



(51) International Patent Classification:
H03D 7/14 (2006.01)

(21) International Application Number:
PCT/EP2011/055101

(22) International Filing Date:
1 April 2011 (01.04.2011)

(25) Filing Language: English

(26) Publication Language: English

(30) Priority Data:
10290191.5 9 April 2010 (09.04.2010) EP

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(81) Designated States (unless otherwise indicated, for every kind of national protection available): AE, AG, AL, AM,

AO, AT, AU, AZ, BA, BB, BG, BH, BR, BW, BY, BZ, CA, CH, CL, CN, CO, CR, CU, CZ, DE, DK, DM, DO, DZ, EC, EE, EG, ES, FI, GB, GD, GE, GH, GM, GT, HN, HR, HU, ID, IL, IN, IS, JP, KE, KG, KM, KN, KP, KR, KZ, LA, LC, LK, LR, LS, LT, LU, LY, MA, MD, ME, MG, MK, MN, MW, MX, MY, MZ, NA, NG, NI, NO, NZ, OM, PE, PG, PH, PL, PT, RO, RS, RU, SC, SD, SE, SG, SK, SL, SM, ST, SV, SY, TH, TJ, TM, TN, TR, TT, TZ, UA, UG, US, UZ, VC, VN, ZA, ZM, ZW.

(84) Designated States (unless otherwise indicated, for every kind of regional protection available): ARIPO (BW, GH, GM, KE, LR, LS, MW, MZ, NA, SD, SL, SZ, TZ, UG, ZM, ZW), Eurasian (AM, AZ, BY, KG, KZ, MD, RU, TJ, TM), European (AL, AT, BE, BG, CH, CY, CZ, DE, DK, EE, ES, FI, FR, GB, GR, HR, HU, IE, IS, IT, LT, LU, LV, MC, MK, MT, NL, NO, PL, PT, RO, RS, SE, SI, SK, SM, TR), OAPI (BF, BJ, CF, CG, CI, CM, GA, GN, GQ, GW, ML, MR, NE, SN, TD, TG).

Published:

- with international search report (Art. 21(3))
- before the expiration of the time limit for amending the claims and to be republished in the event of receipt of amendments (Rule 48.2(h))

(54) Title: LEAKAGE CALIBRATION FOR A MIXER CIRCUIT

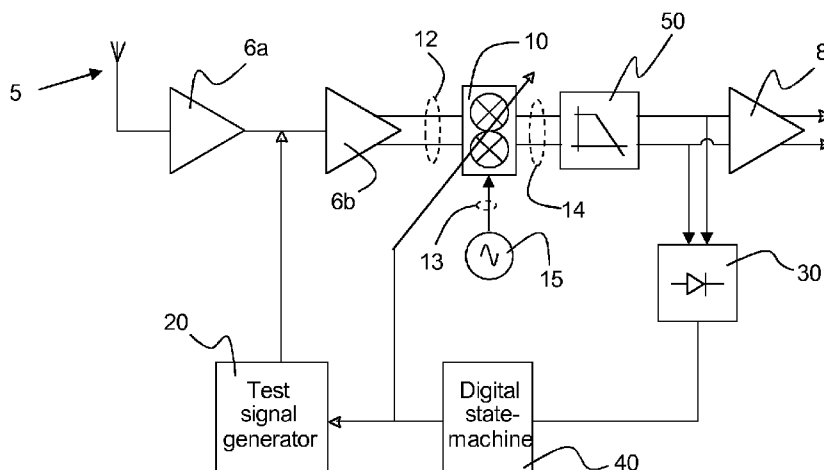


FIG 1

(57) Abstract: A mixer with a calibration circuit. The mixer has a signal input port for a signal to be up- converted or down-con- verted; a local oscillator input port; and an output port. The mixer is for use in a transmitter or receiver in which the spectrum of the signal at the input port and the spectrum of a mixed signal at the output port overlap at least partially. The calibration circuit comprises: a test signal generator, operable to inject a first test signal at the signal input port; a measurement circuit, adapted to measure a first output signal at the output port resulting from the injected first test signal; and a control circuit, adapted to modify at least one voltage-to- current conversion parameter at the input port of the mixer based on the measurement, so as to reduce leakage from the signal input port to the output port. Also provided is a method of calibrating the mixer.

TITLE

LEAKAGE CALIBRATION FOR A MIXER CIRCUIT

FIELD OF THE INVENTION

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The present invention relates to analogue mixer circuits for mixing an input signal with a Local Oscillator (LO) signal in order to either up-convert or down-convert the input signal. It is particularly relevant to applications in which the input to the mixer and the mixed output overlap in the frequency domain.

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BACKGROUND OF THE INVENTION

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A mixer is a non-linear circuit that multiplies (or “mixes”) two inputs to produce an output that is their product. In an ideal mixer, the output would consist purely of this product signal; however, in practice, this desired output is also accompanied by unwanted components, such as higher-order products and noise.

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A key application of mixers is in communications systems: at the transmitter they are used to convert base-band signals into Radio Frequency (RF) signals, known as up-conversion; and at the receiver they are used to down-convert the RF signal back to base-band, or at least to an Intermediate Frequency (IF) signal that can be more easily processed than RF. Receivers which convert directly to base-band are known as direct conversion or low-IF receivers. These contrast with superheterodyne receivers, in which the IF output is at a significantly higher frequency than base-band. Up- and down- conversion are achieved by mixing the data signal with a carrier or local oscillator signal. In the frequency domain, this has the effect of translating the frequency of the data signal by the frequency of the LO. Both sum and difference frequencies are generated. For up-conversion, the sum frequencies are the desired output, so the difference frequencies are attenuated by filtering. Conversely, for down-conversion, at the receiver, the difference frequencies are the desired output, so the sum frequencies are filtered out. The filtering also helps mitigate the other unwanted, out-of-band signals, such as higher-order inter-modulation products.

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It is known that port isolation between the LO and RF ports is important (see, for example, K.L. Fong and R.G. Meyer, “Monolithic RF Active Mixer Design”, IEEE Transactions on Circuits and Systems II: Analog and Digital Signal Processing, 46(3) March 1999). This is because LO-to-RF feed-through results in LO signal leaking to the antenna. If this happens at a receiver, it could cause Electro-Magnetic Interference (EMI) for other nearby systems. If it

happens at the transmitter, the presence of LO leakage reduces the signal to noise ratio of the transmitted signal. Port isolation from LO-to-IF and RF-to-IF (that is, direct feed-through between input to output) has previously been considered unimportant, because the resulting leakage signals could be rejected by filtering.

5 US 2006/128344 and WO 03/079573 disclose calibration circuits in which a variable load at the output of a mixer is modified in order to improve the linearity of the mixer.

SUMMARY OF THE INVENTION

10 According to an aspect of the present invention, there is provided a circuit comprising:
a mixer having a signal input port for a signal to be up-converted or down-converted; a local oscillator input port; and an output port, for a transmitter or receiver in which the spectrum of the signal at the signal input port and the spectrum of a mixed signal at the output port overlap at least partially,

15 wherein at least one of the signal input port and the local oscillator input port is a differential input port and the output port is a differential output port,

the mixer comprising a voltage-to-current conversion part at the input port, having at least one modifiable voltage-to-current conversion parameter; and

a calibration circuit comprising:

20 a test signal generator, operable to inject a first test signal at the signal input port;

a measurement circuit, adapted to measure a first output signal at the output port resulting from the injected first test signal; and

a control circuit, adapted to modify the at least one voltage-to-current conversion parameter of the voltage-to-current conversion part based on the measurement, so as to
25 reduce leakage from the signal input port to the output port.

If the mixer is performing up-conversion, in a transmitter, the input signal will be a baseband or intermediate frequency (IF) signal; and the output signal will be an RF signal. If the mixer is performing down-conversion, at a receiver, the input signal will be the RF signal; and the output signal will be the IF signal.

30 The test signal may comprise a test tone, more preferably a sinusoidal test tone. The voltage-to-current conversion parameter of the input port of the mixer can be any suitable property that is variable and can influence the manner in which the RF signal leaks to the output port.

The measurement circuit may measure the power of the output signal. By measuring
35 the output produced by a known test signal, the measurement circuit can provide information about how the input is leaking to the output.

The control circuit uses this information to optimise the mixer circuit, by modifying one or more parameters of its input stage, so that the leakage is reduced. The controller may comprise a finite state machine.

5 The present inventors have recognised that in certain applications, the spectra of the input to the mixer and its output (an up- or down-converted version of the input) can overlap. In these circumstances, it is no longer possible to reject input-output leakage by filtering, because to do so would also attenuate part of the desired output spectrum. Instead, embodiments of the present invention enable calibration of a mixer, so that such leakage can be minimised by correcting or compensating for imperfections or imbalances in the mixer. This is achieved by
10 modifying a voltage-to-current conversion characteristic at the input stage of the mixer, with the effect that any leakage from input to output appears (at worst) as a common-mode signal at the differential output port. Common-mode leakage can be neglected, if the output is differential.

In general, a mixer circuit comprises a voltage-to-current converter element at the input
15 port. The output of this voltage-to-current converter is connected to one input of a switching element or multiplier element, which performs the actual mixing function. Another input of the switch/multiplier is connected to the local oscillator input. The output of the switch/multiplier drives a current-to-voltage converter at the output port of the mixer.

The switch/multiplier element performing the mixing operation may comprise a
20 transistor. The voltage-to-current converter at the input may comprise an input resistor in a passive mixer, or an input transistor in an active mixer. The voltage-to-current conversion characteristics are thus determined by the resistance and transconductance, respectively. The current-to-voltage converter at the output port of the mixer may comprise a load impedance.

