PAPER

Blind Adaptive Compensation for Gain/Phase Imbalance and DC Offset in Quadrature Demodulator with Power Measurement

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SUMMARY In this paper we propose a new blind adaptive compensator associated with the inverse QRD-RLS (IQRD-RLS) algorithm to adaptively estimate the parameters, related to the effects of gain/phase imbalance and DC offsets occur in the Quadrature demodulator, for compensation. In this new approach the power measurement of the received signal is employed to develop the blind adaptation algorithm for compensator, it does not require any reference signal transmitted from the transmitter and possess the fast convergence rate and better numerical stability. To verify the great improvement, in terms of reducing the effects of the imbalance and offset, over existing techniques computer simulation is carried out for the coherent 16 PSK-communication system. We show that the proposed blind scheme has rapidly convergence rate and the smaller mean square error in steady state.

key words: *quadrature modulation, gain/phase imbalance, DC offset, compensation, inverse QRD-RLS algorithm*

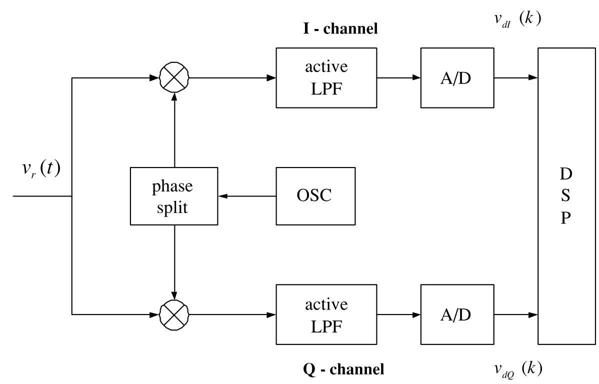

Fig. 1 The block diagram of quadrature demodulator.

1. Introduction

Traditional wireless system design has been carried out at somewhat disjointed levels of abstraction: RF system experts plan the transceiver architecture; IC designers develop each of the building blocks; communication theory people derive their base band technologies with a simple channel model. Recently, there has been much effort in new design for transceiver used in mobile communication. One of the approaches is to combine the RF functions with DSP, which will allow us to employ linear modulation techniques and permit flexibility of modulation format and receiver processing. It is known that the quadrature modulation technique has been widely used in the transceiver of modern communication systems, and, in practice, there is always some Gain/Phase imbalance and DC offset between the I and Q channels of modulator and demodulator. This is mainly due to finite tolerances of capacitor and resistor values used to implement the analog components [1], [2]. The unavoidable imbalance between the I and Q channels is known to degrade the performance of quadrature communication systems [3]. To compensate the problem due to the imbalance between I and Q channels, many techniques have been proposed for the transmitter as well as receiver [3]–[9]. In the transmitter the so-called direct-conversion structure [3], [10], [12] is the most commonly used for compensation. While in the receiver both structures, the direct-conversion receiver [3], [4] and low-IF receiver [5]–[9], have been employed. In this paper, we will focus on the design compensation in the receiver, the approach with the direct-conversion receiver for imbalance and offset compensation is considered.

The block diagram of a typical quadrature demodulator is depicted in Fig. 1, where the phase splitter at the local oscillator does not normally produce an exact $\pi/2$ separation, resulting in the phase imbalance between the I and Q channels. In fact, the mixers are never perfectly balance, causing gain imbalance between the two channels. Additionally, the gain imbalance and DC offset in each channel can be introduced into the systems by the active low pass filter and A/D converter. To compensate the effect due to the imbalance between I and Q channels, in [3] the approach with the decision directed structure was proposed to compensate the imbalance and offset. Like an adaptive equalizer, the compensator can operated in reference directed or decision directed mode. However, this method has problems of slow convergence and having relatively large steady-state values of mean square error in case of low signal to noise ratio (SNR).

In [4] two-steps approach, referred to as the linear parameters estimation and compensation scheme, has been proposed to circumvent the problems described above. It first estimates the parameters, related to the effect of imbalance and offset, and after having obtained the estimated parameters the correction of the received data can be performed. By this approach, the reference signal sent from the transmitter is required and must be constant envelope, this is different from the one proposed in [3]. That is, in [3] a direct decision in conjunction with the compensator

Manuscript received March 7, 2003.
Manuscript revised September 29, 2003.
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was employed. To further improve the compensator, in [4] an alternative approach, referred to as the nonlinear parameters estimation and compensation technique, where the reference signals are not required in the receiver, has been developed. This method estimates the parameters related to the imbalance and offset only with the received demodulated I/Q components. But since I/Q components of received signal are noisy, the biased estimation might occur for low signal to noise ratio.

In this paper, a new blind compensator associated with the inverse QRD-RLS (IQRD-RLS) algorithm is devised to adaptively estimate the nonlinear parameters and to compensate the effects of the gain/phase imbalance and DC offsets occur in the quadrature demodulator. Where the power measurement of the received signal of the receiver is employed with the blind adaptation algorithm for compensation, it does not require any reference signal transmitted from the transmitter (without using the training signal). Moreover, the IQRD-RLS algorithm has been shown to have the fast convergence rate and having better numerical stability among the RLS family. Therefore, with our proposed scheme the great improvement for eliminating the effects of the imbalance and offset can be achieved over the existing techniques.

