Invention Title
Time-Domain Method of Low-Complexity Frequency Dependent IQ Imbalance Compensation

Background
The increasing demand for low-cost and low-power architecture of wireless communication design provides a need for a direct conversion RF transceiver because it simplifies down-conversion by removing intermediate frequency (IF) analog components. However, a direct-conversion RF transceiver suffers from imbalance between analog in-phase (I) and quadrature (Q) branches that arise from the imperfect analog finite-element (FE) components. There are two main sources of IQ imbalance.

First, imperfect I/Q down-conversion may generate the gain and phase imbalance between in-phase and quadrature components. Gain mismatch may arise from unequal gains of a mixer, unequal gains of LO drivers that supply LO clock to the I and Q branches, unequal gains in VGA components of I and Q branches, and unequal LSB levels of the analog-to-digital converters in I and Q branches, or any combinations of the above. The phase imbalance primarily arises from difficulty in achieving precise 90° phase between I and Q clocks. Since these types of imbalance do not depend on signal frequency, they may be referred to as a frequency-independent (FI) imbalance.

Second, analog baseband (ABB) filter pole position mismatch between analog I and Q paths may cause frequency-dependent (FD) IQ mismatch. The FI and FD IQ imbalances results in a mirror image signal in signal bandwidth. Typical image rejection ratio (IRR) at the receiver side ranges from 20 to 40dB, which is not sufficient to correctly receive high-order modulated carriers that require high SNR.

There has been a lot of research in developing schemes that estimate and compensate for I/Q mismatch. Two principal approaches are a frequency-domain method and a time-domain method. The frequency domain method reduces the complexity of I/Q compensation as compared to a time -domain method due to conversion of convolution into product operations in a frequency domain. However, the frequency domain method requires special pilot patterns for I/Q imbalance estimation. The overhead from pilot addition is unavoidable, thus reducing the spectral efficiency and available throughput. Therefore, the frequency-domain method to IQ mismatch cancellation is not supported in leading wide area network (WAN) standards (e.g., long-term evolution (LTE), wideband code division multiple access (WCDMA)).

The time-domain method uses blind estimation by exploiting the orthogonality property of received signal to compensate FI and FD mismatch. The time-domain method does not require special pilot patterns to estimate I/Q mismatch, which enables efficient utilization of a wireless channel.

Summary
The present system includes a digital finite impulse response (FIR) filter that generates a compensation factor to cancel out an image term in the IQ imbalance signal. The compensation factor is added to the IQ imbalance signal to generate a compensated digital signal. Either a real part or an imaginary part of the complex compensated digital signal is used for a digital FIR filter input to generate a compensation factor. The digital FIR filter may update its complex coefficients by either online or offline estimation. The present system may be used either in TX IQ imbalance compensation or RX IQ imbalance compensation. The present system provides low-complexity architecture of time-domain FD IQ imbalance compensation.

Detailed Description

The present disclosure describes a variant of time-domain frequency-dependent IQ imbalance compensation. A time-domain method provides efficient utilization of wireless spectrum. However, due to convolution (filtering) operations in the time-domain, even small number of filter tap operations leads to high hardware complexity and power consumption. Furthermore, due to its nature of blind estimation, the time-domain method requires high resolution of IQ imbalance estimates in order to minimize estimation errors by long-term average, which leads to additional increase in complexity. The present system provides an efficient design of time-domain IQ compensation.

1. I/Q signal models

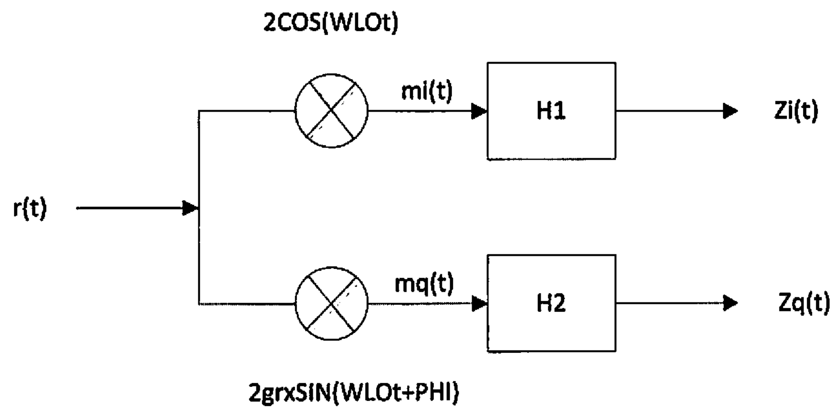


Figure 0: General receiver structure based on I/Q down-conversion

Referring to Figure 0, $r(t) = 2 \cdot \mathbf{real}\{s(t)e^{j\omega_{LO}t}\}$ is a received RF signal. Complex LO signal $\mathbf{x}_{LO}^{\mathbf{RX}}(t)$ may be written as:

(1.1)

$$\mathbf{x}_{LO}^{\mathbf{RX}}(t) = \cos(\omega_{LO}t) - jg\sin(\omega_{LO}t + \phi)$$

$$= K_1 e^{-j\omega_{LO}t} + K_2 e^{j\omega_{LO}t}$$

where $[g, \phi]$ represents the effective amplitude and phase FI imbalance of the RX-path, and the coefficients K_1 and K_2 are of the form:

$$K_1 = \frac{1 + ge^{-j\phi}}{2}, \qquad K_2 = \frac{1 - ge^{j\phi}}{2} \tag{1.2}$$

The down-converted signal $m(t)$ may be expressed as:

$$m(t) = r(t) \cdot \mathbf{x}_{LO}^{\mathbf{RX}}(t) = K_1 s(t) + K_2 s^*(t) \tag{1.3}$$

For equation (1.3), high-order frequency components are ignored. After $m(t)$ goes through low-pass filters of $h_1(t)$ and $h_2(t)$, filter output $z(t)$ becomes:

$$z(t) = m_i(t) * h_1(t) + jm_q(t) * h_2(t) \tag{1.4}$$

$$= \frac{m(t)+m^*(t)}{2} * h_1(t) + j\frac{m(t)-m^*(t)}{2j} * h_2(t)$$

$$= g_1(t) * s(t) + g_2(t) * s(t)^*$$

where $g_1(t) = \frac{1}{2}(h_1(t) + g_{rx}e^{-j\phi}h_2(t))$ and $g_2(t) = \frac{1}{2}(h_1(t) - g_{rx}e^{j\phi}h_2(t))$ are combined response of FI and FD mismatch.

If there's no mismatch between $h_1(t)$ and $h_2(t)$ and so if $h_1(t)$ and $h_2(t)$ are equal to $h(t)$, $g_1(t)$ and $g_2(t)$ reduces to $h(t)K_1$ and $h(t)K_2$. For that case, there's no frequency-dependent IQ mismatch because $h(t)$ will be canceled out in IRR.

In frequency domain, the Fourier transform of Equation (1.4) may be taken as:

$$Z(f) = G_1(f)S(f) + G_2(f)S(-f)^* \tag{1.5}$$

The corresponding mirror-frequency attenuation, or image reduction ratio (IRR) may be given as:

$$IRR = L_{RX} = 10\log_{10}\frac{|G_1(f)|^2}{|G_2(f)|^2} = 10\log_{10}\frac{|H_1(f) + g_{rx}e^{-j\phi}H_2(f)|^2}{|H_1(f) - g_{rx}e^{j\phi}H_2(f)|^2} \tag{1.6}$$

If $H_1(f) = H_2(f)$, (1.6) reduces to $10\log_{10}\frac{|K_1|^2}{|K_2|^2}$, which is IRR of frequency-independent imbalance. An RF transceiver incurs an IQ imbalance due to a quadrature down-conversion structure. A typical range for 25-40 dB IRR is 1-5 % gain mismatch and 1-5 degrees of phase mismatch. Image suppression with more than 40dB is required to support high-order modulated carriers such as 256Q in LTE.

2. Adaptive filter compensation model

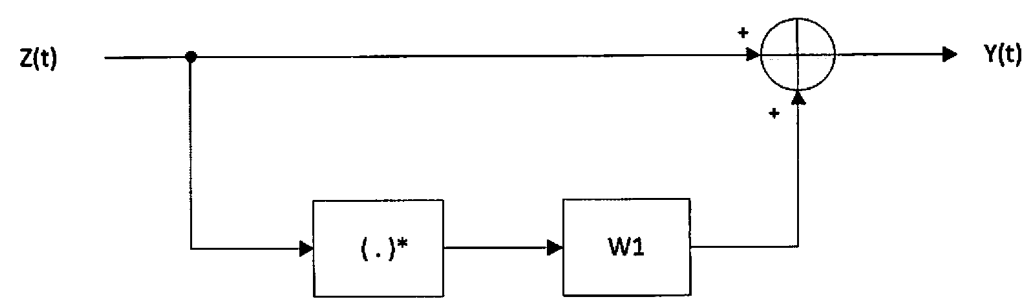


Figure 2: Adaptive filter compensation model

Figure 2 provides an adaptive filter model that compensates IQ frequency-dependent imbalance. Figure 2 shows the block diagram of an adaptive filter. From the block diagram, output of the adaptive filter $y(t)$ becomes:

$$y(t) = z(t) + w1(t) * z^*(t) \tag{2.1}$$

If $z(t)$ is substituted from (1.4), (2.1) can be transformed into:

$$y(t) = (g_1(t) + w1(t) * g_2^*(t)) * s(t) + (g_2(t) + w1(t) * g_1^*(t)) * s^*(t) \tag{2.2}$$

Therefore, optimal $w1(t)$ that cancels out mirror image is:

$$W1_{OPT}(f) = -\frac{G_2(f)}{G_1^*(-f)} \tag{2.3}$$

which makes $g_2(t) + w1(t) * g_1^*(t)$ become zero.

Figure 2 suggests a newton method for adaptation of filter taps. The only property that may be used for time-domain adaptation is properness condition. For frequency-dependent IQ imbalance, the following conditions for adaptation may be used:

$$c_y(\tau) = E\{y(t)y(t-\tau)\} = 0, for\ 0 \le \tau \le \tau_{\max} \tag{2.4}$$

where $\tau_{\max}$ is a system parameter that determines the number of filter taps. An objective function is $C_y = E\{Y(t)y(t)\}$, where $Y(t) = [y(t)\ y(t-1)\ \cdots\ y(t-N+1)]^T$ and N is the number of filter coefficients derived from $\tau_{\max}$. The approximate newton method to find filter coefficients to satisfy $C_y = 0$ according to Figure 2 is:

$$w(n+1) = w(n) - \alpha Y(t)y(t) \tag{2.5}$$

where α is a step size of each update.

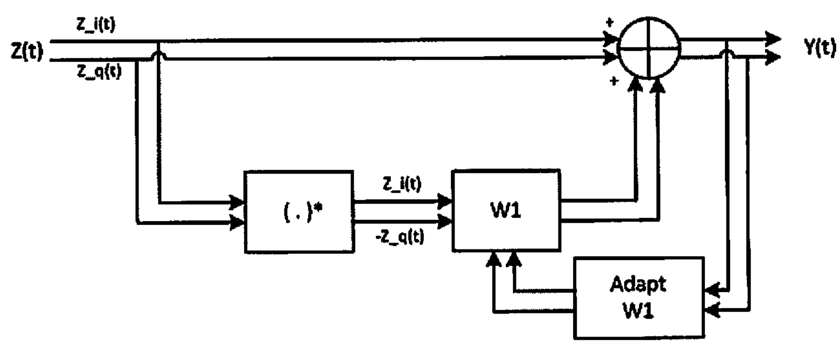


Figure 3: Time-domain IQ compensation

Referring to Figure 3, output $y(t)$ becomes:

$y(t) = (g_1(t) + w1(t) * g_2^*(t)) * s(t) + (g_2(t) + w1(t) * g_1^*(t)) * s^*(t)$

where $(g_2(t) + w1(t) * g_1^*(t)) * s^*(t)$ is an image interference term.

$$W1_{OPT}(f) = -\frac{G_2(f)}{G_1^*(-f)}$$

$W1_{OPT}(f)$ holds when properness holds:

$c_y(\tau) = E\{y(t)y(t-\tau)\} = 0,\ for\ 0 \le \tau \le \tau_{max}$

Filtering:

$y(t) = z(t) + w1(t) * z^*(t)$

Adaptation:

$w1(n+1) = w1(n) - \alpha[y(t)\ y(t-1)\ \cdots\ y(t-N+1)]^T y(t)$

The time-domain adaptive filter architecture increases hardware complexity and power consumption because IQ compensation requires a high resolution of parameter updates for stability. The adaptive filter requires high resolution complex multiplier both at adaptation and filtering stages.