Preferably, the test signal generator is further operable to inject a second test signal at
25 the signal input port; the measurement circuit is further adapted to measure a second output signal at the output port resulting from the injected second test signal; and the control circuit is adapted to compare the measurement of the first output signal with the measurement of the second output signal, and to modify the at least one parameter of the mixer based on the outcome of the comparison, wherein a frequency of the second test signal differs from a
30 frequency of the first test signal.

The first test signal may produce a reference output and the second test signal may then produce a measurement which can be compared with this reference, in order to estimate the leakage between the signal input and output ports.

The signal input port may be a differential input, comprising a positive terminal and a
35 negative terminal.

The mixer may be a direct conversion or low-IF mixer.

Optionally, the at least one parameter that the control circuit is adapted to modify includes a first parameter associated with the positive terminal and a second parameter associated with the negative terminal.

5 In this way, the control circuit can reduce leakage by reducing or compensating for imbalances in the mixer.

The voltage-to-current conversion part may comprise at least one variable impedance element, whose value is modifiable by the control circuit, wherein the at least one parameter that the control circuit is adapted to modify comprises the value of the at least one variable impedance.

10 A variable impedance element, such as a programmable resistance or digital potentiometer, can be one simple but effective way of calibrating the mixer circuit to reduce leakage – in particular, if the mixer is a passive mixer.

The voltage-to-current conversion part may comprise at least one variable transconductance element, whose value is modifiable by the control circuit, wherein the at least one parameter that the control circuit is adapted to modify comprises the value of the at least one variable transconductance.

A variable transconductance element, such as an array of switched transistors, can be another suitable way of calibrating the mixer circuit to reduce leakage – in particular, for active mixers.

20 The mixer circuit preferably exhibits a signal input to output (e.g. RF to IF) leakage rejection of at least 50 dBc, once it has been calibrated, more preferably at least 60 dBc.

The circuit may be used to particular advantage in a receiver (for example, a cable modem) designed for DOCSIS 3.0. This standard (ITU-T Recommendation J.222) uses channel bonding to provide transfer at high data rates. For this reason, it demands broadband tuners, typically having a bandwidth of about 100MHz. For applications of this kind, any leakage between the RF (input) and IF (output) can have a significant impact on the performance.

Also provided in accordance with another aspect of the invention is a method of calibrating a mixer circuit having a signal input port for a signal to be up-converted or down-converted; a local oscillator input port; and an output port, for a transmitter or receiver in which the spectrum of the signal at the signal input port and the spectrum of a mixed signal at the output port overlap at least partially,

wherein at least one of the signal input port and the local oscillator input port is a differential input port and the output port is a differential output port,

35 the mixer circuit comprising a voltage-to-current conversion part at the input port, having at least one modifiable voltage-to-current conversion parameter,

the method comprising:

generating a first test signal at the signal input port;

measuring a first resulting signal at the output port; and

modifying the at least one voltage-to-current conversion parameter of the voltage-to-

5 current conversion part based on the measurement, so as to reduce leakage from the signal input port to the output port.

The method may further comprise: generating a second test signal at the signal input port; measuring a second resulting signal at the output port; comparing the first and second resulting signals; and modifying the at least one parameter based on the result of the
10 comparison, wherein a frequency of the second test signal differs from a frequency of the first test signal.

These further steps can be performed after the step of measuring the first resulting signal at the output port. Thus, in some embodiments, the application of the test tones is sequential.

15 The method may further comprise applying a local oscillator signal at the local oscillator input port when applying the first and second test signals, wherein the second test signal has a frequency equal to the intermediate frequency resulting from the mixing of the first test signal with the local oscillator signal.

That is, the frequency of the second test signal is equal to the difference between the
20 frequency of the first test signal and the local oscillator frequency.

The first test signal can be chosen to have a frequency at which the input and output spectra (RF and IF, or IF and RF, respectively) do not overlap. In this case, any leakage will be out-of-band and will therefore be removed by the filtering. This produces a reference output signal which contains a "pure" mixed signal, free of leakage.

25 The frequency of the second test signal can then be chosen to be the same as the "pure" reference output produced by the first test. This means that the leakage component appearing at the output is at the same frequency as the reference signal.

By comparing the first (reference) output with the second output, the controller can obtain an estimate of the leakage power, which is then used to optimise the mixer.

30 The step of modifying the at least one parameter may comprise adjusting the parameter and assessing whether the leakage from the signal input port to the output port is reduced.

For example, the method can comprise incrementing or decrementing the parameter and measuring the effect on leakage.

35 Optionally, the signal input port is a differential input, comprising a positive terminal and a negative terminal.

The at least one parameter may then include a first parameter associated with the positive terminal and a second parameter associated with the negative terminal.

The step of modifying the at least one parameter may comprise alternately adjusting the first parameter and the second parameter.

5 This is one advantageous strategy for searching for an optimum, at which leakage is minimised.

Preferably, the step of modifying the at least one parameter is performed iteratively.

The algorithm may proceed by making successive incremental changes to the parameter as long as the leakage continues to be reduced, or until a predetermined (low)
10 target level of leakage is achieved.

The at least one parameter that is modified may comprise the value of an input impedance or the value of an input transconductance.

Also provided is a computer program comprising computer program code means adapted to control a calibration circuit to perform all the steps of the methods described above
15 if said program is run on a computer.

BRIEF DESCRIPTION OF THE EMBODIMENTS

Embodiments of the invention are described in more detail and by way of non-limiting
20 examples with reference to the accompanying drawings, wherein

Fig. 1 is a block diagram of a receiver including a calibration circuit for a mixer according to an embodiment of the invention;

Fig. 2 is a circuit diagram of a mixer with differential inputs and outputs;

Fig. 3 shows an equivalent model for a mixer like that of Fig. 2, where resistors have
25 been replaced with variable resistances, according to an embodiment;

Fig. 4 illustrates how a digitally-programmable variable resistor can be implemented;

Fig. 5 is a flowchart showing a calibration method according to an embodiment;

Fig. 6 is a flowchart of an algorithm for optimising an impedance value, for use in the calibration method of Fig. 5;

30 Fig. 7 is a schematic of a generalised embodiment of the invention;

Fig. 8 illustrates the relationship between the generalised embodiment of Fig. 7 and the passive mixer of Figs. 2-3;

Fig. 9 illustrates the relationship between the generalised embodiment of Fig. 7 and an active mixer;

35 Fig. 10 is a circuit diagram of a variable transconductance element for an active mixer according to a second embodiment of the invention; and

Fig. 11 is a circuit diagram of an active mixer with differential inputs, according to that second embodiment.

It should be understood that the same reference numerals are used throughout the Figures to indicate the same or similar parts.

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DETAILED DESCRIPTION OF THE DRAWINGS

Many applications require the reception of several channels simultaneously. Watching and recording TV, and cable modems conforming to DOCSIS 3.0 are two examples. For a DOCSIS 3.0 cable modem, the IF bandwidth can be as great as 100 MHz. Such a wide IF spectrum can cause difficulties in low IF and direct conversion receivers, because in this case part of the RF spectrum is at the same frequency as the IF spectrum. A low-IF architecture is already used in some TV tuners. The use of a low IF architecture for a DOCSIS3.0 compliant receiver is problematic, because in case of RF to IF leakage due to the mixer, a fraction of the RF signal is superimposed with the wanted IF signal. The signal to noise ratio at the receiver output is severely impacted by this RF to IF leakage.

Fig. 1 shows a calibration circuit according to a first embodiment of the invention. This calibration circuit is shown as part of a receiver. The receiver includes a mixer 10 having a signal input port 12 for a signal to be down-converted; a local oscillator input port 13; and an output port 14. A local oscillator 15 is connected to the local oscillator input port 13.

The calibration circuit comprises a test signal generator 20, operable to inject a first test signal at the signal input port 12; a measurement circuit 30, adapted to measure a first output signal at the output port 14 resulting from the injected first test signal; and a control circuit 40, adapted to modify at least one parameter of the mixer 10 based on the measurement, so as to reduce leakage from the signal input port 12 to the output port 14.

In the embodiment of Fig. 1, the mixer 10 is shown as having differential input 12 and output 14. The receiver also comprises an antenna 5, feeding amplifier stages 6a and 6b with the received RF signal. The output of the second amplifier stage 6b is coupled to the input port 12 of the mixer. The input of this amplifier 6b can be coupled to the output of the test signal generator 20, for injection of the test signal to the mixer input 12, during calibration. A filter 50 is coupled to the output 14 of the mixer 10. In this embodiment, the receiver implements a low-IF tuner; therefore, the filter 50 is low-pass. The output of the filter 50 is coupled to an output stage amplifier 8. The output of the filter 50 can also be monitored by the measurement circuit 30, during calibration. In this example, the measurement circuit is a power detector 30. It will be well known to those skilled in the art how to construct a power detector circuit 30 of a suitable kind. The output of the measurement circuit 30 is connected to the control circuit 40,

which in this embodiment is a digital state machine 40. This is connected to, and is able to control, the test-signal generator 20 and the mixer 10.

Fig. 2 shows a circuit diagram for a passive mixer with differential input ports 12, 13 and output port 14. Each port comprises a positive terminal and a negative terminal. The positive terminal of the RF input port 12 is connected to resistors R1 and R3. Meanwhile, the negative terminal of the RF input 12 is connected to resistors R2 and R4. Each resistor R1-R4 is connected (in series) to the drain of a corresponding transistor M1-M4. The gates of these transistors are driven by the local oscillator input 13. In particular, the positive terminal 13a of the LO input 13 is connected to the gate of two transistors M1, M4. The negative terminal 13b of the LO input 13 is connected to the gate of the other two transistors M2, M3. The source of transistor M1 and the source of transistor M2 are connected to the input of an amplifier, the output of which is the positive terminal of the output port 14. The source of transistor M3 and the source of transistor M4 are connected to the input of another amplifier, the output of which is the negative terminal of the output port 14. As can be seen from Fig. 2, the essential unit of the passive mixer is an input resistor R1-R4 connected to a switching element (in this case a MOSFET, M1-M4).