For discussion, in this paper we first introduce the signal model of compensator in Sect. 2. In Sect. 3, the basic idea behind the proposed new blind scheme is addressed and in Sect. 4 computer simulation results are given to verify the superiority of the proposed scheme Finally, in Sect. 5 a brief conclusion is given.

2. Signal Model Description

Let us consider an equivalent baseband communication model where data signal is transmitted with quadrature modulator, i.e.,

$$v_a(t) = A(t)\{\cos\theta(t) + j\sin\theta(t)\} \tag{1}$$

$A(t)$ denotes the amplitude of the transmitted waveform and $\theta(t)$ is the phase, containing the information of I and Q channels. For convenience, the effect of the imbalance and offset at modulator is neglected. The received signal $v_r(t)$ can be expressed as

$$\begin{aligned} v_r(t) &= A(t)\{\cos\theta(t) + j\sin\theta(t)\} + n_I(t) + jn_Q(t) \\ &= v_{rI}(t) + jv_{rQ}(t) \end{aligned} \tag{2}$$

Where $n_I(t)$ and $n_Q(t)$ are the in-phase and quadrature components of the narrow band noise, respectively. Due to the effect of the imbalance and offset, the components of I and Q channels of the demodulated signal are represented by

$$\begin{aligned} v_{dI}(k) &= \alpha v_{rI}(k) + C_I \\ v_{dQ}(k) &= v_{rQ}(k)\cos\phi + v_{rI}(k)\sin\phi + C_Q \end{aligned} \tag{3}$$

In (3) α and ϕ represents the gain and phase imbalance between the I and Q channels, while C_I and C_Q are DC offsets of the I and Q channels, respectively.

In [3], [4] many conventional compensation techniques were proposed by assuming that the signal of the demodulator input is the transmitted signal, $v_a(t)$, of (1). However, in practice, in the receiver $v_a(t)$ is not available. In Fig. 1, the function of DSP is modeled to deal with the effects of imbalance and offset introduced in the quadrature modulator in which the parameters of the DSP model are related to the effects of imbalance and offset of the demodulator and have to be estimated. It seems reasonable that we may simply take the reference signal directly from the input of demodulator instead of using the training sequence sent from the transmitter. Moreover, since the power measurement of the received signal of the demodulator is feasible from the receiver. With this basic idea, the new blind adaptive compensation scheme can be developed in what follows.

3. Blind Adaptive Compensator

In order to circumvent the disadvantage of the compensators described in [3], [4], in this section a new blind scheme with power measurement of the received signal in the demodulator is proposed. It is an alternative indirect non-linear parameters estimation and compensation approach, and can be viewed as a blind version of the parameter estimation and compensation (PEC) where an adaptive IQRD-RLS algorithm is employed for updating the parameters, which are the non-linear function of the terms corresponding to imbalance and offset.

For discussion, the block diagram of the new blind adaptive compensator is illustrated in Fig. 2. The received signal $v_r(t)$ is applied to the demodulator, and in the output we obtain the quadrature-demodulated signals $v_{dI}(k)$, $v_{dQ}(k)$, as defined in (3). Based on the quadrature-demodulated signals of (3) and the power measurement taken from the input of receiver, $v_r(t)$, denoted as $p(n)$, we are able to devise a new blind scheme to adaptively estimate the parameters related to the imbalance and offset. In consequence, the estimated parameters can be applied to the compensator for compensation. By definition the power measurements at the demodulator input with a gain of G, can be expressed as

$$p(k) = G(v_{rI}^2(k) + v_{rQ}^2(k)) + \Delta(k) \tag{4}$$

$\Delta(k)$ is denoted as the measurement noise, which consists of the thermal noise in the circuit and the quantization noise due to the sampling process. Generally, the thermal noise is

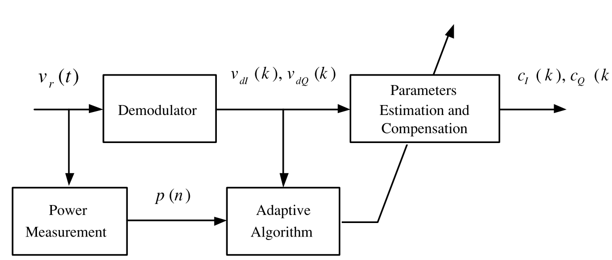


Fig. 2 The block diagram of new blind adaptive compensator.

modeled as a white Gaussian process, while the quantization noise is modeled as a white noise with uniform distribution [10]. From (3) and after some mathematical manipulation, we have

$$v_{rI}(k) = \frac{v_{dI}(t) - C_I}{\alpha}$$
$$v_{rQ}(k) = \frac{\alpha(v_{dQ}(k) - C_Q) - (v_{dI}(k) - C_I)\sin\phi}{\alpha\cos\phi} \tag{5}$$

