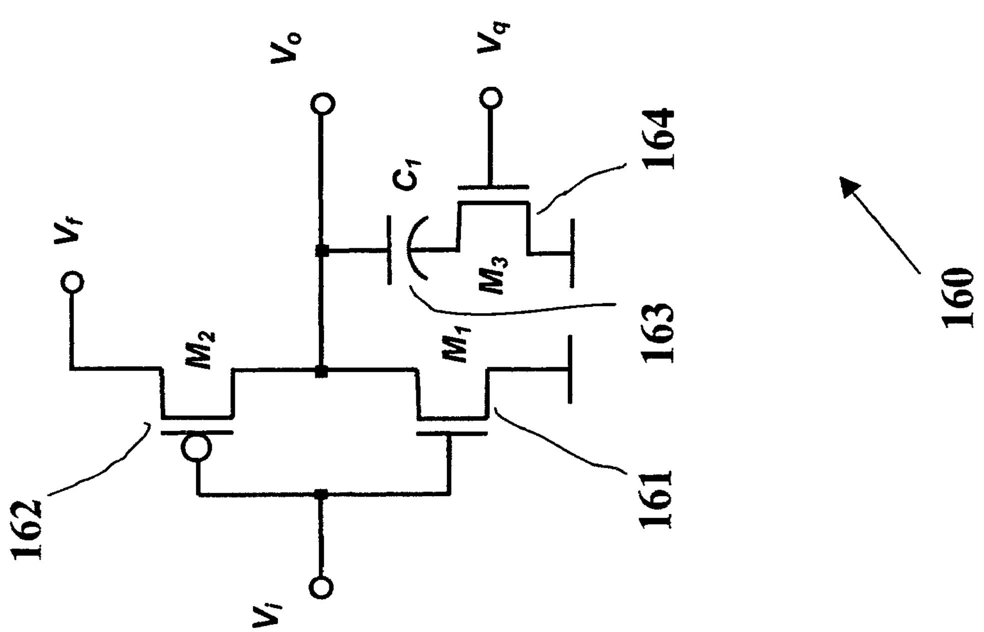


Figure 16