Fig. 3 shows simplified equivalent circuit for a passive mixer similar to that of Fig. 2, suitable for use with a calibration circuit according to an embodiment of the invention. The switching elements M1-M4 are modelled by their on-resistances Ron1-Ron4. According to this embodiment, the input resistors are variable, so that their values can be optimised to suppress leakage of the RF signal directly to the output port.

In the following equations the currents flowing at the positive and negative terminals of the output port 14 will be derived. Here, α represents the duty cycle of the LO signal, such that if $\alpha=1$, the LO duty cycle=50%. Only the terms due to RF to IF leakage are included, for clarity:

$$I = \frac{V_{RF+} \cos \omega_{RF+} t}{R_1 + Ron_1} \left(\frac{1}{2} + \dots \right) + \frac{-V_{RF-} \cos \omega_{RF-} t}{R_2 + Ron_2} \left(\frac{1}{2\alpha} + \dots \right) \quad (1)$$

$$= \frac{V_{RF+} \cos \omega_{RF+} t}{2(R_1 + Ron_1)} - \frac{V_{RF-} \cos \omega_{RF-} t}{2\alpha(R_2 + Ron_2)} + \dots$$

$$In = \frac{V_{RF+} \cos \omega_{RF+} t}{R_3 + Ron_3} \left(\frac{1}{2\alpha} + \dots \right) + \frac{-V_{RF-} \cos \omega_{RF-} t}{R_4 + Ron_4} \left(\frac{1}{2} + \dots \right) \quad (2)$$

$$= \frac{V_{RF+} \cos \omega_{RF+} t}{2\alpha(R_3 + Ron_3)} - \frac{V_{RF-} \cos \omega_{RF-} t}{2(R_4 + Ron_4)} + \dots$$

Here, V_{RF} and ω_{RF} represent the amplitude and the frequency of the RF input signal, respectively.

If the components are matched, such that $R_1=R_2$, $R_3=R_4$, $R_{on1}=R_{on2}$, and $R_{on3}=R_{on4}$, this mixer does not have any RF to IF leakage, provided the RF and LO signals are perfect – that is, provided $V_{RF+}=V_{RF-}$, $\omega_{RF+}=\omega_{RF-}$, and $\alpha=1$. In real circuits, of course, imperfections will occur and RF to IF leakage will exist, because the conditions are no longer met.

In case of mismatch of the switching elements, the ON resistances will be different. A scenario in which both the switching elements and the input resistors have mismatch will now be considered. If $R_{on1} \neq R_{on2}$, and $R_1 \neq R_2$, but if all other components are matched and if the RF and LO signals are perfect ($R_3=R_4$, $R_{on3}=R_{on4}$, $V_{RF+}=V_{RF-}=V_{RF}$, $\omega_{RF+}=\omega_{RF-}=\omega_{RF}$, $\alpha=1$), then the RF to IF leakage becomes:

$$I_{leakage} = \frac{V_{RF} \cos \omega_{RF} t}{2} \left(\frac{1}{R_1 + R_{on1}} - \frac{1}{R_2 + R_{on2}} \right)$$

Consequently in case of mismatch of the mixer components, the RF to IF leakage is cancelled out for the following value of R_1 :

$$R_1 = R_2 + R_{on2} - R_{on1}$$

Imbalance of the RF signal can also be considered. If $V_{RF+} \neq V_{RF-}$, but if all other components are matched and if the LO signal is perfect ($R_1=R_2=R_3=R_4=R$, $R_{on1}=R_{on2}=R_{on3}=R_{on4}=R_{on}$, $\omega_{RF1}=\omega_{RF2}=\omega_{RF}$, $\alpha=1$), then the RF to IF leakage is identical on both of the mixer's differential outputs and is:

$$I_{leakage} = I_{n_{leakage}} = \frac{V_{RF+} - V_{RF-}}{2(R + R_{on})} \cos \omega_{RF} t$$

Thus, in the case of amplitude mismatch at the RF side, the RF to IF leakage is in common mode at the mixer output.

In the same way, in the case of phase mismatch at the RF side, the RF to IF leakage is also in common mode at the mixer output. If $\omega_{RF+} \neq \omega_{RF-}$, if all other components are matched and if LO signal is perfect ($R_1=R_2=R_3=R_4=R$, $R_{on1}=R_{on2}=R_{on3}=R_{on4}=R_{on}$, $V_{RF+}=V_{RF-}=V_{RF}$, $\alpha=1$), then the RF to IF leakage is identical on both mixer outputs and is:

- 10 -

$$I_{leakage} = In_{leakage} = \frac{V_{RF}}{2(R + Ron)} (\cos \omega_{RF+} t - \cos \omega_{RF-} t)$$

Once again, this is a common mode signal, so it can be discounted.

Imbalance of the LO signal can also be considered. If the LO signal has a duty cycle other than 50% then $\alpha \neq 1$. In this case, if all other components are matched and if RF signal is perfect ($R_1=R_2=R_3=R_4=R$, $Ron_1=Ron_2=Ron_3=Ron_4=Ron$, $V_{RF+}=V_{RF-}=V_{RF}$, $\omega_{RF+}=\omega_{RF-}=\omega_{RF}$), then the RF to IF leakage current is:

$$I_{leakage} = \frac{V_{RF} \cos \omega_{RF} t}{2} \left(\frac{1}{R_1 + Ron} - \frac{1}{\alpha(R_2 + Ron)} \right)$$

Thus, in the case of imbalance at the LO side, the RF to IF leakage is cancelled out for the following value of R_1 :

$$R_1 = \alpha R_2 + (\alpha - 1) Ron$$

Calibration circuits according to embodiments of the invention can accommodate mismatch of the components, imbalance of the RF signal, and imbalance of the LO signal, all occurring together. In a real mixer, all of these sources of mismatch will contribute in the mixer RF to IF leakage concurrently. In the case of mismatch of the components and imbalance of the RF and LO signals, the RF to IF leakage is:

$$I_{leakage} = \frac{V_{RF+} \cos \omega_{RF+} t}{2(R_1 + Ron_1)} - \frac{V_{RF-} \cos \omega_{RF-} t}{2\alpha(R_2 + Ron_2)}$$

$$In_{leakage} = \frac{V_{RF+} \cos \omega_{RF+} t}{2\alpha(R_3 + Ron_3)} - \frac{V_{RF-} \cos \omega_{RF-} t}{2(R_4 + Ron_4)}$$

Since the RF signal is not perfectly balanced, the RF to IF leakage cannot be completely eliminated. However, by choosing the following values for R_1 and R_4 it is possible to ensure that the RF to IF leakage appears at the mixer output in common mode:

$$R_1 = \alpha(R_3 + Ron_3) - Ron_1$$

$$R_4 = \alpha(R_2 + Ron_2) - Ron_4$$

Consequently, after calibration the RF to IF leakage will be in common mode at the mixer output. The amplitude of the leakage current will typically be low, so it should not affect the linearity of the following stages significantly. After calibration, with the values of R_1 and R_4 given above, the RF to IF leakage will correspond to the following currents at the mixer output:

$$I_{leakage} = In_{leakage} = \frac{V_{RF+} \cos \omega_{RF+} t}{2\alpha(R_3 + Ron_3)} - \frac{V_{RF-} \cos \omega_{RF-} t}{2\alpha(R_2 + Ron_2)}$$

In this example, the parameter of the mixer that the control circuit is adapted to modify includes a first parameter associated with the positive terminal of the RF input port of the mixer and a second parameter associated with the negative terminal. In particular, the mixer comprises variable impedance elements, whose value is modifiable by the control circuit – at least variable resistor R_1 (positive terminal) and R_4 (negative terminal). In other words, the parameter that the control circuit is adapted to modify comprises the value of each variable resistance.

As shown in Fig. 4, the resistance value of each input resistor of the mixer can be modified by the use of MOS switches. The specified granularity and calibration range will define the size and the number of switches. In the example of Fig. 4, three resistors R_0 , R_{sw0} , and R_{sw1} are connected in parallel. Two of the resistors R_{sw0} and R_{sw1} are connected in series with the drains of respective MOS transistors M_0 and M_1 . Switch signals $sw<0>$ and $sw<1>$, generated by the control circuit 40, are applied to the gates of these transistors, to control the total combined resistance that is presented. The number of switched resistors and their values can be chosen so as to provide the desired steps in resistance, when the respective transistors are turned on. With two switched resistors, as shown in Fig. 4, up to 4 (= 2 x 2) discrete levels of resistance can be provided.

Note that, even if it is not necessary to change the values of the resistors R_2 and R_3 , it is preferable to include the same switches as for R_1 and R_4 . This is desirable for reasons of symmetry. It can also increase the range of the calibration system, because the extra variables provide additional degrees of freedom.

Fig. 5 illustrates a calibration method, suitable for selecting the values of the resistors R_1 and R_4 , according to an embodiment of the invention.

The method comprises first generating 100 a first test signal at the RF input port 12; and measuring 110 a first resulting signal at the output port. The test signal is injected 100 by the test-signal generator. In this embodiment the test signal is a test tone. The test tone may be a sinusoid, although this is not essential. For example, a square wave or triangular wave

may also suffice. Such monotone waveforms contain power at a fundamental frequency and at harmonics which are multiples of that frequency. However, provided these harmonics fall outside the frequency band of interest, they will not have any effect on the calibration process. Therefore, the first test tone can be considered to have one unique frequency, RF1.

5 As will be well known to those skilled in the art, the LO signal may also be a square wave in a passive mixer like the one illustrated. In this case, if the test tone is also a square wave (for example), the harmonics of the two fundamental frequencies will also mix, producing additional sum and difference frequency components at the output. Care should be taken to ensure that these unwanted components fall outside the frequency band of interest, to avoid
10 interfering with the calibration process.