Substituting (5) into (4), the sample of the power measurement data at time index k is expressed as

$$p(n) = \mathbf{w}^T\mathbf{u}(n) + \Delta(n) \tag{6}$$

In (6) $\mathbf{u}(n)$ and $\mathbf{w}$ are denoted as

$$\mathbf{u}(n) = \begin{bmatrix} v_{dI}^2(n) \\ v_{dI}(n) \\ v_{dI}(n)v_{dQ}(n) \\ v_{dQ}(n) \\ v_{dQ}^2(n) \\ 1 \end{bmatrix} \tag{7}$$

and

$$\mathbf{w} = \begin{bmatrix} w_1 \\ w_2 \\ w_3 \\ w_4 \\ w_5 \\ w_6 \end{bmatrix} = \frac{G}{\alpha^2(1-s^2)} \begin{bmatrix} 1 \\ 2\alpha s C_Q - 2C_I \\ -2\alpha s \\ 2\alpha s C_I - 2\alpha^2 C_Q \\ \alpha^2 \\ C_I^2 + \alpha^2 C_Q^2 - 2\alpha s C_I C_Q \end{bmatrix}, \tag{8}$$

respected, and in (8) s =sinϕ. Now, based on (6)–(8), the problem becomes to estimate the unknown parameter vector, $\mathbf{w}$, given two observable sets $\mathbf{u}(n)$ and $p(n)$, where $p(n)$ can be viewed as the desired response. It should be noted that the function of compensator is an inverse filtering of the demodulator, with gain/phase imbalance and DC offset. Therefore, with the power measurement (6) can be viewed as the problem of system identification, in which $\mathbf{u}(n)$ is the input vector, from (3) we know that it is related to the I and Q channels of the demodulated signals, $v_{dI}(n)$ and $v_{dQ}(n)$. While $p(n)$ is the desired response,which is related to the I and Q channels signal components, $v_{rI}(n)$ and $v_{rQ}(n)$, of the received signal $v_r(n)$, it consists of the modulated data signal and background noise, or other interference signal, e.g., if Rayleigh fading channel is considered. Also, in (6) $\mathbf{w}(n)$ represents the system parameter vector to be estimated, which is related to the parameters of the gain/phase imbalance and DC offset of the demodulator, and $\Delta(n)$ is the measurement error. As described above, the major difference of the proposed scheme from the conventional schemes addressed in [3], [4] is placed on the fact that in our scheme the desired response is directly related to the received signal $v_r(n)$, it is changed with the environment we faced, and will have less effect due to noise or other interferences. On the other hand, in the conventional schemes, they assumed that the desired response is the training data signal or using the ideal error free components of $v_{rI}(n)$ and $v_{rQ}(n)$. Therefore, the mismatch between input and desired response signals of the compensator will be increased when the noise or interference component is increased. Next, for estimating the parameter vector, based on (6), we may define the estimation error as

$$e(n) = p(n) - \mathbf{w}^T(n)\mathbf{u}(n) \tag{9}$$

The parameters defined in (8) can be obtained by minimizing the exponentially weighted least squares error, i.e.,

$$\varepsilon(n) = \sum_{i=1}^{n} \lambda^{n-i}|e(n)|^2 \tag{10a}$$

where λ is the forgetting factor. In a form of matrix norm (10a) may be rewritten as

$$\begin{aligned}\varepsilon(n) &= ||\Lambda^{1/2}(n)\mathbf{e}(n)||^2 \\ &= ||\Lambda^{1/2}(n)\mathbf{p}(n) - \Lambda^{1/2}(n)\mathbf{X}(n)\mathbf{w}(n)||^2\end{aligned} \tag{10b}$$

where $\mathbf{e}(n) = [e(1), e(2), \ldots, e(n)]^T$ and $\mathbf{p}(n) = [p(1), p(2), \ldots, p(n)]^T$ are the error vector and desired response vector respectively. And $||(.)||$ denotes the Euclidean norm of (.). The diagonal matrix and the data matrix are designated by $\mathbf{\Lambda}^{1/2}$=diag[$\sqrt{\lambda^{k-1}}$, $\sqrt{\lambda^{k-2}}, \ldots, \sqrt{\lambda}$, 1] and $\mathbf{X}(n) = [\mathbf{u}(1), \mathbf{u}(2), \ldots, \mathbf{u}(n-1), \mathbf{u}(n)]^T$, respected. Recalled that in the conventional QRD-RLS algorithm an orthogonal matrix $\mathbf{Q}(n)$ is used to triangular the data matrix, $\Lambda^{1/2}\mathbf{X}(n)$, and the weighted desired vector, $\Lambda^{1/2}\mathbf{p}(n)$, by Givens rotation [11]. That is,

$$\mathbf{Q}(n)\Lambda^{1/2}(n)\mathbf{X}(n) = \begin{bmatrix} \mathbf{R}(n) \\ \mathbf{O} \end{bmatrix} \tag{11}$$

$$\mathbf{Q}(n)\Lambda^{1/2}(n)\mathbf{p}(n) = \begin{bmatrix} \mathbf{z}(n) \\ \mathbf{v}(n) \end{bmatrix} \tag{12}$$

Where $\mathbf{R}(n)$ is a 6×6 upper triangular matrix, $\mathbf{O}$ is an (n-6)× 6 zero matrix, $\mathbf{z}(n)$ and $\mathbf{v}(n)$ are the 6×1 and (n-6)×1 vectors. Since orthogonal matrices are length preserving, using the results of (11) and (12), the cost function of (10) can be rewritten as

$$\xi(n) = \left|\left|\left| \begin{bmatrix} \mathbf{z}(n) - \mathbf{R}(n)\mathbf{w}(n) \\ \mathbf{v}(n) \end{bmatrix} \right|\right|\right|^2 \tag{13}$$