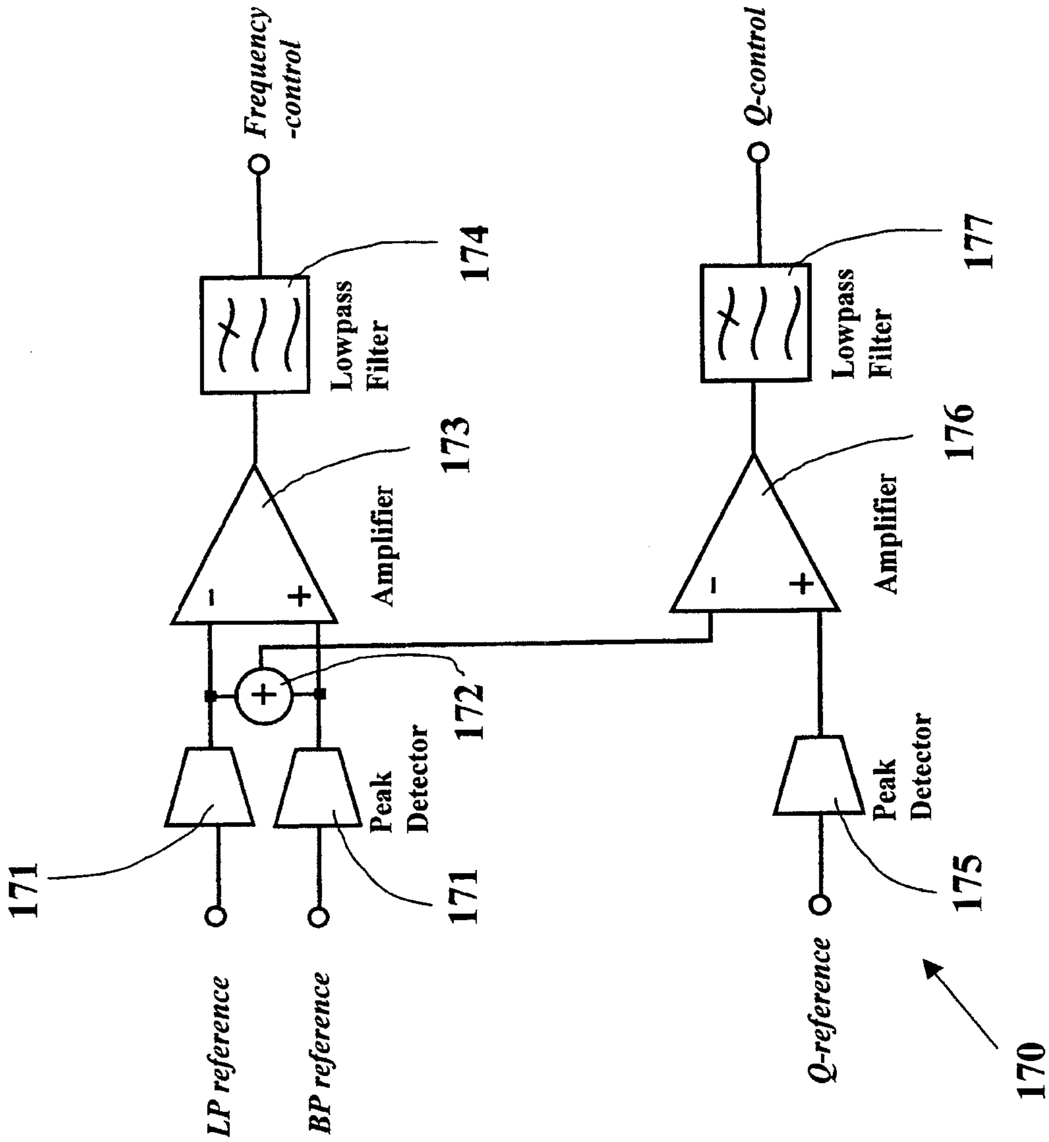
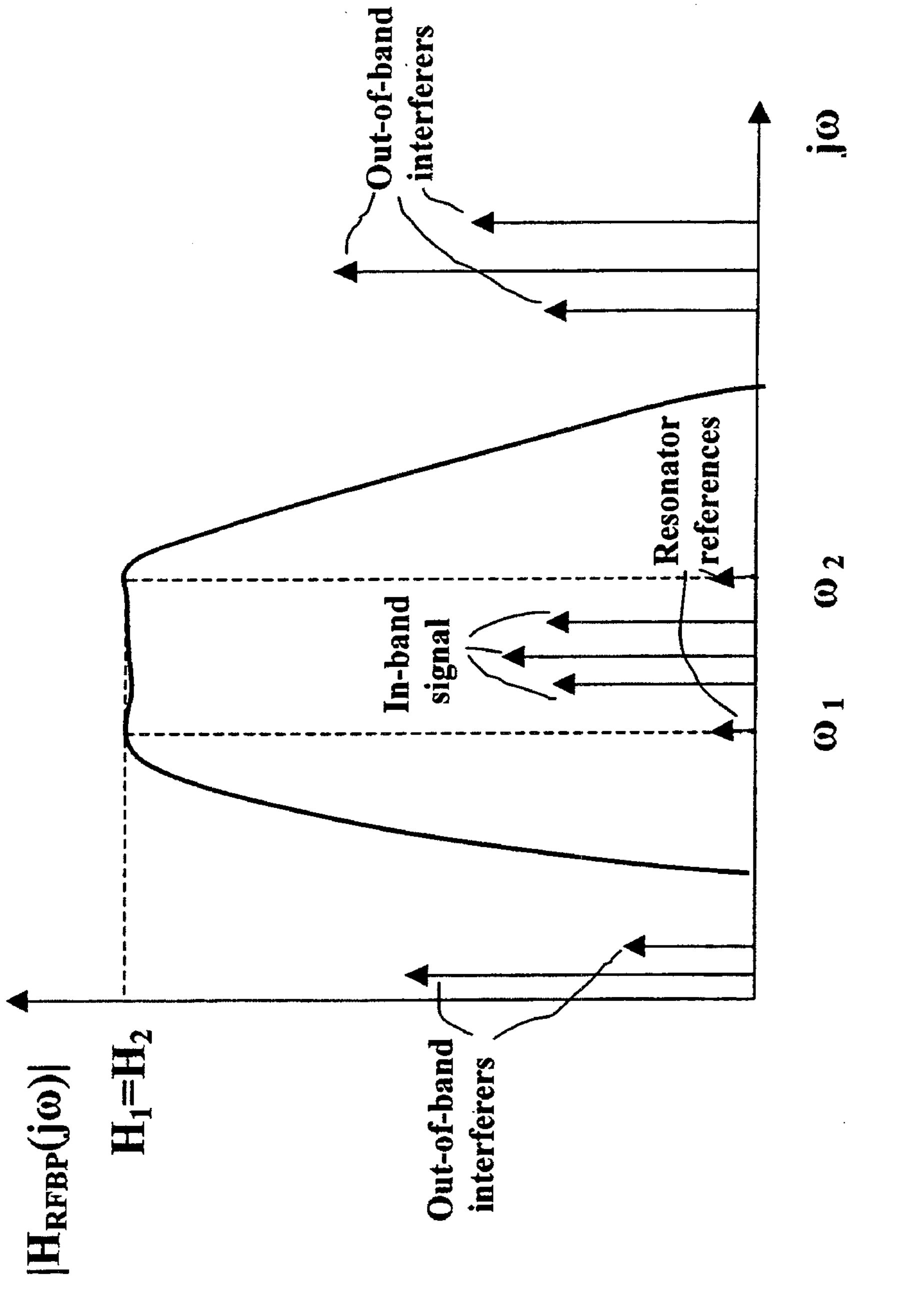