The first test tone will provide the reference power level for the calibration, so its frequency RF1 is chosen to avoid generating RF to IF leakage at the output of the mixer. This is achieved by choosing RF1 to be a frequency where the RF (input) and IF (output) spectra do not overlap.

15 The input at frequency RF1 is down-mixed by the mixer and appears in the output as an intermediate frequency (IF) signal with a frequency that is the equal to the difference between RF2 and the LO frequency. This output power, IF0, resulting from the test tone RF1 is measured 110 by the power detector 30 and this measurement is provided to the control circuit 40, where it is stored.

20 The method proceeds by generating 120 a second test signal at the RF input port 12; and measuring 130 a second corresponding signal at the output port 14. The second test signal is a test tone at a second frequency RF2 that is different to the first. RF2 is chosen so that it generates RF-to-IF leakage. That is, a power component at frequency RF2 appears at the output. More particularly, RF2 is chosen to be equal to the intermediate frequency resulting
25 from RF1. So, the second test tone RF2 is at the same frequency as IF0.

Apart from the leakage component at RF2, which feeds directly through to the mixer output, the mixer will also generate product components corresponding to the mixing of RF2 with the LO signal. However, these components will be out-of-band and will be rejected by the filter 50. For example, one product component will appear at RF1 (being the sum of the LO
30 frequency and RF2); however, as already noted above, RF1 is at a frequency which does not overlap with the desired IF spectrum, so it is eliminated by the filtering. This means that the power at the output resulting from the second test tone RF2 is due solely to leakage. This leakage output power IF1 is measured 130 by the power detector 30 and stored by the digital state machine 40.

35 Next, in step 140a, the leakage output power IF1, generated by the second test tone RF2, is subtracted from the reference output power IF0, generated by the first test tone RF1.

The result is compared with a threshold, T , to determine if the leakage is rejected sufficiently. In this example, the threshold leakage rejection, T , is 60 dBc. That is, $(IF0-IF1)$ should be greater than 60 dBc.

If this condition is met, then further calibration is unnecessary, and the procedure stops.

- 5 If the condition is not met, the method proceeds by modifying at least one parameter of the mixer circuit, so as to reduce the leakage from the RF input port to the output port. In this example, the variable parameter of the mixer circuit is the value of an input resistance $R1$ or $R4$.

- 10 More particularly, the modification comprises a heuristic algorithm in which the resistance is adjusted and it is assessed whether the leakage has been reduced. Note that the second test tone $RF2$ continues to be injected, so that the leakage power can be measured as the adjustments are made.

- In this example, the value of the resistance $R1$ is optimised first, in step 200a. The optimisation algorithm will be described in greater detail below, with reference to Fig. 6. There are two possible outcomes of this optimisation: depending on whether the value of $R1$ has been changed or not. If $R1$ has changed, the method re-evaluates the leakage rejection against the threshold (step 140b). If the leakage rejection is good enough, the calibration is complete –in other words, modification of $R1$ was sufficient to calibrate the mixer circuit. If the leakage rejection is still not good enough, the value of $R4$ is optimised, in step 200b. Note that if $R1$ is unchanged by the optimisation, there is no need to re-evaluate the leakage rejection, so the method instead proceeds directly to step 200b. Once again, there are two possible outcomes from the optimisation: $R4$ changed or unchanged. If $R4$ has been changed, the method loops back to step 140a, in which the leakage rejection is compared with the threshold, T . If $R4$ is unchanged, then it is assumed that no further improvement can be achieved in the leakage rejection, and the calibration method ends.

The optimisation algorithm used in the present embodiment will now be described, with reference to Fig. 6. Note that the same optimisation algorithm is used for optimising $R1$ and $R4$.

- The resistance value, R , of the resistor being optimised is firstly incremented 210. (The size of the increment will be determined by the values of the switched resistors in the digital potentiometer, as illustrated in the example of Fig. 4). The leakage power is measured 220 and stored as $IF2$. The state machine 40 determines 230 whether the leakage rejection has improved – that is, if $(IF0-IF2) > (IF0-IF1)$. If it has not, the resistance value R is decremented 250 by twice amount of the previous, positive increment. This makes R less than its starting value, by the size of one step. Once again, the leakage power is measured 260 and stored as $IF2$; and once again a check is performed 270, to determine if the leakage rejection has

improved. If either of the tests 230, 270 gives a positive result (namely, that the leakage rejection has improved), the method proceeds by replacing 240 the stored value IF1 of the originally measured leakage power with the newly measured power IF2. On the other hand, if neither the incremented value nor the decremented value of R has resulted in improved leakage rejection, the value of R is returned to its original value, by incrementing it again 280 by one step.

As shown in Fig. 5, this procedure will be iterated, alternately adjusting R1 and R4, until the leakage rejection achieves a value above the threshold, or until the optimisation algorithm fails to find a suitable adjustment of the resistance values that reduces leakage.

Fig. 7 shows a mixer circuit according to an embodiment of the invention in a generalised form. The mixer circuit comprises a switching (or multiplying) element 300 at its heart, which performs the actual mixing of the two input signals. A local oscillator voltage V_{LO} is applied 320 at the local oscillator input port 13. At the signal input port 12 there is a voltage-to-current converter 300. This converts the input voltage applied at the input port 12 to an input current signal. The current produced by the voltage-to-current converter 310 is applied to the switching element 300, where it is modulated by (that is, mixed with) the local oscillator voltage signal 320. The result of this mixing is a current, which is input to a current-to-voltage converter 330. This converts the mixed current signal into an output voltage signal at the output port 14.

Fig. 8 illustrates the equivalence between the generalised circuit of Fig. 7 and the embodiment of Figs. 2 to 3. The voltage-to-current converter 310a corresponds to the variable input resistances R1-R4; the switching element 300a corresponds to the transistors M1-M4; and the current-to-voltage converter 320a corresponds to the output stage, including load resistances R5-R6. Note that the value of R_{on} in Fig. 3 is assumed to be included in the resistance value of the voltage-to-current converter 310a in Fig. 8.

Fig. 9 illustrates the equivalence between the generalised embodiment of Fig. 7 and a second exemplary embodiment, in which the mixer is an active mixer. Here, the voltage-to-current converter 310b is a variable transconductance transistor element; the switching element 300b comprises a pair of transistors and the current-to-voltage converter 320b at the output port comprises two load resistors.

Fig. 10 shows the circuit design of a variable transconductance 310b for the embodiment of Fig. 9. The input signal V_{RF} is fed to the gates of transconductors M1 and M2 via coupling capacitors C. In parallel, resistors R couple the gate of each transconductor M1, M2 to a biasing circuit. The resistors R are included to avoid modulation of the current source.

Through each resistor R, the gate of each transconductor M1 and M2 is biased either to a respective current mirror or to ground, depending upon the configuration of switches A1 and A2. Thus, by changing the state of A1 and A2 the overall transconductance is made

variable as indicated in the table and formulas below. In this example, the overall transconductance can be gm_1 , gm_2 , or gm_1+gm_2 :

$$gm_1 \approx \sqrt{2\mu C_{ox} \frac{W_1}{L_1} I_{bias}}$$

$$gm_2 \approx \sqrt{2\mu C_{ox} \frac{W_2}{L_2} I_{bias}}$$

5

Table 1: effect of switch-states on transconductance in Fig. 10

A1	A2	gm
0	0	0
0	1	gm_2
1	0	gm_1
1	1	gm_1+gm_2

Other transconductors with different W/L size ratios could be connected in parallel to increase the range and granularity of the overall transconductance, if desired. In this example, the ratio of the current mirrors is still 1:1 to keep the DC current constant when one transconductor is ON at a time. This may be desirable in order to maintain constant power consumption in the mixer. The variable transconductor of Fig. 10 can be used with embodiments of the present invention to compensate for RF-to-IF leakage in active mixers.

Fig. 11 shows an active mixer based on that of Fig. 9. The only difference is that the circuit of Fig. 11 has a second branch, comprising two additional switching transistors and one additional variable-transconductance element. This arrangement provides for a differential input signal V_{RF} . Thus, Fig. 11 is related to Fig. 9 in the same way that the differential-input mixer of Fig. 2 is related to Fig. 8. Each of the two variable transconductance elements of Fig. 11 is implemented by a circuit like that of Fig. 10.

The balance-condition for cancelling input-to-output leakage in the active mixer of Fig. 11 can be derived theoretically, following the same principles as the exemplary derivation presented earlier above, for the passive mixer of Fig. 2. In practice, the leakage calibration will typically be performed by the same iterative optimization methods as those already described above.

As will be apparent to those skilled in the art, the calibration algorithm could be implemented in custom hardware, or in software running on a generic programmable microprocessor or microcontroller. Methods of programming such a device will be well known to those skilled in the art.

The invention offers improved rejection of RF-to-IF leakage at a differential or complex down-conversion mixer in a receiver. It can compensate not only for mismatch in the mixer circuit components, but also for variations in the LO duty cycle. The invention can be used to particular advantage in a cable modem or TV receiver for DOCSIS 3.0 applications.

5 Note that input-to-output leakage cannot be compensated for at the output of the mixer (for example, by adjusting the impedance of a variable load element) because if leakage is appearing in differential mode at the output port, it is too late to eliminate it. The following analysis of the embodiment of Fig. 2 demonstrates why this is so.