By minimizing (13), the optimal weight vector of the least square (LS) solution can be obtained

$$\mathbf{w}_{LS}(n) = \mathbf{R}^{-1}(n)\mathbf{z}(n) \tag{14}$$

(14) is referred to as the QRD-LS algorithm. It can be solved by back substitution to solve the LS weight vector. Moreover, in [11] an alternative approach referred to as the inverse QRD-RLS (IQRD-RLS) algorithm was proposed. It can be used to solve the LS weight vector without implementing the highly serial back substitution step required in the QRD-RLS algorithm. From [11], the IQRD-RLS algorithm for updating the LS weight vector is given by

$$\mathbf{w}(n) = \mathbf{w}(n-1) + \frac{\mathbf{u}_g(n)}{b(n)} e(n, n-1) \tag{15a}$$

where the priori estimation error, $e(n, n-1)$, is defined as

$$e(n, n-1) = p(n) - \mathbf{w}^T(n-1)\mathbf{u}(n) \tag{15b}$$

Both the scalar variable $b(n)$ and the vector $\mathbf{u}_g(n)$ of (15a) are evaluated entirely by rotation-based method, using the *Givens rotations*, when $\mathbf{R}^{-T}(n)$(*inverse Cholesky factor*) is updated from $\mathbf{R}^{-T}(n-1)$. To do so, we define an intermediate vector, $\mathbf{a}(n)$, i.e.,

$$\mathbf{a}(n) = \lambda^{-1/2}\mathbf{R}^{-T}(n)\mathbf{u}(n) \tag{16a}$$

It provides the key to the parallelization of the inverse QRD-RLS approach. From [11], it shows that there exists rotation matrix $\mathbf{P}(n)$ such that

$$\begin{bmatrix} \mathbf{0} \\ b(n) \end{bmatrix} = \mathbf{p}(n) \begin{bmatrix} \mathbf{a}(n) \\ \mathbf{1} \end{bmatrix} \tag{16b}$$

where $\mathbf{P}(n)$ successively annihilates the elements of the vector $\mathbf{a}(n)$, starting from the top, by rotating them into the element at the bottom of the augmented vector $[\mathbf{a}^T(n)\ 1]^T$. In consequence, we can evaluate $\mathbf{u}_g(n)$ by

$$\begin{bmatrix} \mathbf{R}^{-T}(n) \\ \mathbf{u}_g^T(n) \end{bmatrix} = \mathbf{p}(n) \begin{bmatrix} \lambda^{-1/2}\mathbf{R}^{-T}(n-1) \\ \mathbf{0}^T \end{bmatrix} \tag{17}$$

The IQRD-RLS algorithm of (15a) requires the initial setting for parameters $\mathbf{R}^{-T}(n)$ and $\mathbf{w}(n)$ described as follows: Initialize the algorithm by setting

$$\mathbf{R}^{-T}(0) = \delta^{-1}\mathbf{I}$$
$$\mathbf{w}(0) = \mathbf{0}$$

Where δ is a small positive constant and $\mathbf{I}$ is a 6×6 identity matrix. After having obtained the parameter vector $\mathbf{w}(n)$, the estimate of the gain, phase imbalance and DC offset can be obtained

$$\begin{aligned} \alpha &= \sqrt{\frac{w_5}{w_1}} \\ \phi &= \sin^{-1}\left(-\frac{w_3}{2\sqrt{w_1 w_5}}\right) \\ C_I &= \frac{2w_2 w_5 - w_3 w_4}{w_3^2 - 4w_1 w_5} \\ C_Q &= \frac{2w_1 w_4 - w_2 w_3}{w_3^2 - 4w_1 w_5} \end{aligned} \tag{18}$$

In consequence, the estimated parameters are employed for the correction of the received signal to perform the compensation. That is, the correction of the imbalance and offset is carrier out using (3) by replacing $v_{rI}(n)$ and $v_{rQ}(n)$ with $c_I(n)$ and $c_Q(n)$, respectively.

$$\begin{aligned} c_I(n) &= \frac{v_{dI}(n) - C_I}{\alpha} \\ c_Q(n) &= \frac{\alpha(v_{dQ}(n) - C_Q) - (v_{dI}(n) - C_I)\sin\phi}{\alpha\cos\phi} \end{aligned} \tag{19}$$

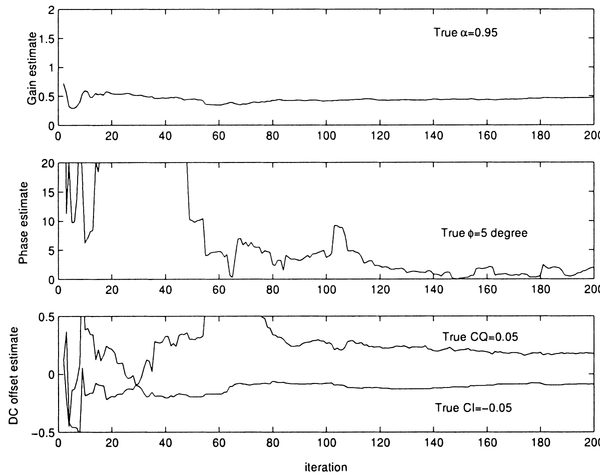
4. Computer Simulation Results

To document the merits of the proposed scheme, in this section, computer simulation is carried out for a coherent 16-PSK-communication system. The simulation results are compared with other existing techniques, such as Decision directed compensation (DDC) [3], the DDC with IQRD-RLS algorithm, the Linear and nonlinear parameter estimation and compensation (PEC) [4]. In the simulation we assume that the transmitter is free of any gain/phase imbalance and DC offset. Also, in the simulation, the following parameters with values; $\alpha = 0.95$, $\phi = 5°$, $C_I = -0.05$, $C_Q = 0.05$, are employed.