Frequency
-control
Q-control
174
177
Lowpass
Filter
Lowpass
Filter
173
176
Amplifier
Amplifier
-
+
-
+
172
171
171
Peak
Detector
Peak
Detector
175
LP reference
BP reference
Q-reference
170

Figure 17



|H_RFBP(jω)|
H1=H2
Out-of-band interferers
In-band signal
Resonator references
Out-of-band interferers
ω 1
ω 2
jω

Figure 18

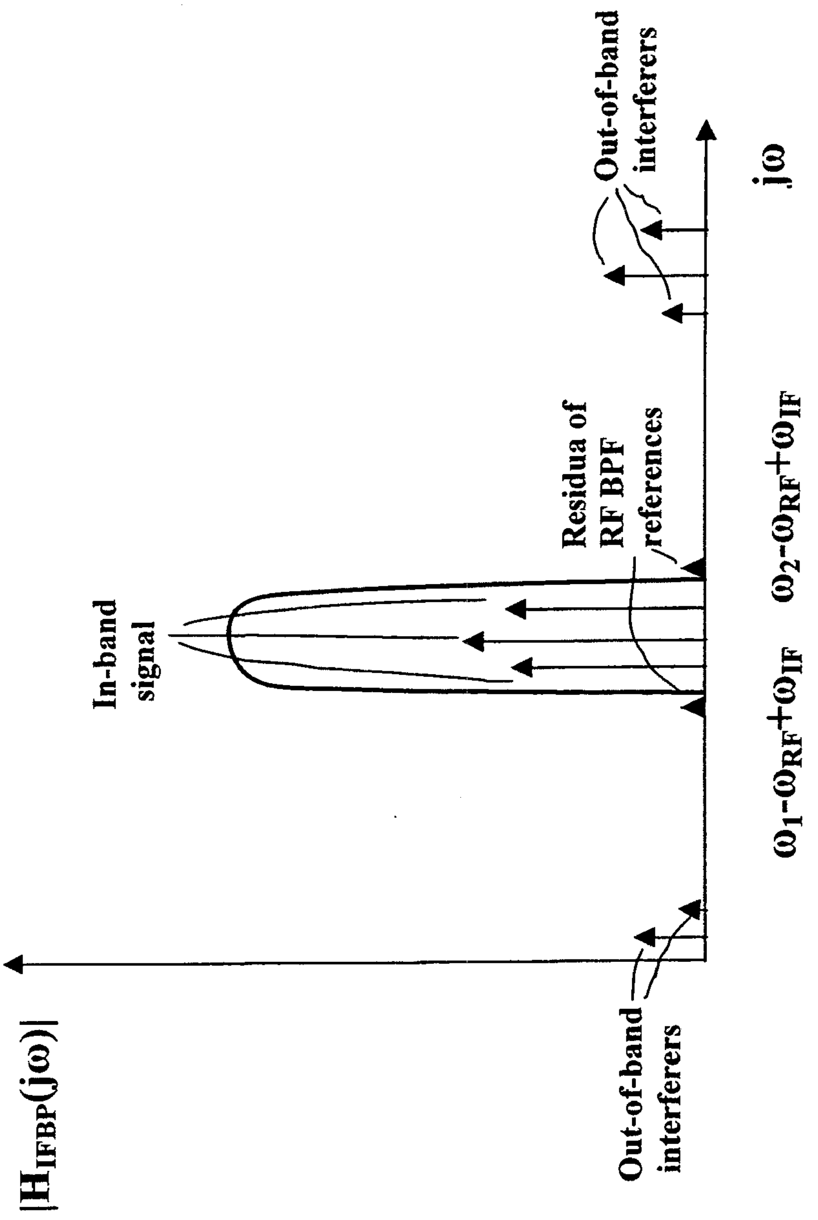

Out-of-band interferers
jω
Residua of RF BPF references
In-band signal
ω_2-ω_{RF}+ω_{IF}
ω_1-ω_{RF}+ω_{IF}
|$H_{IFBP}(j\omega)$|
Out-of-band interferers

Figure 19

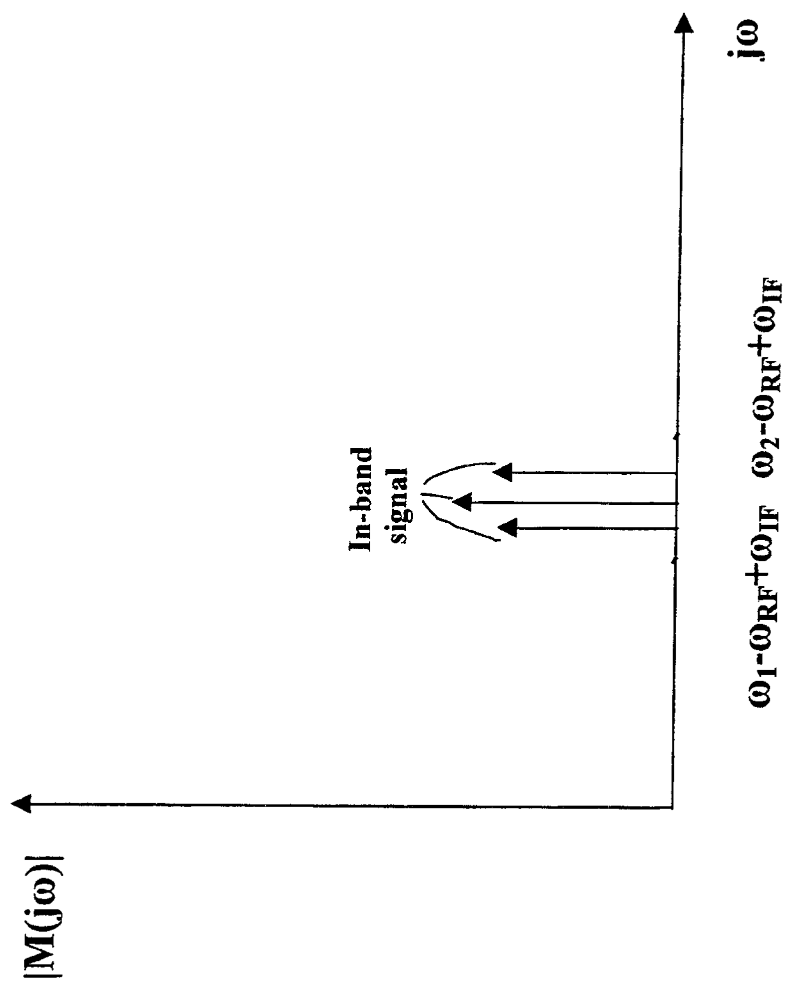

$|M(j\omega)|$
$j\omega$
In-band signal
$\omega_1-\omega_{RF}+\omega_{IF}$
$\omega_2-\omega_{RF}+\omega_{IF}$

Figure 20

SELF-TUNED ACTIVE BANDPASS FILTERS

The present invention relates to fully-integrated continuous-time active real bandpass filters and their automatic frequency- and Q-tuning systems.

BACKGROUND OF THE INVENTION

Nearly all practical transceivers require some form of filtering. Up to the date, a majority of their radio-frequency (RF) and intermediate-frequency (IF) bandpass (BP) filters are realized as off-chip ceramic or surface-acoustic wave (SAW) devices. These devices do not have a practical way to change their center frequency (CF) and/or their bandwidth (BW).