10 Referring to Fig. 2, and considering the case of mismatch of R_{on} for the switching elements, the RF-to-IF leakage currents are:

$$I_{leakage} = \frac{V_{RF} \cos \omega_{RF} t}{2} \left(\frac{1}{R_1 + R_{on1}} - \frac{1}{R_2 + R_{on2}} \right) = \frac{V_{RF} \cos \omega_{RF} t}{2} \frac{(R_2 + R_{on2}) - (R_1 + R_{on1})}{(R_1 + R_{on1})(R_2 + R_{on2})}$$

$$I_{nleakage} = \frac{V_{RF} \cos \omega_{RF} t}{2} \left(\frac{1}{R_3 + R_{on3}} - \frac{1}{R_4 + R_{on4}} \right) = \frac{V_{RF} \cos \omega_{RF} t}{2} \frac{(R_4 + R_{on4}) - (R_3 + R_{on3})}{(R_3 + R_{on3})(R_4 + R_{on4})}$$

If the IF (load) resistors R_5 - R_6 were variable, the leakage would be in common-mode if and only if:

15

$$R_5 \frac{(R_2 + R_{on2}) - (R_1 + R_{on1})}{(R_1 + R_{on1})(R_2 + R_{on2})} = R_6 \frac{(R_4 + R_{on4}) - (R_3 + R_{on3})}{(R_3 + R_{on3})(R_4 + R_{on4})}$$

$$R_5 = R_6 \frac{(R_4 + R_{on4}) - (R_3 + R_{on3})}{(R_2 + R_{on2}) - (R_1 + R_{on1})} \frac{(R_1 + R_{on1})(R_2 + R_{on2})}{(R_3 + R_{on3})(R_4 + R_{on4})}$$

However, this equation shows clearly that the required value for R_5 may be negative, implying a physically unrealizable resistor.

20 For an active implementation the same result is obtained by replacing $R_1 + R_{on1}$ by $1/gm_1$, and so on. This demonstrates that it is not possible to calibrate the RF-to-IF leakage in IF domain.

Next, consider LO signal imbalance. The leakage currents are:

$$I_{leakage} = \frac{V_{RF} \cos \omega_{RF} t}{2} \left(\frac{1}{R} - \frac{1}{\alpha R} \right) = \frac{V_{RF} \cos \omega_{RF} t}{2} \left(\frac{\alpha - 1}{\alpha R} \right)$$

$$I_{nleakage} = \frac{V_{RF} \cos \omega_{RF} t}{2} \left(\frac{1}{\alpha R} - \frac{1}{R} \right) = \frac{V_{RF} \cos \omega_{RF} t}{2} \left(\frac{1 - \alpha}{\alpha R} \right) = -I_{leakage}$$

25 It can be seen that in this case the RF-to-IF leakage signal is perfectly differential and is superimposed with the wanted signal (also in differential mode). It would be meaningless to attempt RF-to-IF leakage-cancellation in the IF domain in this case, because the desired, mixed signal (in differential mode) would also be cancelled.

If the IF load resistors R5 and R6 were variable it can be shown that the RF-to-IF leakage would be in common mode at the mixer output if and only if $R_5 = -R_6$, implying a physically unrealizable resistor. This demonstrates again that it is not possible to calibrate the RF-to-IF leakage in IF domain.

- 5 In the general case, if all sources of mismatch are considered, the RF-to-IF leakage currents are:

$$I_{leakage} = \frac{V_{RF+} \cos \omega_{RF+} t}{2(R_1 + Ron_1)} - \frac{V_{RF-} \cos \omega_{RF-} t}{2\alpha(R_2 + Ron_2)}$$

$$In_{leakage} = \frac{V_{RF+} \cos \omega_{RF+} t}{2\alpha(R_3 + Ron_3)} - \frac{V_{RF-} \cos \omega_{RF-} t}{2(R_4 + Ron_4)}$$

- 10 The RF-to-IF leakage can be put in common mode if these currents are equal. Thus, the conditions are:

$$R_1 = \alpha(R_3 + Ron_3) - Ron_1$$

$$R_4 = \alpha(R_2 + Ron_2) - Ron_4$$

$$\alpha = \frac{R_1 + Ron_1}{R_3 + Ron_3}$$

$$\alpha = \frac{R_4 + Ron_4}{R_2 + Ron_2}$$

- 15 As α cannot have 2 different values, this demonstrates that it is not possible to cancel the RF-to-IF leakage by varying the LO signal imbalance.

It should be noted that the above-mentioned embodiments illustrate rather than limit the invention, and that those skilled in the art will be able to design many alternative embodiments without departing from the scope of the appended claims.

- 20 For example, in the embodiment of Fig. 1, the filter 50 is shown as being part of the normal receiver signal-path. This is typically the case in real tuner circuits, because the output of the mixer is normally required to be narrowband, with out-of-band interference and inter-modulation products eliminated. However, if the mixer circuit does not include a filter 50, the filtering could instead be provided in the calibration loop, just before the input to the measurement circuit 30.

- 25 The examples described above have concentrated on the case that the mixer is a down-conversion mixer, at a receiver. Here, the input-output leakage is RF-to-IF leakage. However, circuits and methods according to other embodiments of the invention can be used to compensate for IF-to-RF leakage, in an up-conversion mixer. Note that, in this case, the detection loop should include a band-pass filter to select only the leakage to be measured.
- 30

In other embodiments, other parameters of the mixer circuit may be modified. Such parameters may include a gain of a part of the mixer circuit, a current flowing through a part of the mixer circuit, or a voltage applied to a part of the mixer circuit.

Whether the parameters modified comprise impedance or any other variable, various search strategies can be used to find an optimum. For example, one simple alternative to the exemplary strategy described above would be to exhaustively test all possible combinations of the values of R1 and R4. This will ensure that an absolute optimum is found; although it is also likely to consume more time and energy than the simple heuristic described in detail above.

The invention is equally applicable to complex mixers.

References in the specification or claims to an intermediate frequency are taken to include the possibility of a low or zero intermediate frequency, as in the case of a direct conversion receiver.

In the claims, any reference signs placed between parentheses shall not be construed as limiting the claim. The word "comprising" does not exclude the presence of elements or steps other than those listed in a claim. The word "a" or "an" preceding an element does not exclude the presence of a plurality of such elements. The invention can be implemented by means of hardware comprising several distinct elements. In the device claim enumerating several means, several of these means can be embodied by one and the same item of hardware. The mere fact that certain measures are recited in mutually different dependent claims does not indicate that a combination of these measures cannot be used to advantage.

CLAIMS

1. A circuit comprising:

a mixer (10) having a signal input port (12) for a signal to be up-converted or down-converted; a local oscillator input port (13); and an output port (14), for a transmitter or receiver in which the spectrum of the signal at the signal input port (12) and the spectrum of a mixed signal at the output port (14) overlap at least partially,

wherein at least one of the signal input port (12) and the local oscillator input port (13) is a differential input port and the output port (14) is a differential output port,

the mixer comprising a voltage-to-current conversion part (310) at the input port, having at least one modifiable voltage-to-current conversion parameter; and

a calibration circuit comprising:

a test signal generator (20), operable to inject a first test signal at the signal input port (12);

a measurement circuit (30), adapted to measure a first output signal at the output port (14) resulting from the injected first test signal; and

a control circuit (40), adapted to modify the at least one voltage-to-current conversion parameter of the voltage-to-current conversion part (310) based on the measurement, so as to reduce leakage from the signal input port (12) to the output port (14).

2. The circuit of claim 1, wherein:

the test signal generator (20) is further operable to inject a second test signal at the signal input port (12);

the measurement circuit (30) is further adapted to measure a second output signal at the output port (14) resulting from the injected second test signal; and

the control circuit (40) is adapted to

compare the measurement of the first output signal with the measurement of the second output signal, and to

modify the at least one voltage-to-current conversion parameter based on the outcome of the comparison,

wherein a frequency of the second test signal differs from a frequency of the first test signal.

3. The circuit of claim 1 or claim 2, wherein

the signal input port (12) is a differential input port comprising a positive terminal and a negative terminal; and

the at least one voltage-to-current conversion parameter that the control circuit is adapted to modify includes a first parameter associated with the positive terminal and a second parameter associated with the negative terminal.

4. The circuit of any of claims 1 to 3, wherein the voltage-to-current conversion part (310) comprises at least one variable impedance element at the input port, whose value is modifiable by the control circuit,

wherein the at least one voltage-to-current conversion parameter that the control circuit is adapted to modify comprises the value of the at least one variable impedance.

5. The circuit of any of claims 1 to 3, wherein the mixer comprises at least one variable transconductance element at the input port, whose value is modifiable by the control circuit,

wherein the at least one voltage-to-current conversion parameter that the control circuit is adapted to modify comprises the value of the at least one variable transconductance.

6. A method of calibrating a mixer circuit (10) having a signal input port (12) for a signal to be up-converted or down-converted; a local oscillator input port (13); and an output port (14), for a transmitter or receiver in which the spectrum of the signal at the signal input port (12) and the spectrum of a mixed signal at the output port (14) overlap at least partially,

wherein at least one of the signal input port (12) and the local oscillator input port (13) is a differential input port and the output port (14) is a differential output port,

the mixer circuit (10) comprising a voltage-to-current conversion part (310) at the input port, having at least one modifiable voltage-to-current conversion parameter,

the method comprising:

generating (100) a first test signal at the signal input port (12);

measuring (110) a first resulting signal at the output port (14); and

modifying (200) the at least one voltage-to-current conversion parameter of the voltage-to-current conversion part (310) based on the measurement, so as to reduce leakage from the signal input port to the output port.

7. The method of claim 6, further comprising:

generating (120) a second test signal at the signal input port (12);

measuring (130) a second resulting signal at the output port (14);

comparing (140) the first and second resulting signals; and

modifying (200) the at least one voltage-to-current conversion parameter based on the result of the comparison,

wherein a frequency of the second test signal differs from a frequency of the first test signal.

8. The method of claim 7, further comprising applying a local oscillator signal at the local oscillator input port (13) when applying the first and second test signals,

wherein the second test signal has a frequency equal to the intermediate frequency resulting from the mixing of the first test signal with the local oscillator signal.

