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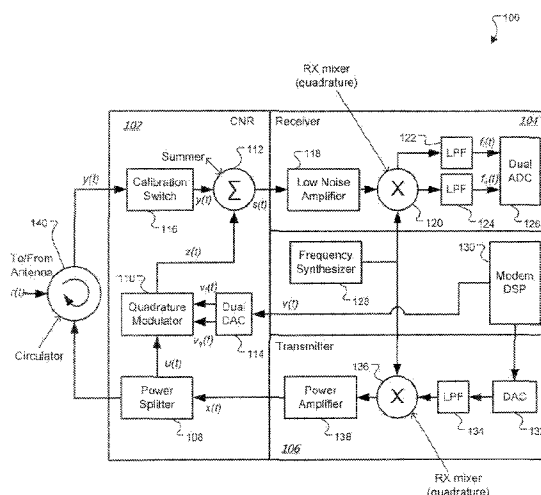
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(54) Title: ADJUSTING PARAMETERS ASSOCIATED WITH LEAKAGE SIGNALS



(57) Abstract: The present disclosure includes a system and method for adjusting parameters associated with leakage signals. In some implementations, an RFID reader includes an RF antenna, a transmitter section, a receiver section, a control module and a cancellation noise reduction (CNR) section. The transmitter section is coupled to the RF antenna and operable to generate a transmit signal to be transmitted by the RF antenna. The receiver section is coupled to the RF antenna and operable to receive a receive signal from the RF antenna. In addition, the receiver section further includes a de-rotation module and a control module. The de-rotation module is operable to de-rotate, by θ , an in-phase signal and quadrature signal associated with the leakage signal. The control module is operable to generate control signals used to produce a signal for reducing the leakage signal in a receive path of the reader. The CNR section is operable to subtract from the reduction signal from the leakage signal.

Adjusting Parameters Associated with Leakage Signals

CROSS-REFERENCE TO RELATED APPLICATION

This application claims benefit of U.S. Provisional Patent Application Serial No. 60/795,625, filed on April 27, 2006, which is hereby incorporated by reference.

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TECHNICAL FIELD

This invention relates to Radio Frequency Identification (RFID) Readers and, more particularly, to adjusting parameters associated with leakage signals.

BACKGROUND

Passive UHF RFID (radio frequency identification) protocols require the tag to
10 be powered by the reader's field and to use the field to backscatter information on the same frequency. The technical term for such a system, where both the transmit and receive sections of the device are simultaneously operating on the same frequency is "homodyne." One class of homodyne systems intends to only transmit a pure continuous sinusoidal wave (CW) signal while in the receive mode. UHF RFID reader
15 systems are of this class. A challenge is presented to the homodyne systems when the receiver section is not well isolated from the transmitter section. Transmitter (TX) leakage into the receive (RX) path can be as much as 110 dB above the desired backscattered receive signal. Such a high TX leakage to receive signal ratio leaves the receiver section quite susceptible to typical nonlinearities associated with standard cost
20 effective analog signal processing components. Therefore an unusually high dynamic range in the receiver section would be required.

Passive and semi-active (battery assisted) UHF RFID communications use radar cross section (RCS) modulation to send data from the transponder to the reader. That means the reader transmits a sinusoidal RF signal toward the transponder. Some of the
25 RF energy which hits the transponder reflects back to the reader. By modulating its RCS, the transponder is able to communicate data back to the reader.

This presents many design challenges. In particular, the reader electronics must be designed to receive a very weak signal while it is transmitting a very strong signal at the same frequency. Whereas many other wireless communications schemes use
30 frequency division multiplexing, the RFID reader cannot since its own transmit field is

being used as a medium for communications from transponder to reader. The transmit signal may be 1 watt or more, while the receive signal for semi-active transponders (those which only use the RF signal for communications, not for power) may be as low as 1 picowatt (10^{-12} watt), e.g., 12 orders of magnitude less power. For passive transponders the receive strength is usually at least 1 nanowatt (10^{-9} watt), which is still pretty challenging.

SUMMARY

The present disclosure includes a system and method for adjusting parameters associated with leakage signals. In some implementations, an RFID reader includes an RF antenna, a transmitter section, a receiver section, a control module and a cancellation noise reduction (CNR) section. The transmitter section is coupled to the RF antenna and operable to generate a transmit signal to be transmitted by the RF antenna. The receiver section is coupled to the RF antenna and operable to receive a receive signal from the RF antenna. In addition, the receiver section further includes a de-rotation module and a control module. The de-rotation module is operable to de-rotate, by θ , an in-phase signal and quadrature signal associated with the leakage signal. The control module is operable to generate control signals used to produce a signal for reducing the leakage signal in a receive path of the reader. The CNR section is operable to subtract the reduction signal from the leakage signal.

The details of one or more embodiments of the invention are set forth in the accompanying drawings and the description below. Other features, objects, and advantages of the invention will be apparent from the description and drawings, and from the claims.

DESCRIPTION OF DRAWINGS

FIGURE 1 is a block diagram illustrating an example RFID reader in accordance with some implementations of the present disclosure;

FIGURE 2 is a block diagram illustrating an example mathematical model of the reader in FIGURE 1;

FIGURE 3 is a flow chart illustrating an example method for estimating DC offsets for baseband signals;

FIGURE 4 is a flow chart illustrating an example method for estimating phase offsets for baseband signals; and

FIGURE 5 is a flow chart illustrating an example method for reducing leakage signal in a receive path.