In cases when a BP filter is required to change its CF and/or its BW discrete component solutions are used that apply varactors as tunable capacitors that together with external coils (inductors) allow to tune the filter CF. Since the off-chip filter components can easily pick up noise and radio interference they need to be shielded by a metal can. This requirement unavoidably increases the cost and dimensions of the

In the case of presented BP filters instead of using the M-S scheme the tuning accuracy can be substantially improved by

with common-mode (CM) signals filter as shown in U.S. Pat. No. 5,608,665 by passing the CM reference through the filter while simultaneously processing the differential signal. The simplest configuration is obtained by two single-ended structures forming a pseudo-differential filter. Because these tuning schemes are free from the M-S matching errors the expected accuracy of self-tuned frequency- and Q-tuning systems could be as good as 0.5% and 2% respectively.

DESCRIPTION OF THE PRIOR ART

The architecture of a classical Master-Slave (M-S) tuning scheme such as one described in U.S. Pat. No. 3,997,856 is illustrated in FIG. 1, and is identified by the numeral **10**. Note that only a frequency-tuning scheme is presented in U.S. Pat. No. 3,997,856. The Q-tuning scheme is not disclosed in that patent. The frequency-control scheme consists of a Master filter, or a Master oscillator **12** followed by a frequency-tuning circuit **14** similar to that of U.S. Pat. No. 3,997,856. It is illustrated in FIG. 2 and identified by the numeral **20**. It consists of a pair of limiters **21**, a phase-detector **22**, which the M-S scheme. The reason is that while initially the wafer-probe scheme demonstrates the accuracy similar to that of the M-S, after the Master is disabled its accuracy relies entirely on the temperatures insensitive biasing. Since such biasing introduces an extra error by not perfectly tracking the temperature variations the frequency-tuning accuracy of the wafer-probe scheme can be as low as 10%, which is still useful for some less demanding applications. However, because of this low accuracy, the Q-tuning is not practical here.

The architecture of a self-tuned filter scheme is illustrated in FIG. **7**, and it is identified by the numeral **70**. It consists of two filters **71** and **72**. When the first of filters **71** and **72** has its input connected to the input signal the other one is tuned with the reference. Then using switches **73** and **74**, their roles are interchanged. The outputs of filters **71** and **72** are switched using switches **75** and **76**. The frequency reference is applied to the frequency-tuning circuit **79** that generates control signals via hold circuits **77** and **78** for tuning the one of the filters **71** and **72** that is not processing the input signal. The critical difficulty of this scheme is switching the filters on and off the signal such that transients are avoided. For this reason this architecture is not practical. However, if it was not for this problem the self-tuning has the potential of achieving higher-pairs of resonators. Additionally, since both tuning schemes rely on the amplitude detection the reference signal can be made small, which prevents the build-up of the intermodulation distortion in the filter.

BRIEF DESCRIPTION OF THE DRAWINGS

For a more complete understanding of the present invention and for further advantages thereof, reference is now made to the following Description of the Preferred Embodiments taken in conjunction with the accompanying Drawings in which:

FIG. **1** is a block diagram of a prior art Master-Slave tuning scheme;

FIG. **2** is a block diagram of a prior art frequency-tuning scheme using a phase-detector;

FIG. **3** is a block diagram of a prior art frequency-tuning scheme using a phase/frequency-detector;

FIG. **4** is a block diagram of a prior art frequency-tuning scheme using an integrator and peak-detectors;

FIG. **5** is a block diagram of a prior art amplitude detection Q-tuning scheme;

FIG. **6** is a block diagram of a prior art wafer-probing frequency-tuning scheme;

143 that rejects all differential signals and interferers and passes first CM-reference only. Next, the first resonators' BP outputs are also connected to the subtracting block **144** that rejects first CM-reference and passes all differential signals and interferers.

This differential output together with second CM-reference enters the second pair of identical single-ended resonators **145**. The second resonators' LP outputs are connected to the summing block **146** that rejects all differential signals and interferers and passes second CM-reference only. Similarly, the second resonators' BP outputs are connected to the summing block **147** that rejects all differential signals and interferers and passes second CM-reference only. Next, the second resonators' BP outputs are also connected to the subtracting block **148** that rejects second CM-reference and passes all differential signals and interferers.

Referring to the FIG. **14** without losing generality, all single-ended signals in the present automatically tuned filter system **140** can be replaced with differential signals. In particular, the two single-ended input signals and two pairs of identical single-ended resonators **141**,**145** can be replaced with their differential versions. The single-ended CM reference can be replaced by its differential version. In such a case the summers **142**, **143**, **146**, **147** and subtractors **144**, **148** have two differential inputs and one differential output, the extracted signals become double-difference and the two references become differential signals.