9. The method of any of claims 6 to 8, wherein the step of modifying the at least one voltage-to-current conversion parameter comprises adjusting the parameter and assessing whether the leakage from the signal input port to the output port is reduced.

10. The method of any of claims 6 to 9, wherein

the signal input port (12) is a differential input port comprising a positive terminal and a negative terminal; and

the at least one voltage-to-current conversion parameter includes a first parameter associated with the positive terminal and a second parameter associated with the negative terminal.

11. The method of claim 10, wherein the step of modifying the at least one parameter comprises alternately adjusting the first parameter and the second parameter.

12. The method of any of claims 6 to 11, wherein the step of modifying the at least one parameter is performed iteratively.

13. The method of any of claims 6 to 12, wherein the at least one parameter that is modified comprises the value of an impedance.

14. The method of any of claims 6 to 12, wherein the at least one parameter that is modified comprises the value of a transconductance.

15. A computer program comprising computer program code means adapted to control a calibration circuit to perform all the steps of the method of any of claims 6 to 14 when said program is run on a computer.

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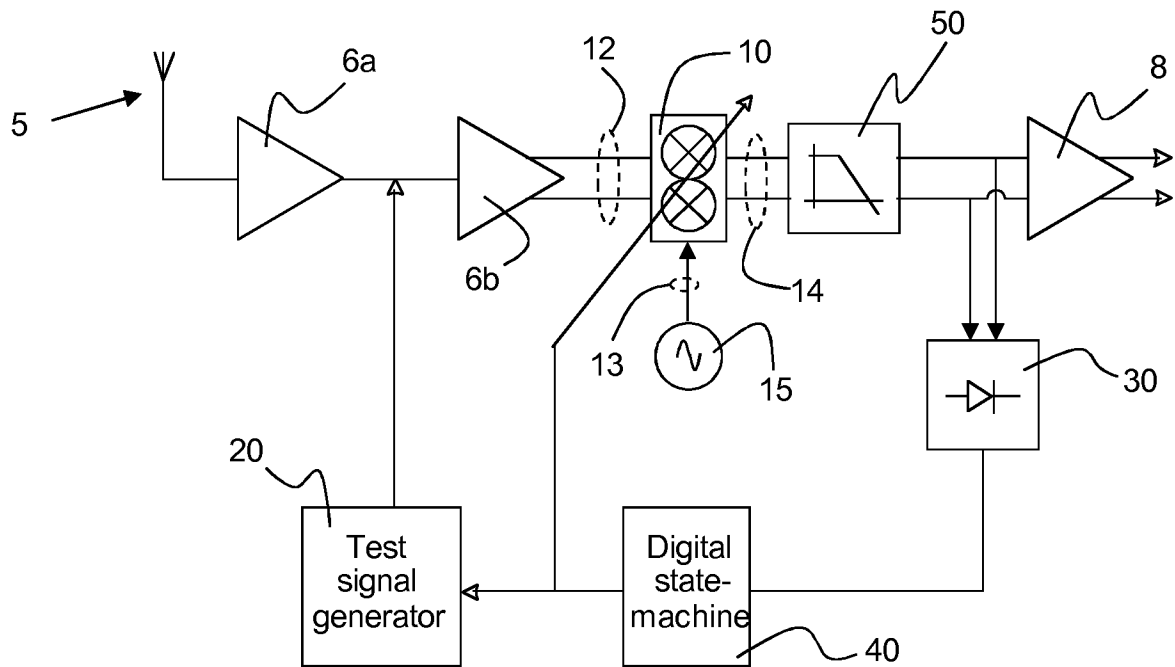


FIG 1

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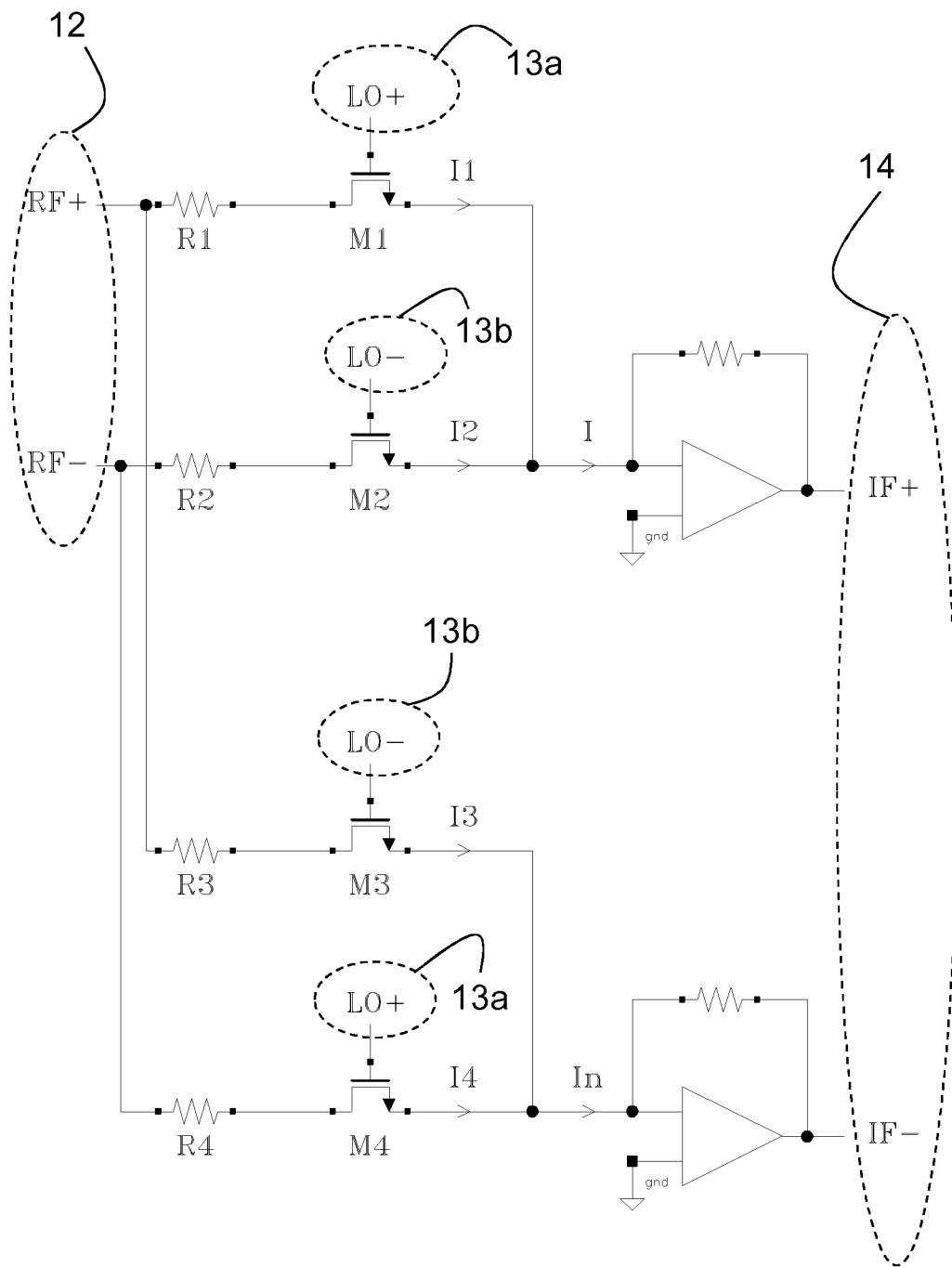


FIG 2

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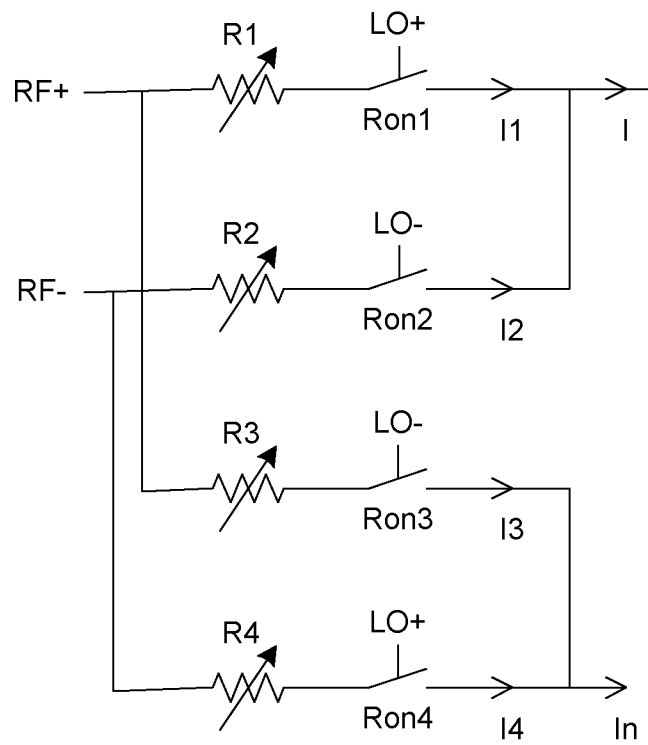