5 Like reference symbols in the various drawings indicate like elements.

DETAILED DESCRIPTION

FIGURE 1 is an example RFID reader 100 for reducing leakage signal in a receive path in accordance with some implementations in the present invention. For example, the reader 100 may reduce a DC offset and/or phase offset associated with error signals used to compensate for leakage signal. In general, a leakage signal is interference generated from a transmit signal that is added to a receive path. Transmitter leakage into the receive path can be as much as 110 dB above the desired backscattered receive signal. Such a high leakage signal to receive signal ratio can leave the baseband signals susceptible to typical nonlinearities associated with standard cost effective analog signal processing components. In the case that a reader has perfect transmitter-to-receiver isolation, only the reflected signal from the transponder would make it into the receiver. Leakage associated with the transmit signal frequently generates interference in the receive signal and may result from one or more sources such as reflections off other nearby objects in the vicinity, internal circuit reflections caused by non-ideal impedance matching, and/or other sources. In some implementations, the reader 100 offers approximately 40dB (4 orders of magnitude) of isolation. In eliminating, minimizing or otherwise reducing the leakage signal, the reader 100 may generate control signals that are quadrature modulated with a portion of the transmission signal to generate a cancellation signal, *i.e.*, a signal that when added to a receive path can reduce leakage signals. In the process of generating the control signals, the reader 100 may generate DC offsets and/or phase offsets that interfere with the estimated control signals. In some implementations, the DC offsets and the phase offsets may be referred to as nuisance parameters. The DC offsets can include offsets of the baseband signals that can result, for example, from amplifiers, analog-to-digital converters (ADC), and other elements in the reader 100. The phase offsets can include offsets in the phase of the baseband signals that can result, for example, from quadrature modulators, summers, low noise amplifiers, down conversion mixers,

baseband filters, and others. By compensating for such DC offsets and/or phase offsets, the reader 100 may enhance, maximize, or increase the reduction in the leakage signal in the receive path.

In some implementations, the reader 100 may estimate the transmit signal as:

$$5 \quad x(t) = A(t) \cos(2\pi f_c t + \phi(t) + \theta),$$

where $A(t)$ represents slow amplitude variations, $\phi(t)$ represents the oscillator phase noise, and θ represents the phase angle of the transmit signal out of, for example, a power amplifier. In addition to the receive signal in the receive path, the reader may also include leakage signals in the receive path and the combination of these signals may be expressed as:

$$10 \quad y(t) = r(t) + c(t) \cdot A(t) \cos(2\pi f_c t + \phi(t) + \theta(t) + \theta),$$

where $r(t)$ is the receive signal from transponders and other RF environmental signals and the rest may estimate the leakage signal from the transmitter, where, in some implementations, $c(t) \ll 1$ and $0 \leq \theta(t) < 2\pi$ can represent slow variations in transmit leakage amplitude and/or phase and both can vary slowly over time. During the course of this description, the leakage current is described in polar coordinates but may also be described in other coordinates such as rectangular. In some implementations, the leakage current may be expressed as a portion of an in-phase signal and a quadrature signal, as discussed in more detail below.

20 In the illustrated implementation, the reader 100 includes a carrier-noise-reduction (CNR) module 102, a receiver module 104, and a transmitter module 106. The CNR module 102 includes any software, hardware, and/or firmware operable to reduce leakage signals in the receive path. For example, the CNR module 102 may add signals to the receive path for canceling, minimizing, or otherwise reducing leakage signals. In the illustrated implementation, the CNR module 102 includes a power splitter 108, a quadrature modulator 110, a summer 112, a dual digital-to-analog converter (DAC) 114, and a calibration switch 116. The power splitter 108 splits or otherwise directs a portion of the transmit signal to the quadrature modulator 110. In some implementations, the portion of the transmit signal may be expressed as:

$$30 \quad u(t) = b_1 \cdot x(t),$$

where b_1 is a fixed small constant (e.g., $b_1 = 0.05$). In addition to receiving a portion of the transmit signal, the quadrature modulator 110 receives an in-phase control signal

$v_i(t)$ and a quadrature control signal $v_q(t)$. In some implementations, the control signals may be polar controls. The quadrature modulator 110 can modulate the portion of the transmit signal (e.g., $u(t)$) and the baseband quadrature control signals $v_i(t)$ and $v_q(t)$ to generate a cancellation signal for the leakage signal. In some

5 implementations, the quadrature modulator 110 includes a vector modulator.

In some implementations, the quadrature modulator 110 may estimate the cancellation signals as:

$$z(t) = b_2 A(t) (v_i(t) \cos(2\pi F_c t + \phi(t) + \theta) + v_q(t) \sin(2\pi F_c t + \phi(t) + \theta))$$

where b_2 is a fixed small constant (e.g., $b_2 = 0.01$). In some implementations, the

10 constant b_2 accounts for the combined signal attenuation through the power splitter (b_1) and the quadrature modulator 110. In the example expression for the cancellation signal, the quadrature modulator 110 uses the input $u(t)$ to generate a 90 degree shifted version (sine), then modulates the control signals $v_i(t)$ and $v_q(t)$ onto these cosine and sine carriers, respectively, to produce the cancellation signal.

15 After generating the cancellation signal, the quadrature module 110 directs the cancellation signal to the summer 112. The summer 112 subtracts the cancellation signal from the signal received from the receiver which includes the leakage signal. In the example, the summer 112 subtracts the quadrature modulator output signal $z(t)$ from the receiver input $y(t)$ to produce $s(t)$. In some implementations, the residual

20 signal $s(t)$ substantially equals the desired receive signal $r(t)$, i.e., substantially all of the transmitter leakage is cancelled. The CNR module 102 can represent the residual signal as:

$$s(t) = b_2 A(t) (c(t) \cdot \cos(2\pi F_c t + \phi(t) + \theta(t) + \theta) + v_i(t) \cos(2\pi F_c t + \phi(t) + \theta) + v_q(t) \sin(2\pi F_c t + \phi(t) + \theta)) + r(t)$$

25 In some implementations, the CNR module 102 includes the dual DAC for converting digital control signals to analog control signals and directing the analog control signals to the quadrature modulator 110. In some implementations, the control signals are generated as a sampled data signal and these signals are passed through a dual digital-to-analog converter (DAC) to create the analog control signals for the quadrature

30 modulator 110. In other words, the control signals $v_i(t)$ and $v_q(t)$ can comprise digital

signals received from the dual DAC 114. In some implementations, the control signals $v_i(t)$ and $v_q(t)$ may be generated from analog control circuitry. The calibration switch 116 can substantially prevent input signals into the receive module 104 when the DC offsets and/or the phase offsets are estimated. The reader 100 also includes a circulator 140. The circulator 140 directs the transmit signals towards the antenna and also directs receive signals from the antenna to the CNR module 102. The circulator 140 could be replaced with a coupler or separate transmit and receive antennas could be used, commonly known as a bi-static antenna configuration.