Referring to the FIG. **14** without losing generality, the two-resonators **141**, **145** can be replaced by three- or more resonator system. The principle of the tuning will remain the same, with additional resonators keeping their magnitudes equal to the first two resonators. Since the references should fall outside the desired band the even number of resonators is preferred to the odd number with their peak frequencies outside the desired band.

In such a case the magnitude comparison can be performed in pairs. As an illustration, for four resonators the magnitudes

-continued

$$H_{LP}(s) = \frac{\frac{g_{m0}g_{m2}}{C_1C_2}}{s^2 + s\frac{g_{m1}}{C_1} + \frac{g_{m2}g_{m3}}{C_1C_2}} \tag{2}$$

with their resonant frequency ω_o and the quality factor Q defined as:

$$\omega_o = \sqrt{\frac{g_{m2}g_{m3}}{C_1C_2}} \tag{3}$$

$$Q = \sqrt{\frac{g_{m2}g_{m3}C_1}{g_{m1}^2C_2}} \tag{4}$$

From the Q formula it is seen that the highest achievable Q is obtained when g_{m1} is replaced by the sum of output conductances $g_{o0}+g_{o3}$, which are much smaller than g_{m1}. Depending on the technology parameters and the design details the resonator Q's in the order of ten to fifty can be achieved.

A single-ended g_m-C integrator is illustrated in FIG. **16** and is generally identified by the numeral **160**. It is a building block for each of the four biquadratic resonators used in the presented filter. It consists of an inverter **161** M_1, inverter **162**, M_2 and a capacitor **163** C_1. The transistor **164** M_3 serves as a series resistor that changes the phase of the integrator and modifies a Q-factor of any filter built with it. The Q-control is applied as a gate voltage V_q controlling the gate voltage and the triode mode conductance g_{ds3} of M_3. The supply voltage V_f controls the unity gain frequency of the integrator and the center frequency of any resonator built with it. The device M_1-M_2 matching sets the dc-output voltage of the integrator which with $V_G \approx V_f/2$ becomes

$$g_m = (k_n + k_p)\frac{V_f}{2} - (k_n V_{Tn} + k_p|V_{Tp}|) \tag{8}$$

and it is easily modified by changing the frequency-tuning/supply voltage V_f.

The phase of the integrator in FIG. **16** can be tuned using Q-control voltage V_q by modifying g_{ds3} value of M_3 with $V_{DS}=0$. Using a simplified formula for the drain current of a NMOS device in triode region

$$i_D = k_n\left(v_{GS3} - \frac{v_{DS3}}{2} - V_{Tn}\right)v_{DS3} \tag{9}$$

its drain-source conductance can be calculated as:

$$g_{ds3} = \frac{\partial i_{D3}}{\partial v_{DS3}} = k(v_{GS3} - v_{DS3} - V_{Tn}) \tag{10}$$

where V_{DS3} is the drain-source voltage and all other symbols have their regular meanings. Calculating the transfer function of the integrator in FIG. **16** as:

$$\frac{v_o}{v_i} = \frac{g_{m0}\left(1 + \frac{s}{z}\right)}{sC_1\left(1 + \frac{s}{p}\right)} \tag{11}$$

where p is the internal pole of the transconductor and z is the zero introduced by g_{ds3} with its value

$$z = \frac{g_{ds3}}{C_1} = \frac{k_n(V_q - V_{Tn})}{C_1} \tag{12}$$

Writing BP and LP transfer functions for a biquad as:

$$H_{BP}(s) = \frac{s\omega_o}{s^2 + s\frac{\omega_o}{Q} + \omega_o^2} \tag{13}$$

$$H_{LP}(s) = \frac{\omega_o^2}{s^2 + s\frac{\omega_o}{Q} + \omega_o^2} \tag{14}$$

and calculating the magnitudes of BP and LP transfer functions at the resonant frequency $j\omega_o$ as:

$$|H_{BP}(j\omega_o)| = \left|\frac{j\omega_o^2}{-\omega_o^2 + j\omega_o\frac{\omega_o}{Q} + \omega_o^2}\right| = \left|\frac{j\omega_o^2}{j\omega_o\frac{\omega_o}{Q}}\right| = Q \tag{15}$$