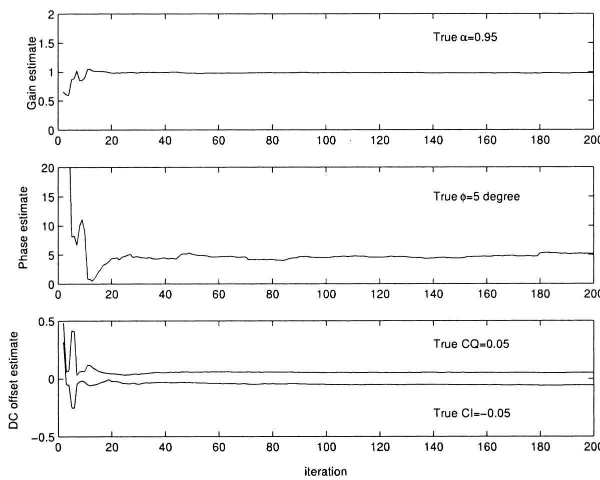
As described earlier in Sect. 3, that with our proposed scheme the component of background noise has been involved in the I and Q channels of the demodulated signals, $v_{dI}(n)$ and $v_{dQ}(n)$, as the input signal in the compensator as well as in the desired response of $p(n)$. Therefore, with the faster convergence IQRD-RLS algorithm, the proposed scheme could achieve faster convergence rate to eliminate the effect due to the gain/phase imbalance and DC offset of the demodulator as evident from Fig. 3 to Fig. 5. Since the PEC-nonlinear scheme proposed in [4] was with the blind approach, first we would like to compare the convergence property of the proposed scheme with the PEC-nonlinear scheme, in terms of gain/phase imbalance and DC offset. Figure 3 illustrates the convergence behavior of the parameter estimates with the nonlinear parameters estimation scheme suggested in [4] and the proposed scheme, for SNR=0 dB. As observed from Fig. 3, all the parameters estimated by the proposed scheme with the IQRD-RLS algorithm converge to the desired values in about 30 symbols. However, this is not the case when the nonlinear estimation and compensation suggested in [4] is employed, that is, the estimation is biased due to the low signal to noise ratio. To further investigate the advantage of the proposed scheme, the learning curve, e.g., the convergence behavior of the mean square error, is considered. For comparison, we define the *symbol error* as the remainder in the symbol between the input of the demodulator and the output of the compensator, i.e.,

$$e_{symbol}(n) = [r_I(n) - C_I(n)] + j[r_Q(n) - C_Q(n)] \tag{20}$$

Fig. 4 is the comparison of the learning curves with different approaches for SNR=15 dB. In general, in the conventional engineering problems the approach with the blind adaptation scheme has worse performance compared with the non-blind adaptation approach. However, from Fig. 4, we learn that the proposed PEC performed the best and having faster convergence rate. Similarly, in Fig. 5 for SNR=5 dB the proposed PEC scheme outperforms other existing methods. As described earlier with the proposed scheme the component of background noise has been involved in the system identification model, where the I and Q channels of the demodulated signals, $v_{dI}(n)$ and $v_{dQ}(n)$, can be viewed as the input signal in the compensator and the desired response,



(a)



(b)

Fig. 3 Convergence of parameter estimation with (a) nonlinear parameters estimation [4] and (b) proposed technique in SNR=0 dB.

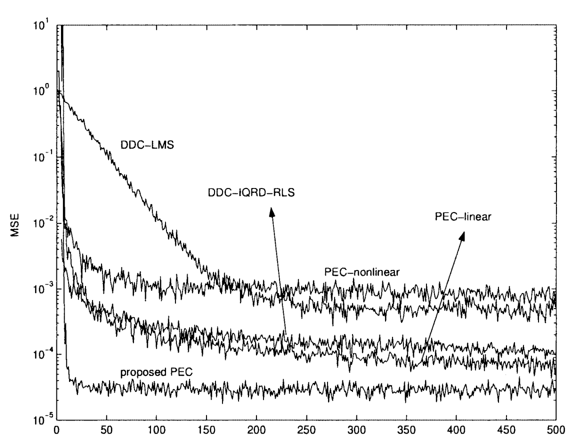


Fig. 4 Learning curve with different schemes for SNR=15 dB.

$p(n)$ is the corresponding output signal. It has the less effect by the background noise as evident shown in Fig. 4 and Fig. 5. Moreover, sine the DDC-LMS and DDC-IQRD-RLS

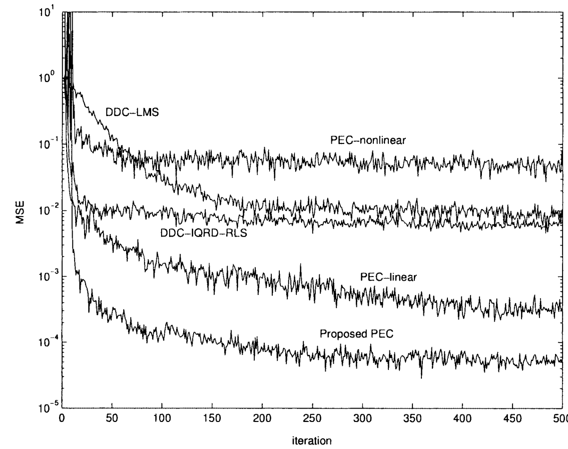


Fig. 5 Learning curve with different schemes for SNR=5 dB.