The receiver module 104 can include any software, hardware, and/or firmware operable to down convert the received signal to baseband signals for processing by the DSP 130. For example, the receiver module 104 may convert an RF signal to a baseband signal. In some implementations, the baseband signal is a low frequency signal (e.g., DC to 400 KHz). In addition, the receiver module 104 may perform other functions such as amplification, filtering, conversion between analog and digital signals, and/or others. The receiver module 104 may produce the baseband signals using a mixer and low pass filters. In the illustrated implementations, the receiver module 104 includes a low noise amplifier (LNA) 118, a mixer 120, a low pass filters (LPFs) 122 and 124, and a dual ADC 126. The LNA 118 receives the residual signal from the summer 112 and amplifies the residual signal in light of the relative weakness of the signal to the transmission signal. The mixer 120 mixes the residual signal with a signal received from a frequency synthesizer 128 to generate two component signals. In the illustrated implementation, the mixer 120 generates an in-phase signal and a quadrature signal. For example, the receiver module 104 can amplify the residual signal $s(t)$ using the LNA 118 and then mix down the signal to baseband using a combination of the quadrature mixer 120 and the LPFs 122 and 124. The LPFs 122 and 124 can reject the out of band energy of transceivers in neighboring channels. In doing so, the effect of out of band noise can be made relatively small through intelligent selection of band-limiting baseband filters. In some implementations, the signals generated from the down conversion may be substantially estimated as:

$$\begin{aligned}
e_i(t) &= \frac{b_1 A(t)}{2} \left(c(t) \cdot \cos(\phi(t) + \theta(t)) + v_i(t) \cos \phi(t) + v_q(t) \sin \phi(t) \right) \\
&\quad + r(t) \cos(2\pi F_c t) \\
&= \frac{b_1 A(t)}{2} \left((c(t) \cdot \cos \theta(t) + v_i(t)) \cos \phi(t) + (v_q(t) - c(t) \cdot \sin \theta(t)) \sin \phi(t) \right) \\
&\quad + r(t) \cos(2\pi F_c t)
\end{aligned}$$

and

$$\begin{aligned}
e_q(t) &= \frac{b_2 A(t)}{2} \left(-c(t) \cdot \sin(\phi(t) + \theta(t)) - v_i(t) \sin \phi(t) + v_q(t) \cos \phi(t) \right) \\
&\quad + r(t) \sin(2\pi F_c t) \\
&= \frac{b_2 A(t)}{2} \left((v_q(t) - c(t) \cdot \sin \theta(t)) \cos \phi(t) - (c(t) \cdot \cos \theta(t) + v_i(t)) \sin \phi(t) \right) \\
&\quad + r(t) \sin(2\pi F_c t)
\end{aligned}$$

- 5 In this case, the following control signals $v_i(t)$ and $v_q(t)$ may be used to substantially eliminate the leakage signal:

$$v_i(t) = -c(t) \cdot \cos \theta(t)$$

and

10
$$v_q(t) = c(t) \cdot \sin \theta(t).$$

A number of primary and secondary circuit and/or system impairments can limit performance of the reader 100. To indicate this difference, the baseband signals, *i.e.*, the in-phase signal and the quadrature signal, into the dual ADC 126 are denoted as $f_i(t)$ and $f_q(t)$ as compared with $e_i(t)$ and $e_q(t)$.

- 15 The receiver module 104 passes or otherwise directs the baseband signals to the digital signal processor (DSP) 130. The DSP 130 can include any software, hardware, and/or firmware operable to process the residual signal. For example, the DSP 130 may generate control signals for adjusting the cancellation signal used to compensate for leakage signal. In some implementations, the DSP 130 compensates the baseband signals for DC offset and/or phase offset. As mentioned above, the reader 100 may include elements that subtract DC offsets and/or de-rotate phase offsets in the baseband signals. Otherwise, these offsets can reduce the efficacy of the cancellation signal in reducing the leakage signal. In other words, the DSP 130 may eliminate, minimize, or otherwise reduce the DC offset and/or the phase offset to reduce error in the
- 20

cancellation signal. In the case of DC offset, the DSP 130 can, in some implementations, subtract estimates of the DC offsets in the baseband signals such as the in-phase signal and the quadrature signal. For example, the DSP 130 may determine samples (*e.g.*, hundreds of samples) of the DC offset for the baseband signals and generate an average for each baseband signal based, at least in part, on the samples. In this example, the DSP 130 may subtract the DC offset from the corresponding baseband signal during steady state. In regards to the phase offset, the DSP 130 may introduce a phase shift in the baseband signals to minimize, eliminate, or otherwise reduce the phase shift generated by the elements in the reader 100. In some cases, varying a control value on one baseband signal (*e.g.*, in-phase signal) can produce a change on the other baseband signal (*e.g.*, quadrature signal). This cross-coupling between the two baseband signals can, in some implementations, lead to a more complex control algorithm for compensating for the phase shift offset.

The transmitter module 106 can include any software, hardware, and/or firmware operable to generate transmission signals for transponders. In the illustrated implementation, the transmitter module 106 includes a DAC 132, a LPF 134, a transmission mixer 136 and a power amplifier 138. The DAC 132 receives a digital signal from the DSP 130 and converts the digital signal to analog signals. For example, the digital signal can encode queries for transponders to identify associated information. The DAC 132 passes the analog signal to the LPF 134 to attenuate higher frequencies than a cutoff frequency from the analog signals. The LPF 134 passes the analog signals to the transmission mixer 136 to upconvert the baseband signals to an RF signals. In this case, the transmission mixer 136 receives a signal from the frequency synthesizer 128 and mixes this signal with the analog signal to generate the RF signal. The power amplifier 138 amplifies the RF signal and directs the amplified signal to the power splitter 108. In some implementations, the power splitter 108 may comprise a coupler.