FIG 3

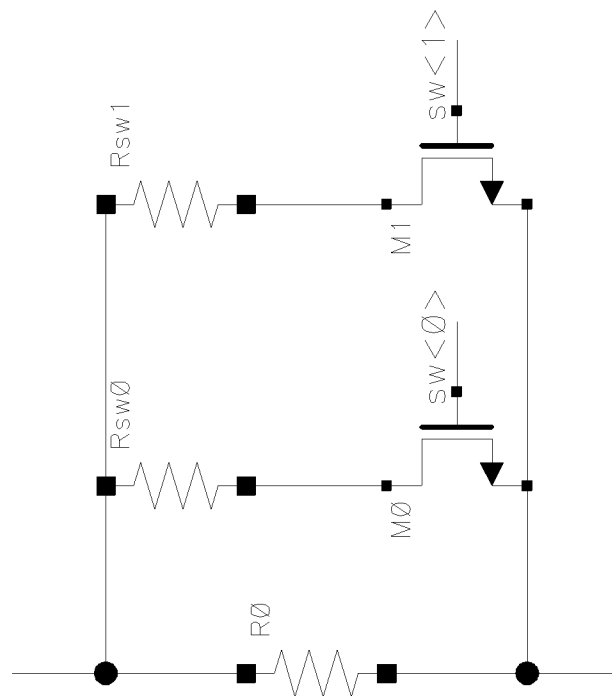


FIG 4

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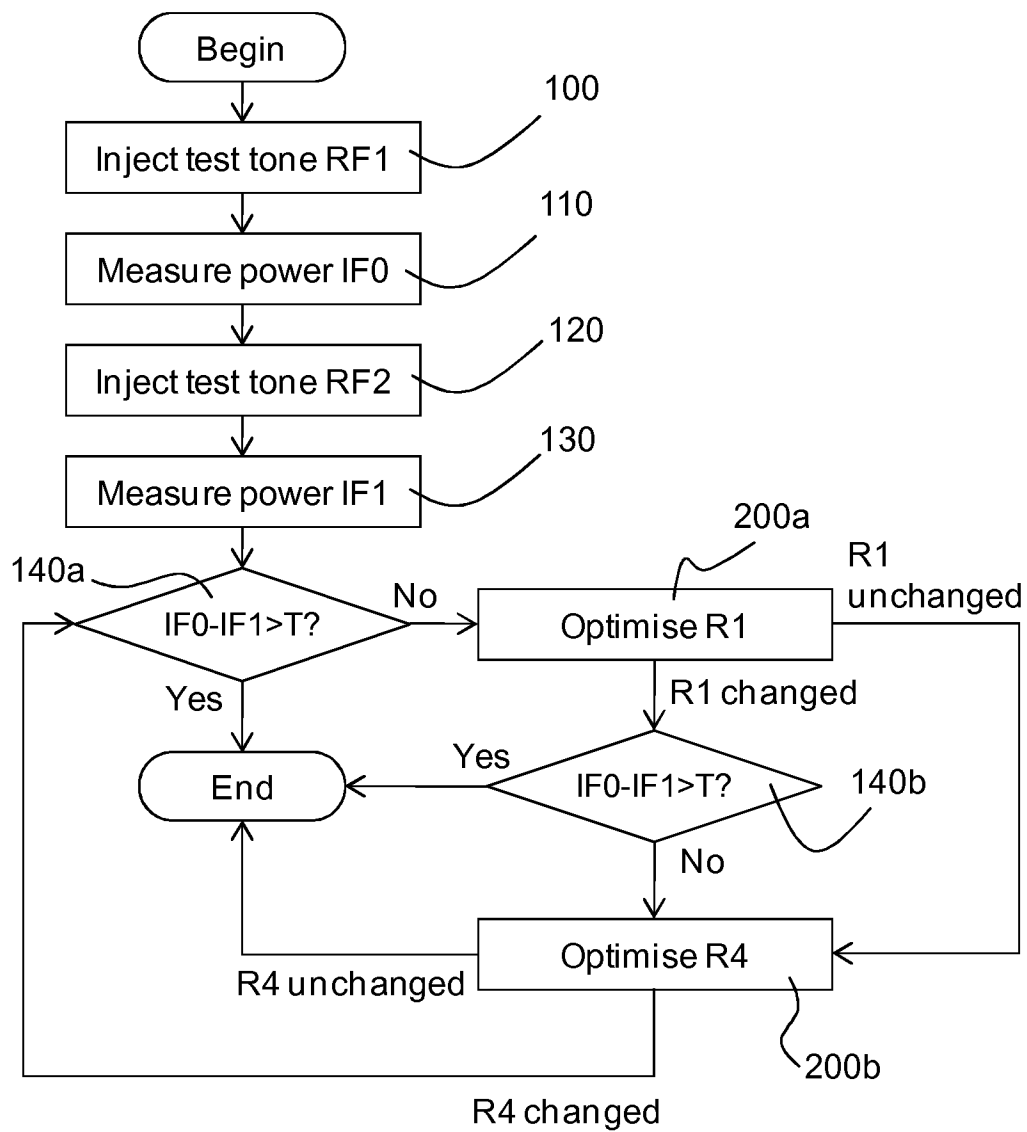


FIG 5

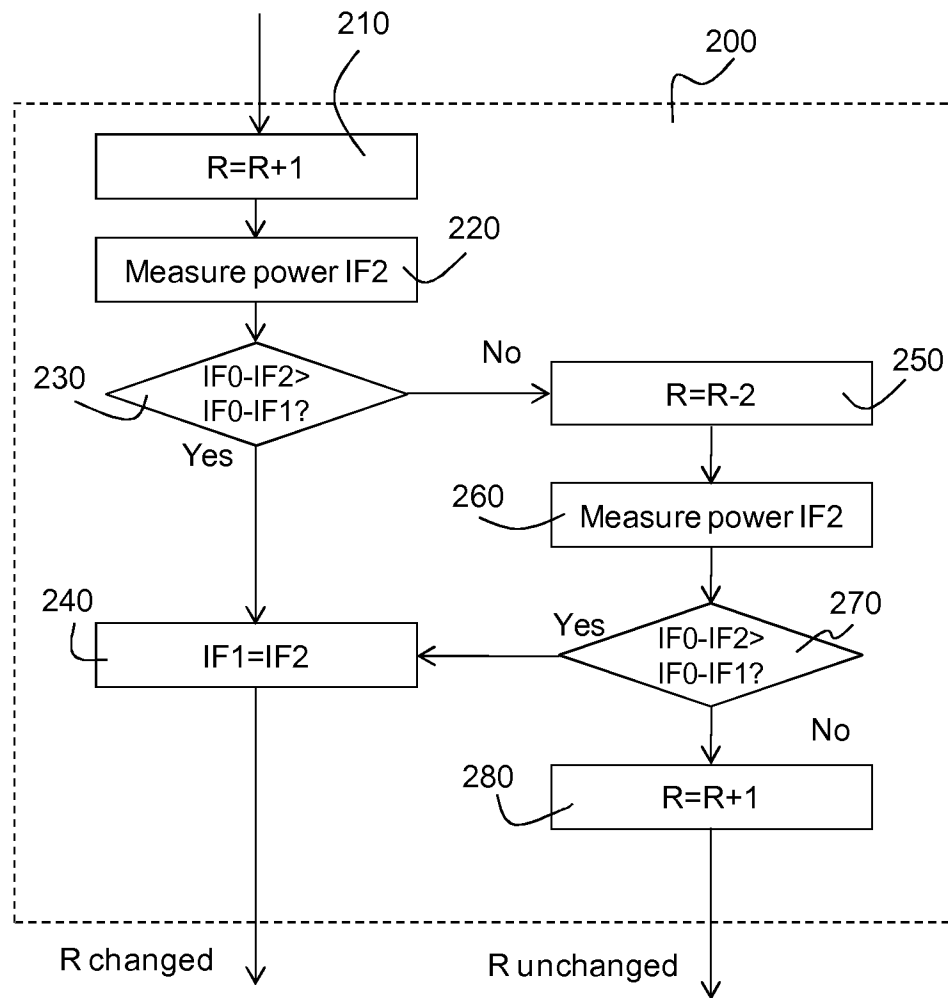


FIG 6

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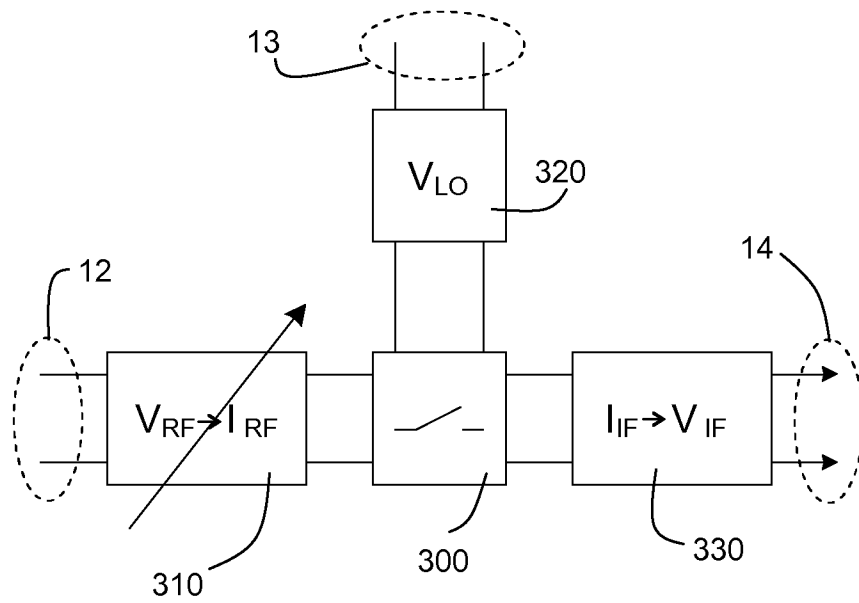