3. Low-complexity IQ compensator

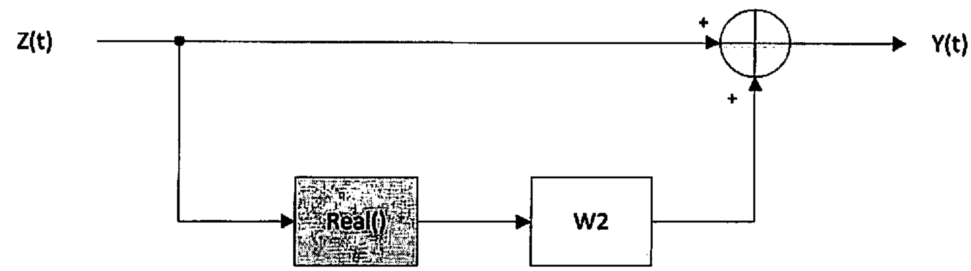


Figure 4: Low-complexity IQ compensator

The present system uses a real input of a complex signal for IQ imbalance compensation, according to Figure 4, instead of a conjugate input as in Figure 2. Consequently, IQ compensation can be implemented by two real multipliers for each tap multiplication instead of complex multipliers.

The present system provides hardware complexity of the filtering stage as half of the complexity of the filtering stage in the adaptive filter model. Due to reduced hardware complexity, the present system consumes lower power per each filtering operation. In spite of the reduced hardware complexity, the present system shows almost equivalent performance (or no loss) compared with the adaptive filter model.

The present system reduces complex filtering of frequency dependent IQ compensation real-only filtering with complex filter coefficients as in Figure 4. Correspondingly, total hardware complexity in IQ compensation stage may be reduced to half of the adaptive filter model.

To determine whether the optimal weight exists in the present system, the optimal weight $W2_{OPT}(f)$ in Figure 4 may be found from the following derivation:

$$y(t) = z(t) + w2(t) * real\{z(t)\} = z(t) + w2(t) * \frac{(z^*(t) + z(t))}{2} \qquad (3.1)$$

$$y(t) = \left(g_1(t) + \frac{1}{2} w2(t) * (g_2^*(t) + g_1(t))\right) * s(t) + \left(g_2(t) + \frac{1}{2} w2(t) * (g_1^*(t) + g_2(t))\right) * s \qquad (3.2)$$

$$W2_{OPT}(f) = -\frac{2G_2(f)}{G_1^*(-f) + G_2(f)} \qquad (3.3)$$

(3.2) may be derived from substitution of $z(t)$ in (1.4). (3.3) may be derived from Fourier transform of (3.2) and let $(G_{\downarrow}2\ (f) + 1/2\ w2(t) * (G_{\downarrow}1^{\dagger} *\ (-f) + G_{\downarrow}2\ (f))$ be zero. Since $W2$ is a complex filter coefficient and (3.3) is a realizable transfer function by FIR architecture, there exists a $W2$ that satisfies (3.3).

If (3.3) is compared with (2.3), the optimal weight for the present system estimates an average of $G_1^*(-f)$ and $G_2(f)$ instead of $G_1^*(-f)$ in the denominator term. Usually, $G_2(f)$ is much weaker than $G_1^*(-f)$ (at least 20dB) and, therefore, $W2_{OPT}(f)$ is approximately equal to $2W1_{OPT}(f)$.

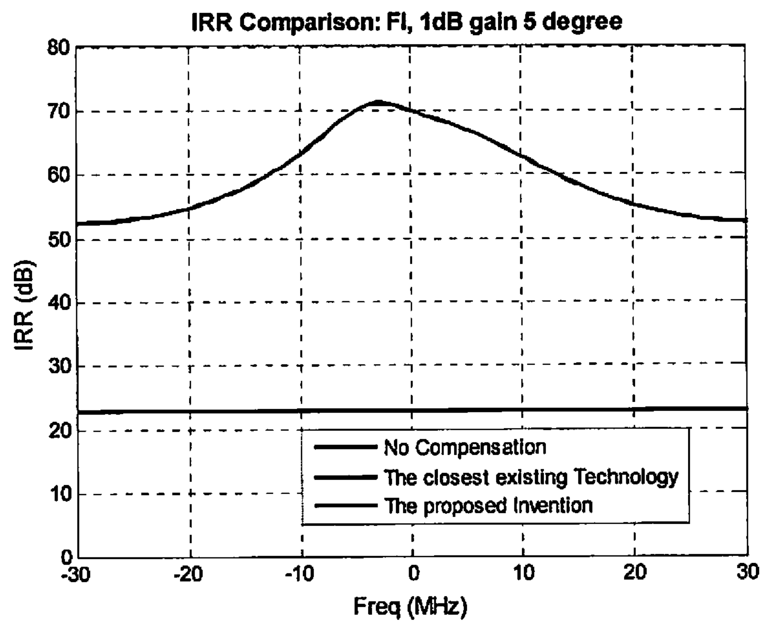
One thing to note for the present system is that using a real part of an input signal does not lose any information needed for IQ imbalance compensation compared with that of using a conjugate input. The main idea of FD IQC is to cancel image part $s^*(t)$ by adding compensating factor that contains a scalable $s^*(t)$ term. For the present system, a compensating factor to be added is $w2(t) * \frac{(z^*(t) + z(t))}{2}$ and it may be expressed as:

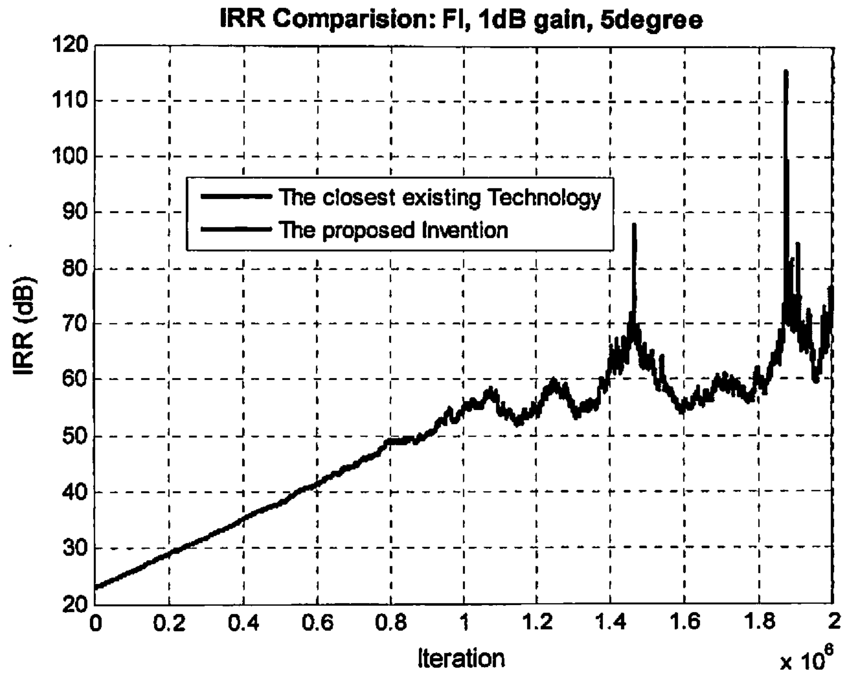
$$w2(t) * \frac{(z^*(t) + z(t))}{2} = \frac{1}{2} w2(t) * \left(g_2^*(t) + g_1(t)\right) * s(t) + \frac{1}{2} w2(t) * \left(g_1^*(t) + g_2(t)\right) * s^*(t) \quad (3.4)$$

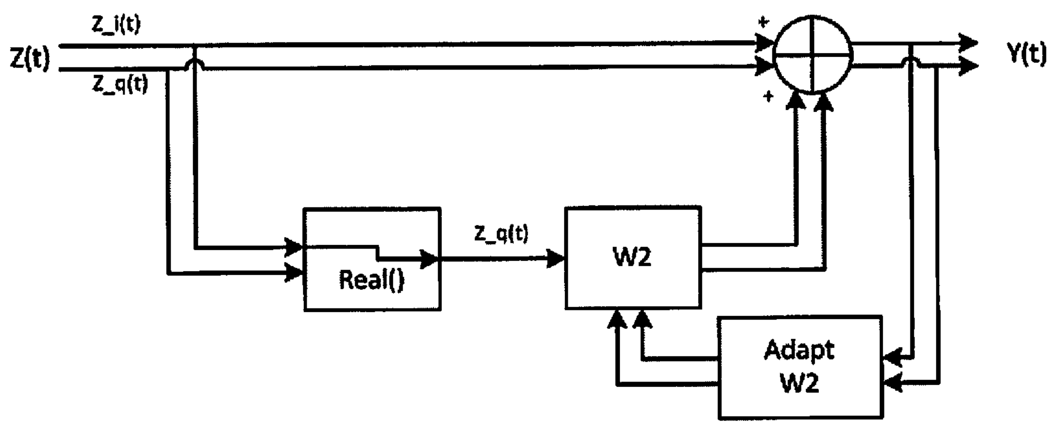
Therefore, the compensating factor in the present system contains $s^*(t)$ term that may cancel out image term in $z(t)$ by adjusting filter coefficients of $w2(t)$.