FIGURE 2 is a baseband equivalent model 200 of the CNR control loop of the reader 100 in FIGURE 1. In particular, this baseband model 200 is mathematically a substantially equivalent model of the reader 100 with the RF carrier removed. The baseband model 200 includes the loops 202 and 204 which are associated with the in-phase signal and the quadrature signal. A portion of the loops 202 and 204 include associated control signals and are illustrated as the in-phase control signal $v_i(t)$ and

the control quadrature signal $v_q(t)$ discussed above. As previously discussed, DC offsets are impairments that result from the elements in reader 100 and are typically associated with DC coupled applications. Phase offsets are impairments that result from the elements in reader 100 and are typically associated with RF and quadrature baseband applications.

Regarding the DC offsets, the loops 202 and 204 are effectively DC coupled loops and, as a result, DC offsets in the signal paths can directly effect the estimated control signals $v_i(t)$ and $v_q(t)$. Such DC offsets are represented in the model 200 as the DC offsets 206a and 206b. As discussed above, the DSP 130 eliminates, minimizes, or otherwise reduces these DC offsets from the loops 202 and 204. In the illustrated implementation, the DSP 130 includes a DC-offset-removal module 208 to subtract DC offsets from the in-phase signal and the quadrature signal. In addition, the module 208 may sample the baseband signals to estimate the DC offsets. For example, the module 208 may take hundreds of samples to determine average DC offsets to subtract from the baseband signals.

Regarding the phase-shift offsets, the elements in the reader 100 can impart a phase shift in the loops 202 and 204 and, as a result, this phase shift can directly effect the estimated control signals $v_i(t)$ and $v_q(t)$. For example, the phase shift can be due to quadrature modulator, summer, low noise amplifier, down conversion mixer, baseband filtering, and other elements. Such phase shifts in the loops 202 and 204 are represented in the model 200 as unknown phase shift 210. As discussed above, the DSP 130 eliminates, minimizes, or otherwise reduces these phase-shift offsets from the loops 202 and 204. In the illustrated implementation, the DSP 130 includes a phase rotation module 212 to de-rotate the in-phase signal and the quadrature signal by angle θ . In some implementations, the de-rotation is performed by a standard complex multiply of $e^{-j\theta}$. In addition, the module 212 may sample the baseband signals to estimate the phase-shift offsets. For example, the module 212 may take hundreds of samples to determine an average phase shift for each signal and de-rotate each signal in accordance with the associated averages.

In addition, the DSP 130 includes gains 214a-b and integrators 216a-b. The gains 214a-b allow the tracking bandwidth of the leakage cancellation system to be adjusted. The gains 214a and 214b may generate a gain value on each loop 202 and

204. In some implementations, the gains 214a and 214b generate gain values in light of a desire for fast convergence and loop stability. Further, the gain value can be adjusted over time to be large at first for quick approximation and then later made smaller to improve accuracy in the final results. Lower gain values reduce the bandwidth of the leakage cancellation system and make the system less responsive to noise signals. The integrators 216a-b filter the error signals to produce accurate control outputs.

The leakage path is illustrated in the model as the transmitter leakage function 218. This function 218, shown as a single element, typically results from a number of leakage paths, one of which can be the circulator 140. These leakage paths combine to yield a composite transmitter leakage function 218. The leakage signal is often a sinusoid of some general amplitude and phase where each is generally a function of the transmit frequency. In some implementations, the leakage signal can be an unmodulated sinusoid, because the transmitter is frequently unmodulated during the receive mode of operation. Though, the concept could be applied successfully as well with a relatively slowly modulated transmit carrier being used during receive operations. As mentioned above, the leakage signal of interest could be viewed as a sinusoid of some amplitude and phase and can be expressed in polar form.

FIGURES 3-5 are flowcharts illustrating example methods 300, 400, and 500 for reducing impairments in baseband signals in accordance with some implementations of the present disclosure. Generally, method 300 describes an example technique for determining average DC offsets for the baseband signals. Method 400 generally describes an example technique for determining average phase offsets for the baseband signals. Method 500 generally describes an example technique for adjusting baseband signals for DC offsets and phase offsets prior to determining control signals associated with leakage signals. The reader contemplates using any appropriate combination and arrangement of logical elements implementing some or all of the described functionality.

Regarding FIGURE 3, the method 300 begins at step 302 where a calibration switch is opened. For example, the calibration switch 116 can be opened. At step 304, control value outputs are set to zero. In the example, the control value outputs $v_i(t)$ and $v_q(t)$ generated by the DSP 130 can be set to zero. Next, at step 306, the DC offsets are estimated for the in-phase signal and quadrature signal by taking several

samples (e.g., hundreds of samples). Returning to the example, the DSP 130 may take several samples of the baseband signals and average the samples to determine the DC offsets for each baseband signal. The in-phase DC offset and the quadrature DC offset are stored for use in the steady state control routine at step 308. As for the example, the DC offset removal module 208 may subtract the DC offset averages from the baseband signals while the reader 100 is in operation.

Referring to FIGURE 4, the method 400 begins at step 402 where a calibration switch is opened. For example, the calibration switch 116 can be opened. At step 404, control value outputs are set to a known constant non-zero value. In the example, the control value outputs $v_i(t)$ and $v_q(t)$ generated by the DSP 130 can be set to $v_i(t)=1$ and $v_q(t)=0$. This example would ideally produce a baseband error signal with zero quadrature signal if there were no phase shift. Next, at step 406, the DC offsets are estimated for the in-phase signal and quadrature signal by taking several samples (e.g., hundreds of samples) from ADC inputs, averaging these samples, and subtracting the previously estimated DC offsets. Returning to the example, the samples for the phase offset may be determined from the inputs to the ADC 126. At step 408, the phase shift offset for the baseband signals may be estimated as the arctangent of the ratio of the quadrature-phase response to the in-phase response. The estimated phase shift offsets are stored for use in the steady state control routine at step 410. As for the example, the phase rotation module 212 may de-rotate the baseband signals in accordance with the estimated phase shift.