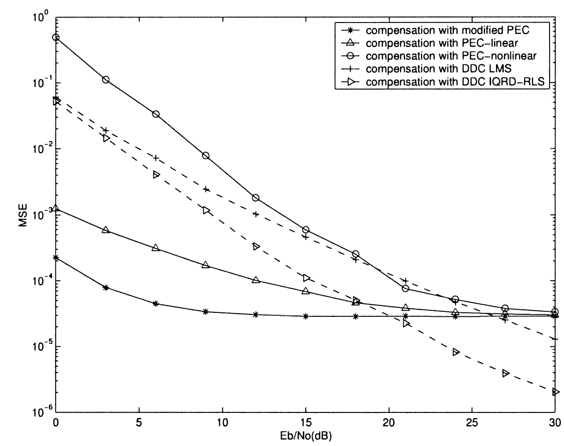


Fig. 6 Comparison of the typical mean square symbol error with different compensation techniques.

are with the training data in the output of compensator, when the value of SNR is changed from 15 dB to 5 dB, the performance are degraded, dramatically, compared with the one of using the PEC-linear. In fact, it can be easily observed from the optimal weight vectors derived for the DDC-LS and PEC-linear with LS approach. The DDC-LS is affected by the noise more than the PEC-linear. (Due to the limitation of content, the detail analysis is not given here.) Also, in all case the conventional blind approach, the PEC-nonlinear scheme performed the worse. Next, the typical mean square symbol error for various compensation techniques are examined. Again, as depicted in Fig. 6, the proposed scheme has the smaller value of mean square symbol error, when the value of SNR is below 20 dB. Since for higher SNR cases, all the methods used for adaptive compensatorcould perform well, it is not useful in practical application. Finally, we would like to show the performance in terms of bit-error-rate (BER). From Fig. 7, by using the proposed PEC scheme the BER approaches the one (the ideal case) without having the gain/phase imbalance and DC offset. That is, with our approach we are able to eliminate the effect of due to the imbalance and DC offset in the quadrature demodulator,

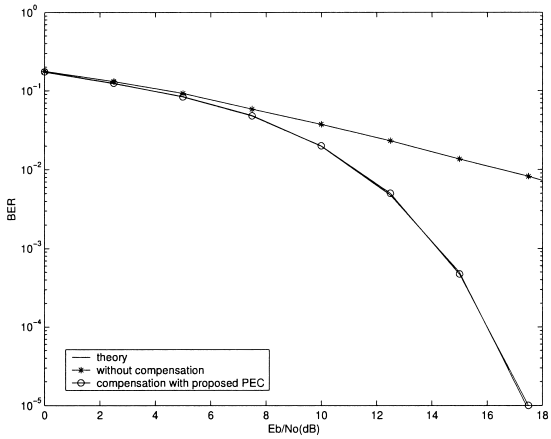


Fig. 7 Performance comparison of BER for the cases with and without compensation.

perfectly.

Next, it is of interest to investigate the performance of the proposed scheme when it is fluctuated by the multipath fading channel. From (2), when the signals transmitted through the multipath-fading channel, the received signal can be rewritten as

$$\begin{aligned} v_r(t) &= \alpha_1(t)e^{j\phi_1(t)}A_1(t)[\cos\theta(t) + j\sin\theta(t)] \\ &\quad + \sum_{l=2}^{L}\{\alpha_l(t)e^{j\phi_l(t)}A_l(t)[\cos\theta(t-\tau_l) \\ &\quad + j\sin\theta(t-\tau_l)]\} + n_I(t) + jn_Q(t) \\ &= A_1(t)[\cos\theta(t) + j\sin\theta(t)] \\ &\quad + [\alpha_1(t)e^{j\phi_1(t)} - 1]A_1(t)[\cos\theta(t) + j\sin\theta(t)] \\ &\quad + \sum_{l=2}^{L}\{\alpha_l(t)e^{j\phi_l(t)}A_l(t)[\cos\theta(t-\tau_l) \\ &\quad + j\sin\theta(t-\tau_l)]\} + n_I(t) + jn_Q(t) \\ &= A_1(t)[\cos\theta(t) + j\sin\theta(t)] + N_I(t) + jN_Q(t) \end{aligned} \tag{21}$$

Where α_l and ϕ_l denote the Rayleigh fading parameter and Doppler frequency shift, respectively. It is noted that in this case the components of narrowband noise, $n_I(t)$ and $n_Q(t)$, described in (2) have been replaced by $N_I(t)$ and $N_Q(t)$, respectively, to include the effect of multipath fading environment. To investigate this effect for the proposed scheme, a synchronous pilot symbol-aided QPSK DS-CDMA system over multipath fading channel addressed in [13] is considered. For simplicity, only single user with 5 multipaths is focused. The channel bandwidth is 3.968 MHz, the carrier frequency is 2.0 GHz (consistent with that given in [13]), and the mobile speed is set as 50 km/h. Here the non-orthogonal Gold codes with length N=31 are employed. As evident from Fig. 8 and Fig. 9, the proposed scheme is more robust to the multipath fading and has better convergence property as compared to other conventional techniques, in terms of the convergence curves of gain/phase imbalance and DC offset parameter estimation, and MSE. Other conventional schemes have serious effect of phase imbalance due to multipath fading environment, but the proposed scheme still performed well.