For adaptation, the present system may use the same approximate newton update. The approximation in (2.5) forces scaling factor of the exact newton method to be a constant value. Since the only difference between the present system and the adaptive filter model in the exact newton update is a scaling term, the approximate newton update in the present system becomes the same as (2.5).

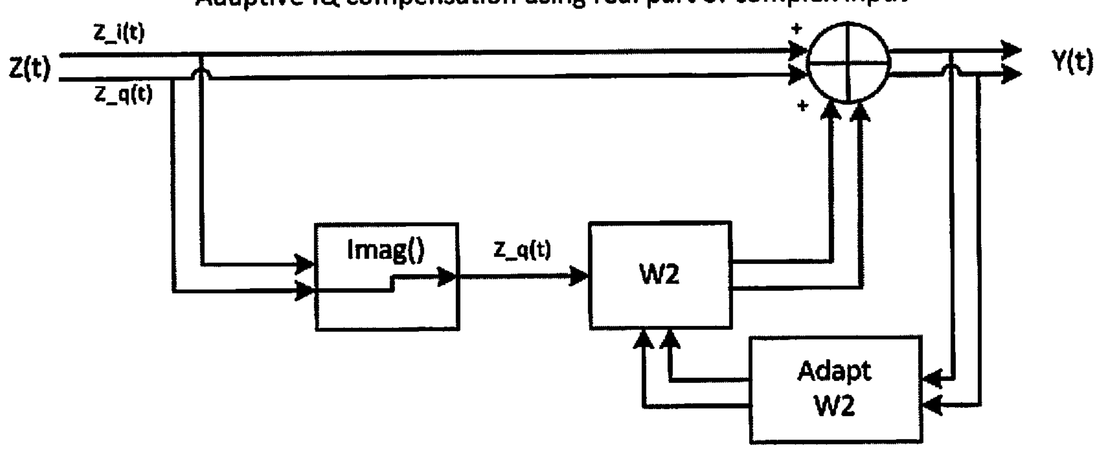
MATLAB evaluations that compare the present system and the adaptive filter model in FI environments are presented below. For FI environment, 1dB gain mismatch and 5 degree phase mismatch are used. The image rejection ratio (IRR) of the present system almost coincides with the adaptive filter model.







Adaptive IQ compensation using real part of complex input



Adaptive IQ compensation using imaginary part of complex input

The present system reduces the complex multiplication of frequency dependent IQ compensation to real multiplication by using a real part or an imaginary part of the input signal. The complexity in the filtering stage may be reduced by at least half of the adaptive filter. The optimality check (when real part of input is used) may be derived as follows:

$$y(t) = z(t) + w2(t) * real\{z(t)\} = z(t) + w2(t) * \frac{(z^*(t) + z(t))}{2}$$

$$y(t) = \left(g_1(t) + \frac{1}{2}w2(t) * (g_2^*(t) + g_1(t))\right) * s(t) + \left(g_2(t) + \frac{1}{2}w2(t) * (g_1^*(t) + g_2(t))\right) * s^*(t)$$

$$W2_{OPT}(f) = -\frac{2G_2(f)}{G_1^*(-f) + G_2(f)}$$

$W2_{OPT}(f)$ is about two times of $W1_{OPT}(f)$, considering $G_2(f)$ is much smaller than $G_1(f)$. The present system may achieve $W2_{OPT}(f)$ when properness holds, which is the same as an adaptive filter's optimality condition.

The present system does not tradeoff between complexity and performance. The performance of the present system does not degrade in spite of using only a real part or an imaginary part of input is that the compensating factor still contains a scalable $s^*(t)$ that cancels an image part of IQ mismatch input.

For a real input filtering case, the filter output becomes:

$$w2(t) * \frac{(z^*(t) + z(t))}{2} = \frac{1}{2}w2(t) * \left(g_2^*(t) + g_1(t)\right) * s(t) + \frac{1}{2}w2(t) * \left(g_1^*(t) + g_2(t)\right) * s^*(t)$$

$s^*(t)$ may be scalable by $w2(t)$ without loss of performance.