Referring to FIGURE 5, the method 500 begins at step 502 where estimates of the DC offsets of the baseband signals are estimated (this step was previously detailed in Figure 3). For example, the DSP 130 may estimate the DC offsets for the in-phase signal and the quadrature signal by averages several samples. At step 504, the phase shift offsets are estimated (this step was previously detailed in Figure 4). For example, the phase shift offsets may be measured from the inputs of ADC 126 and taking the average of several samples of the measured shifts. Next, at step 506, the calibration switch is closed to allow input signals. In the example, the calibration switch 116 may be closed to allow the input signal to be processed by the reader 100. The ADC inputs are measured at step 508. For example, the inputs to the dual ADC 126 can be measured. At step 510, the estimated DC offsets are subtracted from the measured

ADC inputs. Returning to the example, the DC-offset-removal module 208 may subtract the estimated DC offsets from the baseband signals. Next, at step 512, the inputs are rotated by the estimated phase shift to decouple the two control loops. As for the example, the phase rotation module 212 may de-rotate the baseband signals using the estimated phase shift. At step 514, a gain is applied to the DC compensated, phase offset de-rotated, baseband error signals, which can represent the control loop “error” signals. The gained error signals are integrated to produce new control signals at step 516. For example, the integrators 216a and 216b may integrate the gained error signals to generate the control signals $v_i(t)$ and $v_q(t)$. The integrated signals are sent to a DAC at step 518. In the example, the integrators 216a and 216b may send the control signals $v_i(t)$ and $v_q(t)$ to the dual DAC 114. At step 520, a period is allowed for the CNR circuit to settle with the new control values. If the system is still operating at decisional step 522, the method 500 returns to the step 508. Otherwise, execution of method 500 ends.

A number of embodiments of the invention have been described. Nevertheless, it will be understood that various modifications may be made without departing from the spirit and scope of the invention.

WHAT IS CLAIMED IS:

1. An RFID reader comprising:
 - a RF antenna;
 - a transmitter section coupled to the RF antenna and operable to generate
 - 5 a transmit signal to be transmitted by the RF antenna;
 - a receiver section coupled to the RF antenna and operable to receive a receive signal from the RF antenna, the receiver section further comprising:
 - a de-rotation module operable to de-rotate, by θ , an in-phase signal and quadrature signal associated with the leakage signal; and
 - 10 a control module operable to generate control signals used to produce a signal for reducing the leakage signal in a receive path of the reader; and
 - a carrier noise reduction (CNR) section operable to subtract the reduction signal from the leakage signal.
- 15 2. The RFID reader of claim 1, further comprising a DC offset module operable to subtract DC offsets from the in-phase signal and the quadrature signal associated with the leakage signal.
- 20 3. The RFID reader of claim 2, wherein the control signals for the reduction signal are determined after the in-phase signal and quadrature signal associated with the leakage signal are de-rotated and the DC offsets are subtracted.
- 25 4. The RFID reader of claim 2, further comprising a DC offset estimation module operable to determine a plurality of samples of the DC offsets of the in-phase signal and the quadrature signal associated with the leakage signal and determine the estimated DC offsets based, at least in part, on the average of the samples.
- 30 5. The RFID reader of claim 2, wherein the DC offsets are subtracted from the in-phase signal and the quadrature signal after downconverting the RF to the in-phase signal and the quadrature signal.

6. The RFID reader of claim 1, further comprising a phase estimation module operable to determine a plurality of samples of the phase offset of the in-phase signal and the quadrature signal associated with the leakage signal and determine the de-rotation angle based, at least in part, on the average of the samples.

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7. The RFID reader of claim 1, wherein the leakage signal is reduced by over 20dB.

8. The RFID reader of claim 1, wherein the de-rotation angle is based, at least in part, on a phase offset associated with the in-phase signal and a phase offset associated with the quadrature signal.

9. The RFID reader of claim 1, further comprising a quadrature modulator operable to receive the control signals and a portion of a transmission signal to generate the reduction signal.

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10. The RFID reader of claim 1, wherein the in-phase signal and the quadrature signal are de-rotated after downconverting the RF signal to the in-phase signal and the quadrature signal.

20

11. A method, comprising:
transmitting an RF signal in an interrogation zone;
downconverting an RF signal on a receive path to an in-phase and a quadrature signal;

25 determining a phase-shift offset associated with leakage signals from the transmitted RF signal;

de-rotating the in-phase signal and the quadrature signal using the determine phase-shift offset;

30 generating control signals used to produce a signal for reducing the leakage signal in the receive path based, at least in part, on the de-rotated in-phase signal and the de-rotated quadrature signal; and

subtracting the reduction signal from the leakage signal.

12. The method of claim 11, further comprising:
determining DC offsets of the in-phase signal and the quadrature signal; and
subtracting the DC offsets from the in-phase signal and the quadrature signal
associated with the leakage signal.

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13. The method of claim 12, wherein the control signals for the reduction
signal are determined after the in-phase signal and quadrature signal associated with the
leakage signal are de-rotated and the DC offsets are subtracted.

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14. The method of claim 12, wherein the DC offsets are determined based,
at least in part, on a plurality of samples of the DC offsets of the in-phase signal and the
quadrature signal associated with the leakage signal.

15

15. The method of claim 12, wherein the DC offsets are subtracted from the
in-phase signal and the quadrature signal after downconverting the RF signal to the in-
phase signal and the quadrature signal.