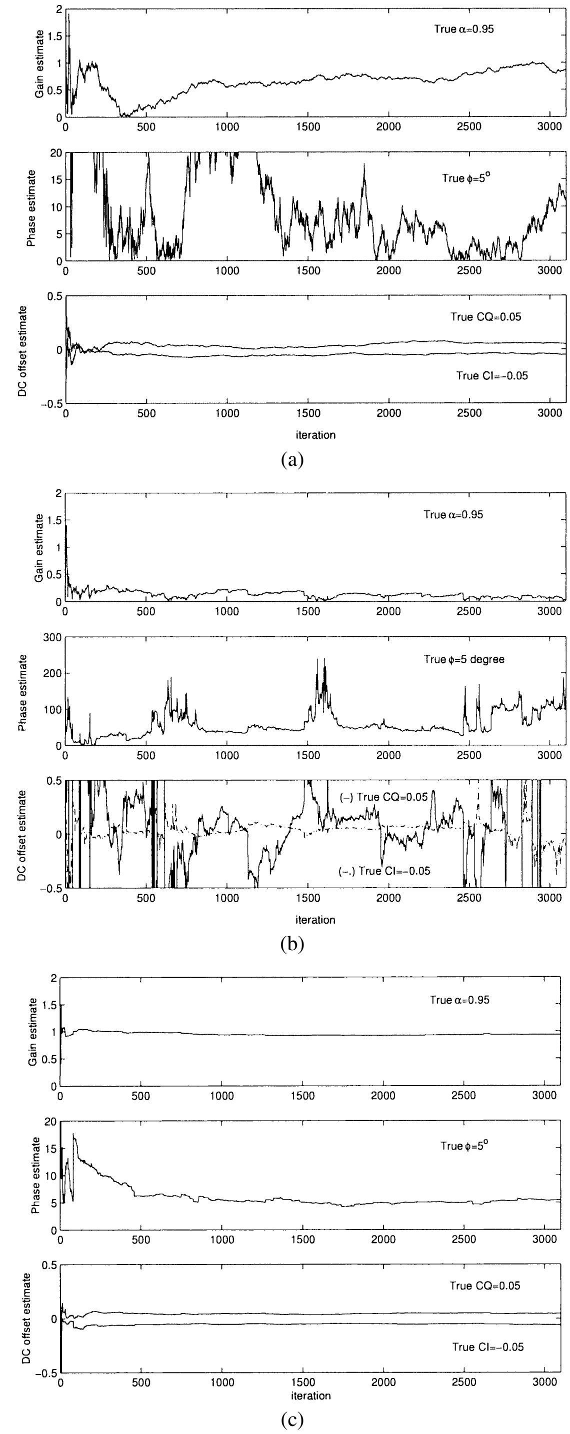


Fig. 8 Convergence of parameter estimation with α=0.95, f=5°, C_I = −0.05 and C_Q=0.05. (a) Linear parameters estimation [4] (b) nonlinear parameters estimation [4] and (c) proposed scheme, for SNR=0 dB.

Finally, to evaluate and analyze the computation complexity of the proposed scheme and compared with the con-

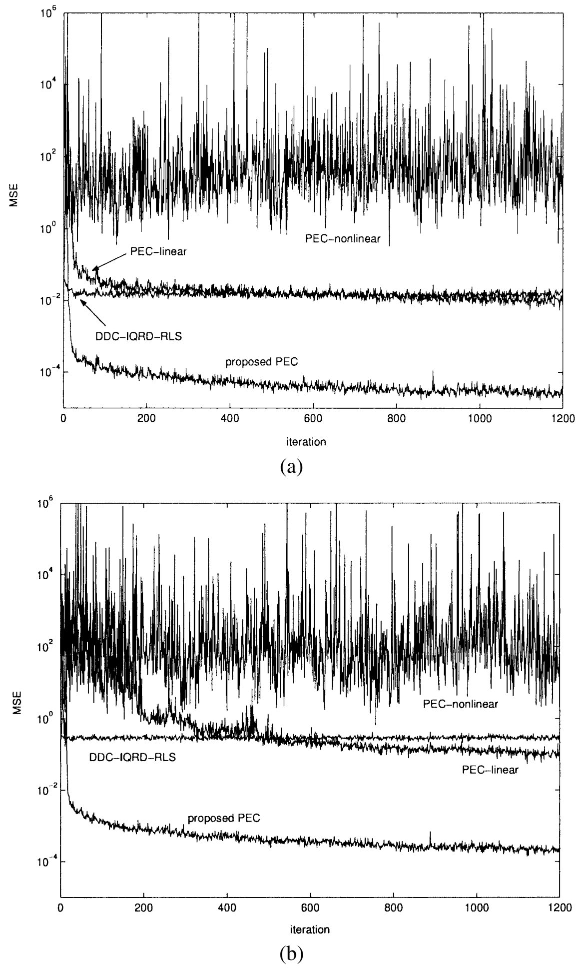


Fig. 9 Learning curves with different schemes for (a) SNR=15 dB (b) SNR=5 dB. (α=0.95, f=5°, C_I=−0.05 and C_Q=0.05)

Table 1 Comparison of computation complexity.