$$|H_{LP}(j\omega_o)| = \left|\frac{\omega_o^2}{-\omega_o^2 + j\omega_o\frac{\omega_o}{Q} + \omega_o^2}\right| = \left|\frac{\omega_o^2}{j\omega_o\frac{\omega_o}{Q}}\right| = \left|\frac{Q}{j}\right|Q \tag{16}$$

it can be shown that:

$$|H_{BP}(j\omega_o)| = |H_{LP}(j\omega_o)| = Q. \tag{17}$$

Referring to the FIG. **17**, for each pair of resonators, the summer **172** extracts the average value of the resonator CM-reference BP and LP magnitudes and passes it to the Q-tuning system consisting of blocks **175**-**177**. Referring to the FIG. **10**, it forces the average resonator magnitude value at resonant frequency ω_o to be equal to that of the particular resonator desired Q.

The present filter is directly tuned with a reference signal while simultaneously operating on the main signal. By choosing appropriate input amplitude of the reference, the reference output amplitude is set to be sufficiently small to not interfere with the main signal for a given type of signal modulation.

Any viable frequency-tuning technique including, but not limited to phase detection used in phase locked-loop Type I illustrated in FIG. **2**, phase and frequency detection used in phase locked-loop Type II illustrated in FIG. **3**, or amplitude detection using unity-gain integrators illustrated in FIG. **4** can be used to implement frequency-tuning circuit in FIG. **17**.

FIG. **13** illustrates a fully integrated low-IF receiver **130** using the present filter **132**. The signal from the antenna enters the input of the low-noise amplifier (LNA) circuit **131**, the output of which is connected to the present filter **132**. The output the present filter **132** is connected to the input of polyphase filter **133** that splits the RF signal into I and Q parts shifted by 90 degrees.

The I, Q outputs of the polyphase filter **133** are connected to the four inputs of the double-complex mixer circuit **134** that consists of four identical mixers fed by a pair of input **1**, Q RF signals shifted by 90 degrees and a pair of I, Q LO signals also shifted by 90 degrees (LO I and LO Q). The outputs of double-complex mixer are subtracted by a pair of subtractors 135 which results in cancelling of image signals.

The I, Q output signals form the subtractors **135** connect to the inputs of the complex IF BP filter **136** that selects the baseband channel and further rejects the image. The I and Q outputs of filter **136** are connected to the inputs of two variable gain amplifiers (VGA's) **137**.

Referring to the FIG. **18** the present RF BP filter input

spectrum of the receivers in FIGS. **12** and **13** is illustrated. Strong out-of band interferers are present on both sides of the selected spectrum. The in-band spectrum falls into the present filter passband. Note that the two CM-references also pass through the present filter passband, but they are located outside the in-band spectrum and that their amplitude is small compared to the signal.

Referring to the FIG. **19** the input spectrum of the IF BP

5. The tuning system of claim **4** wherein said automatic frequency-tuning system includes a center-frequency and adjusts said center-frequency with respect to said common-mode reference signal.

6. The tuning system of claim **5** wherein for each pair of biquadratic resonators said automatic frequency-tuning system compares the magnitudes of a bandpass and a lowpass transfer function and locks at the frequency for which said transfer functions are equal.

7. The tuning system of claim **6** wherein said frequency-tuning system controls the magnitudes of said bandpass and said lowpass transfer functions by changing a resonance frequency through the modification of an electronically adjustable parameter that controls the resonance frequency.

8. The tuning system of claim **7** wherein said automatic Q-tuning system adjusts its magnitude, and thereby the Q-factor, with respect to a reference amplitude.

9. The tuning system of claim **8** wherein for each pair of biquadratic resonators said Q-tuning system compares the average magnitude of said bandpass and said lowpass transfer

functions at the reference frequency and locks the transfer function to the value of the Q-reference.

10. The tuning system of claim **9** wherein said Q-tuning system controls the average magnitude of said bandpass and said lowpass transfer functions at the reference frequency by electronically adjusting the phase of a resonator within said Q-tuning system.

11. The tuning system of claim **10** wherein said real band a differential output subtractor for canceling said reference [redacted]

bandpass filters are differential circuits.

23. The tuning system of claim 22 wherein said identical bandpass filters are each fed with a pair of differential input signals in anti-phase and a pair of in-phase differential reference signals.

24. The tuning system of claim 23 and further including: [redacted]

a differential input-double-difference output subtractor for canceling the reference signal and extracting the input signal.

* * * * *