FIG 7

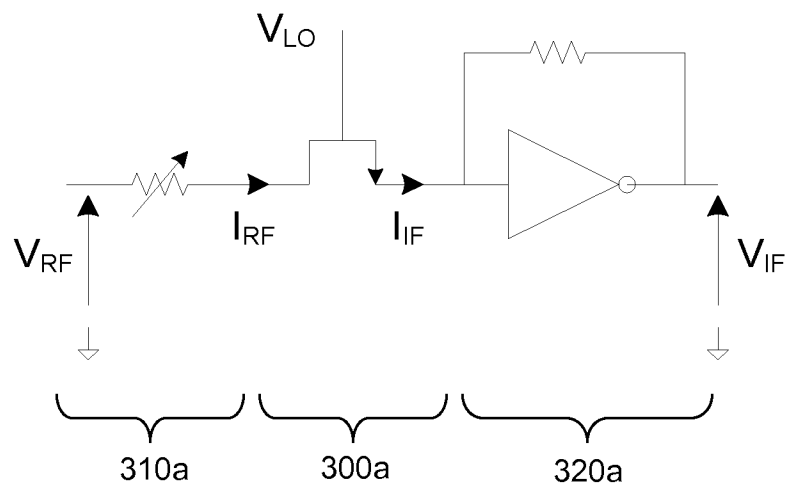


FIG 8

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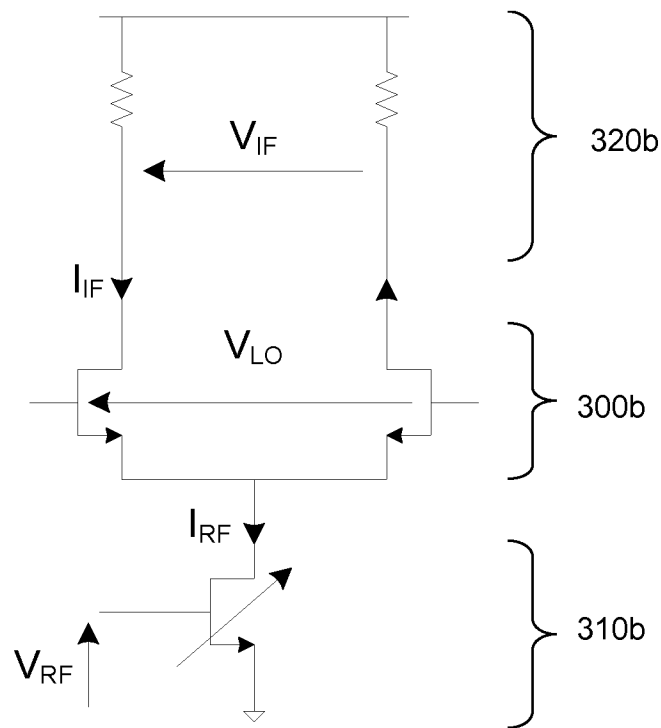


FIG 9

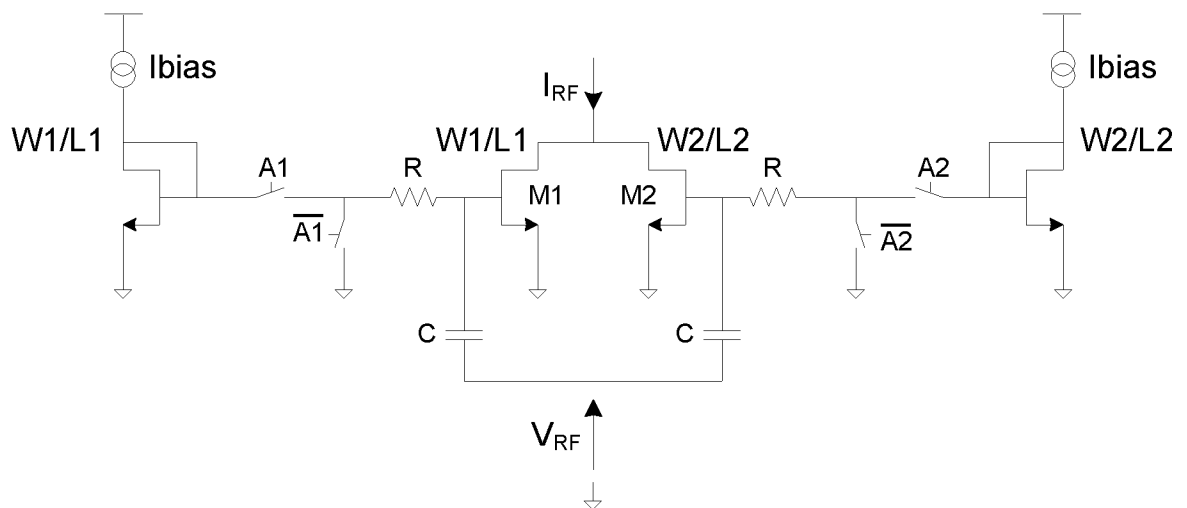


FIG 10

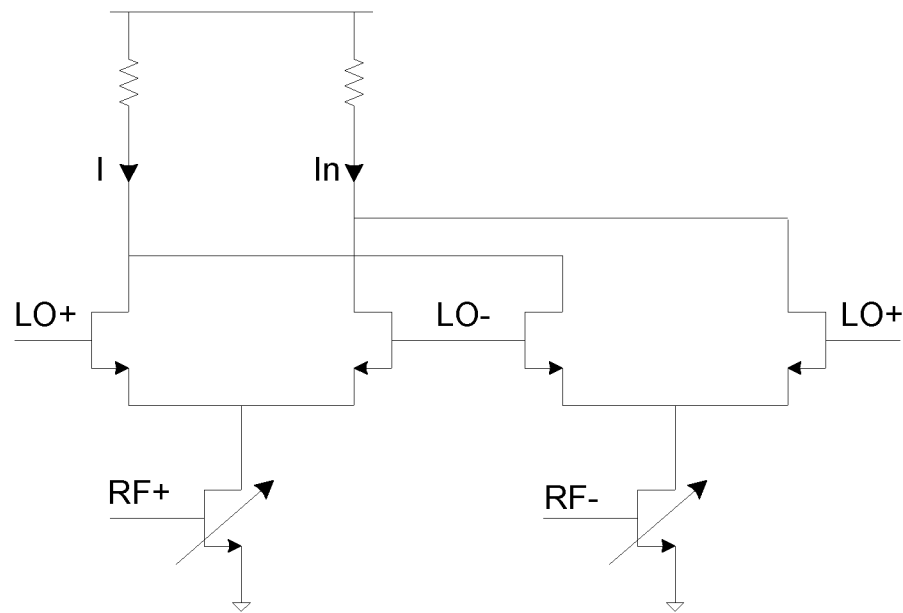


FIG 11

INTERNATIONAL SEARCH REPORT

International application No
PCT/EP2011/055101

A. CLASSIFICATION OF SUBJECT MATTER
INV. H03D7/14
ADD.

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)
H03D

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 practical, search terms used)

EPO-Internal

C. DOCUMENTS CONSIDERED TO BE RELEVANT

Category*	Citation of document, with indication, where appropriate, of the relevant passages	Relevant to claim No.
X	US 2006/128344 A1 (KIVEKAS KALLE [FI] ET AL KIVEKAES KALLE [FI] ET AL) 15 June 2006 (2006-06-15) paragraph [0035]; figures 3,5,6 -----	1,3-6, 9-15
X	RODRIGUEZ S ET AL: "A Novel BiST and Calibration Technique for CMOS Down-Converters", CIRCUITS AND SYSTEMS FOR COMMUNICATIONS, 2008. ICCSC 2008. 4TH IEEE INTERNATIONAL CONFERENCE ON, IEEE, PISCATAWAY, NJ, USA, 26 May 2008 (2008-05-26), pages 828-832, XP031268815, ISBN: 978-1-4244-1707-0 figures 1,5,7 ----- -/-	1,3-6, 10,11, 13-15

☒ Further documents are listed in the continuation of Box C.

☒ See patent family annex.

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Date of the actual completion of the international search

26 August 2011

Date of mailing of the international search report

05/09/2011

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Authorized officer

Schnabel, Florian

INTERNATIONAL SEARCH REPORT

International application No
PCT/EP2011/055101

C(Continuation). DOCUMENTS CONSIDERED TO BE RELEVANT		
Category*	Citation of document, with indication, where appropriate, of the relevant passages	Relevant to claim No.
X	WO 03/079573 A1 (NOKIA CORPRATION [FI]; NOKIA INC [US]; SEPPINEN PAULI [FI]; PARSSINEN) 25 September 2003 (2003-09-25) page 20, line 4 - line 15; figures 19,20 -----	1-8,10, 11,13-15
X	FILLATRE V ET AL: "A SiP Tuner with Integrated LC Tracking Filter for both Cable and Terrestrial TV Reception", 2007 IEEE INTERNATIONAL SOLID-STATE CIRCUITS CONFERENCE (IEEE CAT. NO.07CH37858) IEEE PISCATAWAY, NJ, USA,, 1 February 2007 (2007-02-01), pages 208-597, XP031180049, ISBN: 978-1-4244-0852-8 page 208, column 2, paragraph 2; figures 1,4,5 -----	1-8,10, 11,13-15
X	US 6 393 260 B1 (MURTOJAERVI SIMO [FI] ET AL) 21 May 2002 (2002-05-21) figures 2,4 -----	1,3-5, 10-15
X,P	US 2010/203860 A1 (TILLMAN FREDRIK [SE] ET AL) 12 August 2010 (2010-08-12) paragraph [0029]; figure 3 -----	1
A	US 2005/159124 A1 (SHAH PETER J [US] SHAH PETER JIVAN [US]) 21 July 2005 (2005-07-21) paragraph [0067] - paragraph [0075]; figure 4 -----	1-15
A	EP 0 557 800 A1 (HUGHES AIRCRAFT CO [US]) 1 September 1993 (1993-09-01) column 6, line 30 - column 14, line 50; figure 3 -----	1-15
A	US 6 711 396 B1 (BERGSMA ADRIAN J [CA] ET AL) 23 March 2004 (2004-03-23) column 12, line 50 - column 13, line 14; figure 12 -----	1-15
A	US 5 584 066 A (OKANOBU TAIWA [JP]) 10 December 1996 (1996-12-10) figure 5 -----	1-15

INTERNATIONAL SEARCH REPORT

Information on patent family members

International application No

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