Algorithm		kalman gain	Update LS weights	Total complexity	Others
LMS	MUL		$2N$+1	$2N$+1	
	ADD		N+1	N+1	
RLS	MUL	$6N$	$2N^2$+$2N$+1	$2N^2$+$8N$+1	
	ADD		N^2+N+1	N^2+N+1	1 division
IQRD-RLS	MUL	$6N$	$2N$	$8N$	$O(N)$ operations for recursive updating of the triangular matrix $\mathbf{R}(n)$
	ADD		$2N$	$2N$	

ventional schemes, in Table 1 the computation complexity of the IQRD-RLS algorithm used in the proposed scheme with the conventional LMS and RLS algorithms is given for comparison. It is well known that the conventional LMS has the less computational requirement than the RLS family. Also, as discussed in [11], among the RLS family the IQRD-RLS algorithm can provide the same numerical stability and achieving the fast convergence rate as the QRD-RLS algorithm, and outperformed the conventional RLS algorithm. Moreover, the structure of the IQRD-RLS algorithm is more suitable for IC design and VLSI circuit implementation than the QRD-RLS algorithm. Since in this application, only six parameters are involved for estimation, the configuration scale of DSP will not be a problem.

5. Conclusions

In this paper, the problem of gain/phase imbalances and DC offsets in quadrature demodulator has been investigated. Here, a new blind adaptive compensation scheme along with the power measurement of the received signal was developed to adaptively estimate and compensate, using the IQRD-RLS algorithm, for the gain/phase imbalances and DC offsets in a quadrature demodulator. Simulation results verified that the effects of the gain/phase imbalance and DC offset could be eliminated with the proposed blind adaptive compensator along with the IQRD-RLS algorithm. It possessed the properties of having rapidly convergence rate and the smaller MSE compared with other existing techniques addressed in [3], [4]. Moreover, with this approach, it did not require any training sequence sent from the transmitter, resulting in data rate improvement. Moreover, the proposed new blind adaptive compensation scheme is very suitable for communication systems with QAM transceiver.

Acknowledgement

The finacial support of this work by the National Science Council, Taiwan under the contract no. NSC92-2213-E-110-014, is greatly acknowledged. Also, we would like to thank the anonymous reviewers for their useful comments to enhance the quality of this paper.

References

[1] A.A. Abidi, "Direct-conversion radio transceivers for digital communications," IEEE J. Solid-State Circuits, vol.30, no.12, pp.1399–1410, Dec. 1995.
[2] B. Razavi, "Design considerations for direct-conversion receivers," IEEE Trans. Circuits Syst. II, Analog Digit. Signal Process., vol.44, no.6, pp.428–435, June 1997.
[3] J.K. Cavers and M.W. Liao, "Adaptive compensation for imbalance and offset losses in direct conversion transceivers," IEEE Trans. Veh. Technol., vol.42, no.4, pp.581–588, Nov. 1993.
[4] H. Nguyen, "Improving QPSK demodulator performance for quadrature receiver with information from amplitude and phase imbalance correction," WCNC, Sept. 2000.
[5] J.P.F. Glas, "Digital I/Q imbalance compensation in a low-IF receiver," Proc. IEEE Globecom, pp.1461–1466, Nov. 1998.
[6] M. Valkama and M. Renfors, "Advanced DSP for I/Q imbalance compensation in a low-IF receiver," Proc. IEEE Int. Conf. Commun., pp.768–772, June 2000.
[7] M. Valkama, M. Renfors, and V. Koivunen, "Blind source separation based I/Q imbalance compensation," Proc. IEEE Symp. Adaptive Syst. Signal Process., Commun., Contr., pp.310–314, Oct. 2000.

[8] L. Yu and W.M. Snelgrove, "A novel adaptive mismatch cancellation system for quadrature IF radio receivers," IEEE Trans. Circuits Syst. II, Analog Digit. Signal Process., vol.46, no.6, pp.789–801, June 1999.
[9] M. Valkama, M. Renfors, and V. Koivunen, "Advanced methods for I/Q imbalance compensation in communication receivers," IEEE Trans. Signal Process., vol.49, no.10, pp.2335–2344, Oct. 2001.
[10] X. Huang and M. Caron, "Adaptive gain/phase/DC offset compensation in quadrature modulators," ISPACS'01, Nashville, Tennessee, USA, Nov. 2001.
[11] S. Haykin, Adaptive Filter Theory, 3rd ed., Prentice-Hall, Upper Saddle River, NJ, 1996.
[12] D.S. Hilborn, S.P. Stapleton, and J.K. Cavers, "An adaptive direct conversion transmitter," IEEE Trans. Veh. Technol., vol.43, no.2, pp.223–233, May 1994.
[13] S.R. Kim, Y.G. Jeong, and I.K. Choi, "A constrained MMSE receiver for DS/CDMA systems in fading channels," IEEE Trans. Commun., vol.48, no.11, pp.1793–1796, Nov. 2000.

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