20

16. The method of claim 11, wherein the phase-shift offset is determined,
based at least in part, on a plurality of samples of the phase-shift offset of the in-phase
signal and the quadrature signal associated with the leakage signal.

17. The method of claim 11, wherein the leakage signal is reduced by 20 dB
or greater.

25

18. The method of claim 11, wherein the de-rotation angle is based, at least
in part, on a phase-shift offset associated with both the in-phase signal and a phase
offset associated with the quadrature signal.

30

19. The method of claim 11, further comprising generating the reduction
signal based, at least in part, on a portion of the transmission signal and the control
signals.

20. The method of claim 11, wherein the in-phase signal and the quadrature signal are de-rotated after downconverting the RF signal to the in-phase signal and the quadrature signal.

115

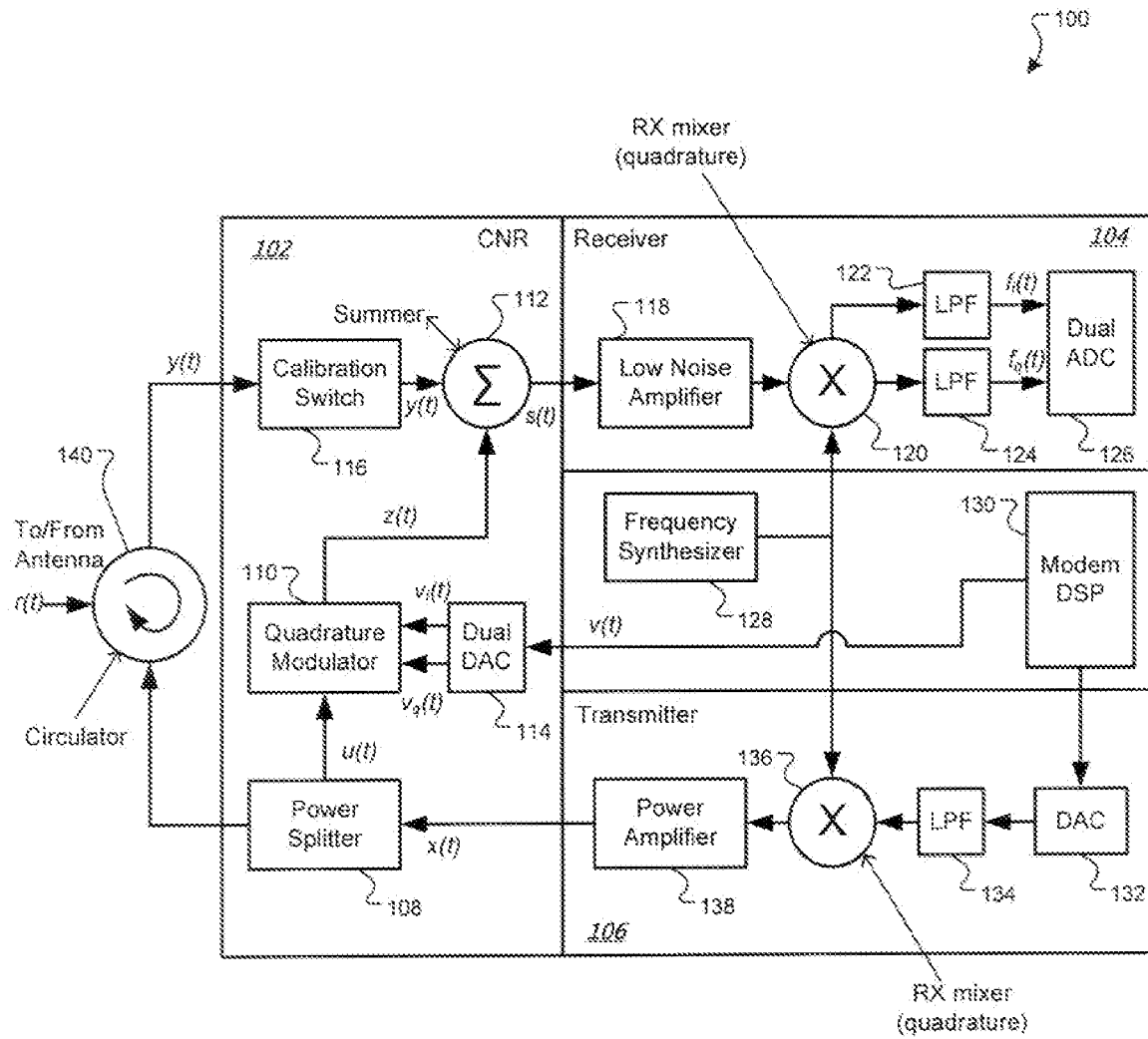


FIG. 1

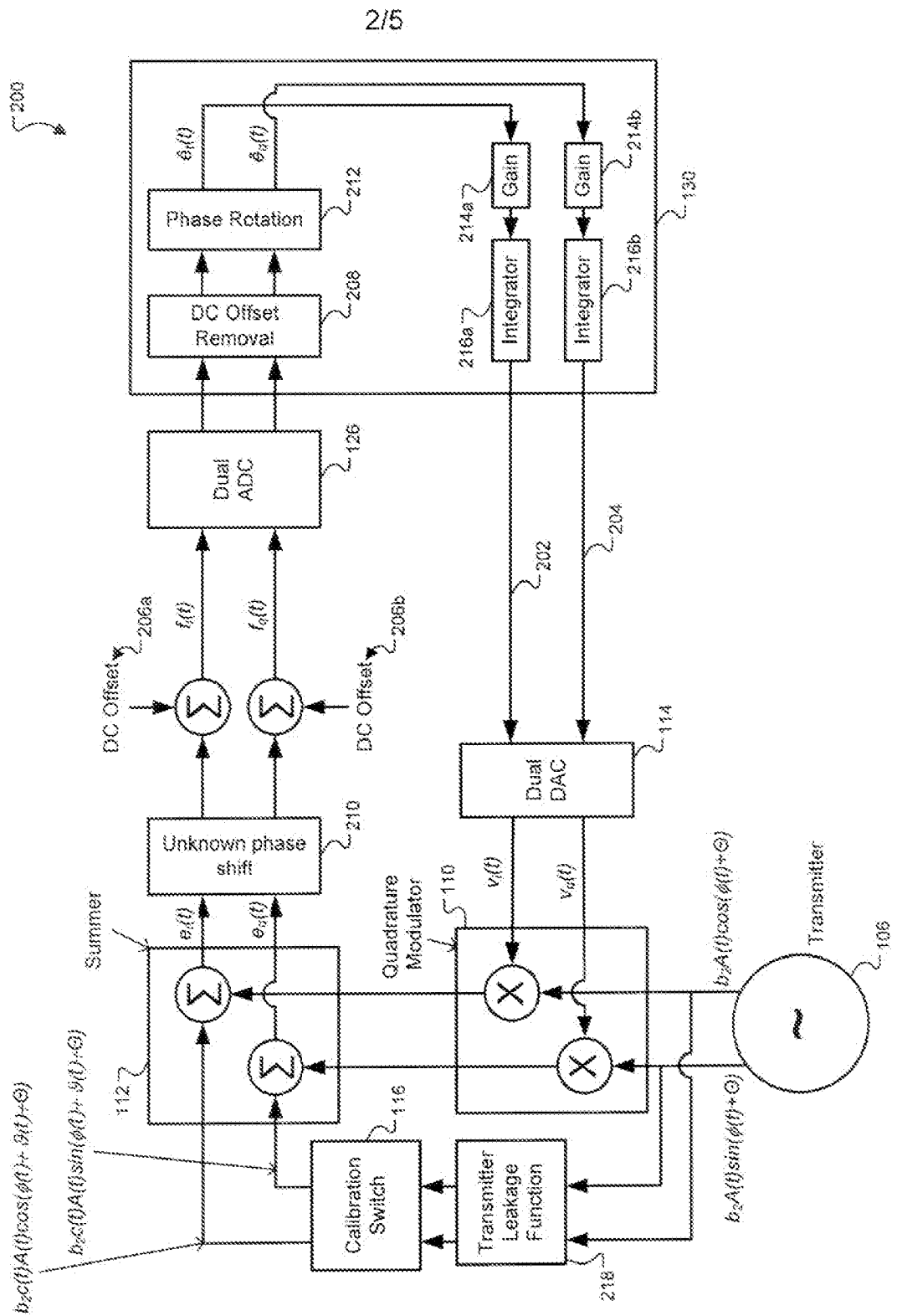


FIG. 2

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300 ↗

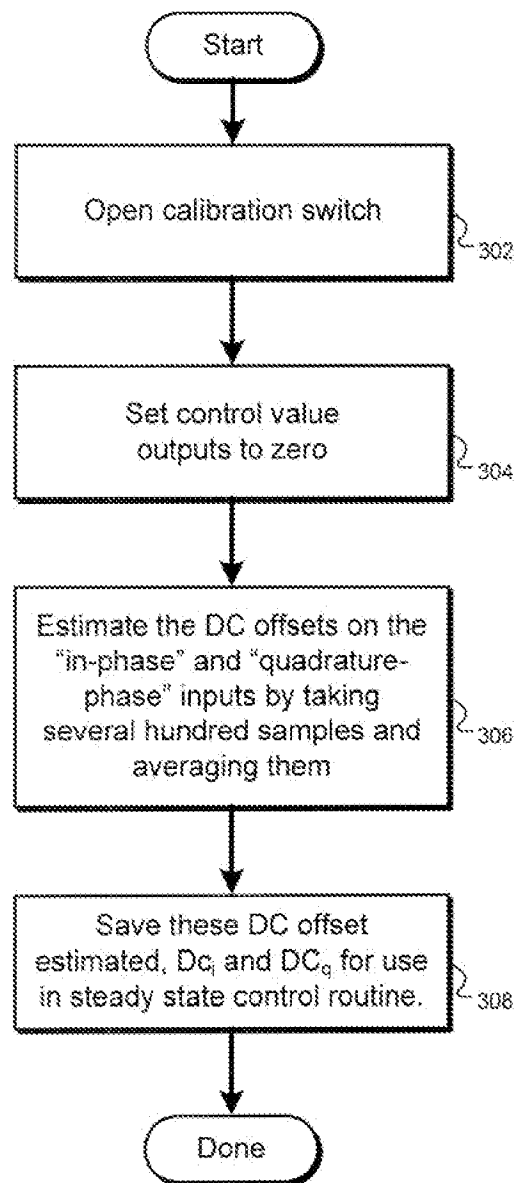


FIG. 3

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400

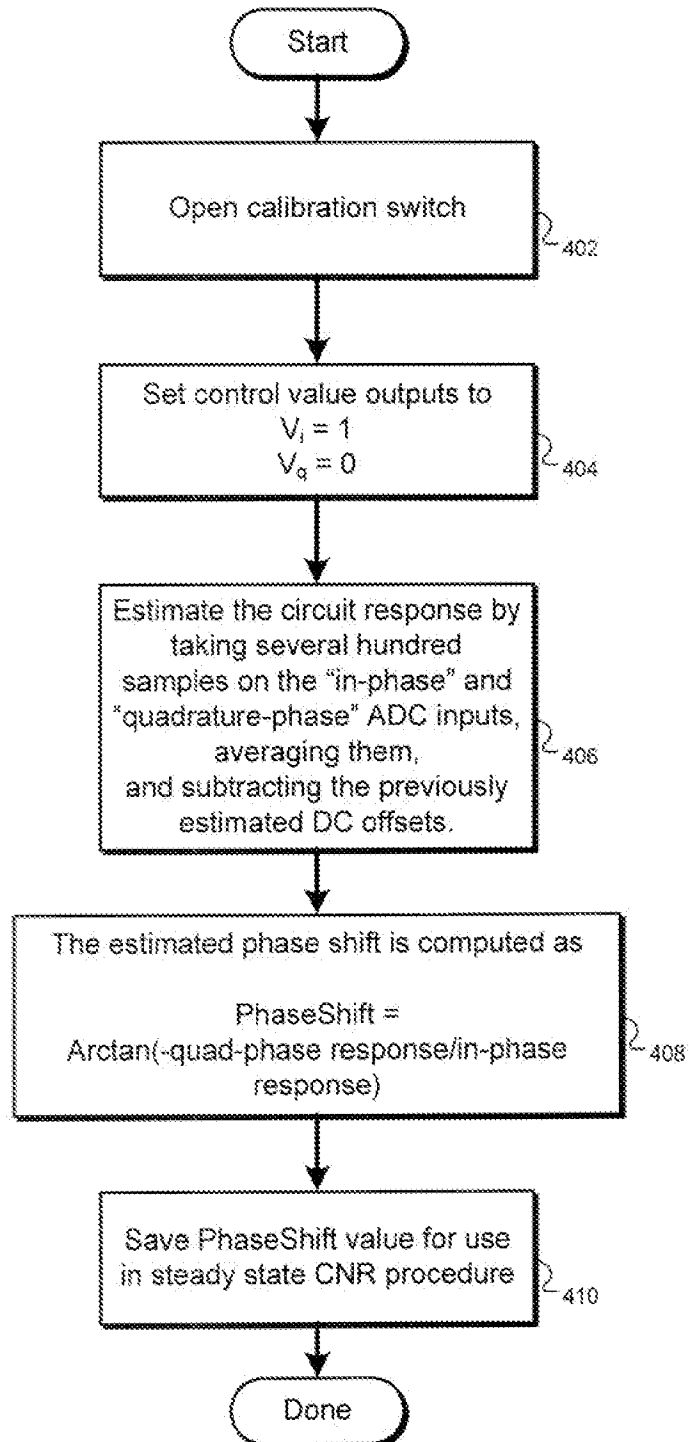


FIG. 4

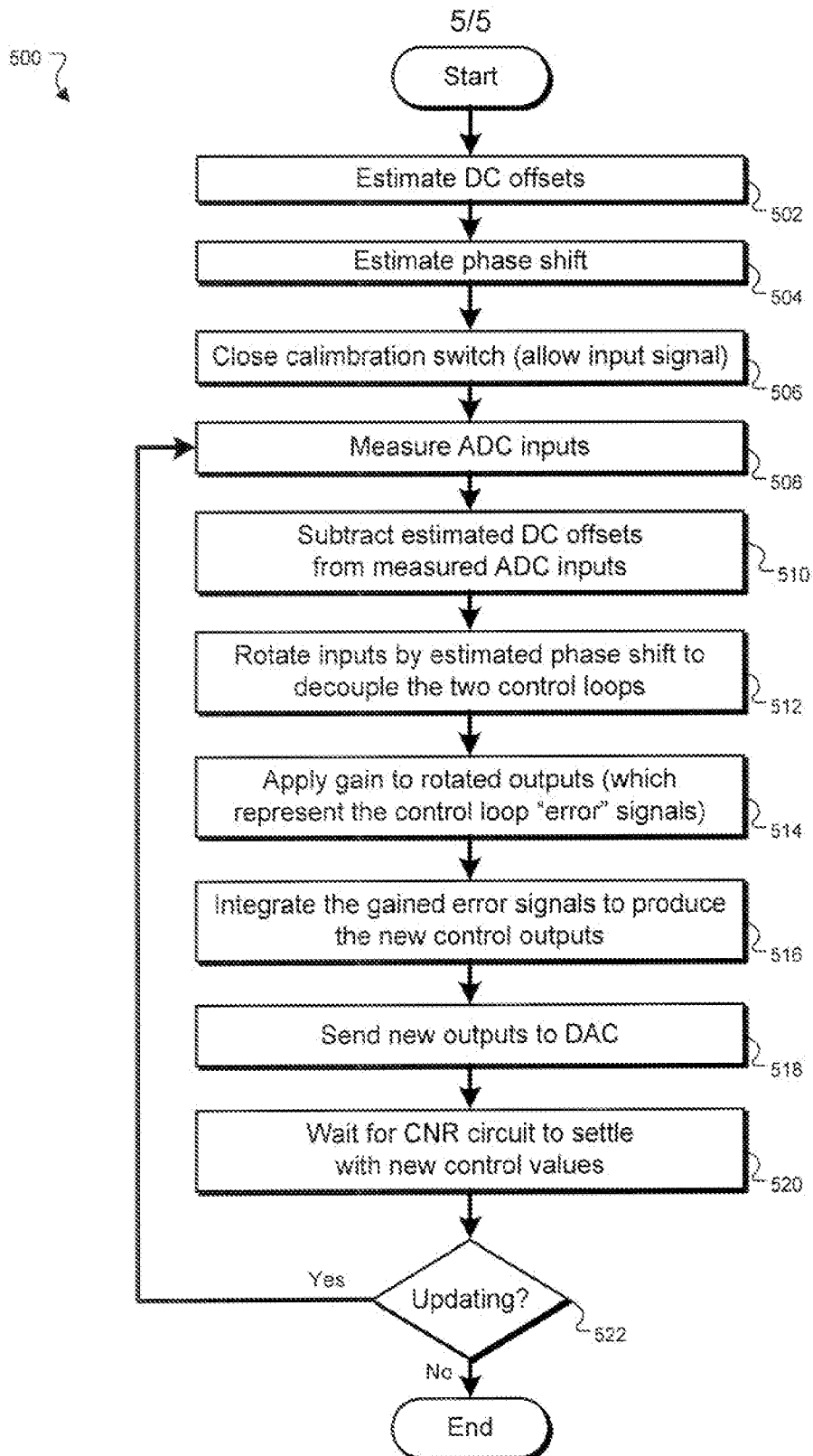


FIG. 5