INTERNATIONAL JOURNAL OF
INFORMATION AND SYSTEMS SCIENCES
Volume 3, Number 4, Pages 541-554

A NEW TECHNIQUE FOR DIGITAL PRE-COMPENSATION IN IQ MODULATOR

ANTHONY G. LIM, VICTOR SREERAM AND GUO-QING WANG

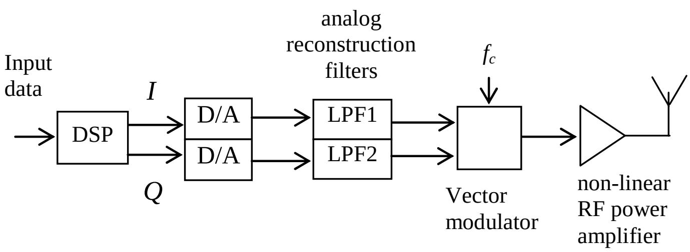
Abstract In DSP based IQ modulators generating CPFSK signals, departures from a flat-magnitude, linear phase frequency response in the pass bands of analogue reconstruction filters in the I and Q channels cause ripples in the modulator output signal envelope. These ripples then produce undesirable side-lobes in the output signal spectrum when the signal passes through non-linear elements in the transmission path.

In this paper, we present a new technique of designing digital FIR pre-compensation filters in IQ modulators. These FIR filters are found by minimizing the mean square error (MSE) between the desired channel response and the response from the low-pass analogue reconstruction filter. This technique does not require the estimation of the I and Q channel analogue reconstruction filter impulse responses. Furthermore, we propose a time-domain approach based on decimation and interpolation to overcome the problem of ill-conditioned matrix and to reduce the computational load of the long FIR filters placed on the DSP. These techniques have shown to be effective in improving the modulator performance by achieving substantial reduction on the output envelope ripples.

Key Words, Digital compensation, IQ modulator, Decimation, Interpolation

1. Introduction

In-phase/Quadrature (IQ) modulation is a versatile and widely used technique in transceiver architecture, due to its flexibility in generating a wide variety of amplitude, phase or frequency modulated signals. A flexible approach to the implementation of a radio transmitter is to synthesize the base-band (low frequency) I and Q signals digitally using a Digital Signal Processor (DSP), followed by a digital-to-analogue (D/A) converter section and analogue vector modulator to directly up-convert the base-band signals to the radio frequency (RF) for transmission. This configuration is shown in Fig. 1.



Received by the editors October 10, 2006 and revised July 15, 2006.

Fig. 1 Radio transmitter architecture incorporating digital *IQ* modulation

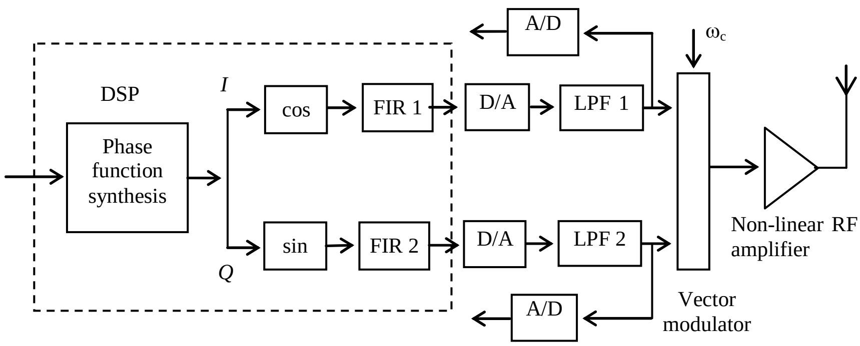


Fig. 2**.** *IQ* modulator and digital pre-compensation structure.

This approach can be limited, however, by the two analogue reconstruction filters labelled LPF1 and LPF2 in Fig. 1. Ideally, the two analogue filters should have identical low-pass frequency response characteristics with flat magnitude and linear phase. In practice, the transfer characteristics of practical reconstruction filters and errors in their implementation result in the pass-band characteristics departing from constant magnitude and linear phase. Furthermore, implementation errors also result in a mismatch between the I and Q channel reconstruction filter frequency response. In the case of Continuous Phase Frequency Shift Keying (CPFSK) signals, this results in the loss of the desired constant envelope property of the output signal and thus, causing significant degradation of the performance of the transmitter system when the signal passes through the nonlinear RF power amplifier [1].

Recently, several techniques have been developed to compensate for the imperfections in the analogue subsystems of quadrature modulators and demodulators [2]-[7]. The approach in [2] centers on the introduction of two Finite Impulse Response (FIR) filters into the IQ modulator structure as shown in Fig. 2 to pre-compensate for both imbalances in the analogue reconstruction filters' frequency responses as well as departures from constant magnitude, linear phase in the pass-band of each reconstruction filter. These FIR filters are derived from least-squares (LS) technique. An alternative solution is presented in [7] using state-space approach. These approaches involve two steps:

1. Estimation of the I and Q channel reconstruction filters' impulse responses using Pseudo Random Binary Noise (PRBN) test signal.
2. Computation of the optimum FIR filter tap-weights using estimated I and Q channel analogue filters' impulse responses.

In [6], a digital compensation scheme using IIR (Infinite Impulse Response) pre-compensation filters was proposed. These IIR filters are designed using an indirect approach, by performing model reduction on the FIR filters in [2]. Although these techniques have shown to be effective in improving modulator performance, they result in filters that have a large number of coefficients and hence, are computationally demanding to implement on the DSP. Furthermore, the solution matrix to a LS optimization problem is often ill-conditioned and must first be regularized before the solution vector is computed in order to be successful in practical applications. Lee et al [4] presented some encouraging results of an investigation into reducing the computational burden placed on the DSP by changing the delay/tap spacing of the FIR pre-compensation filters.

In this paper, we propose a new technique of designing the FIR pre-compensation filters. It is a one-step approach, where the compensation filters tap-weights are determined without having to estimate the impulse responses of the I and Q channel reconstruction filters. In the presented algorithm, the FIR filters are found by minimizing the MSE of the desired channel response and the output response of the analogue reconstruction filter. To overcome the problem of computational efficiency and numerical instability due to ill-conditioned matrix, we present an effective approach based on sampling rate reduction and sampling rate increase to compute the FIR pre-compensation filters. Hereafter, we will generally refer to these processes as decimation and interpolation respectively. In this approach, the output data is first decimated, then the MSE estimate of the decimated FIR filters' impulse response is calculated from these decimated data, and finally the filters' impulse response is interpolated to reconstruct the original impulse response of high sampling rate. It is found that this approach, not only improves the condition number of the solution matrix and computational efficiency of the FIR pre-compensation filters, but also reduces the output envelope ripples significantly.

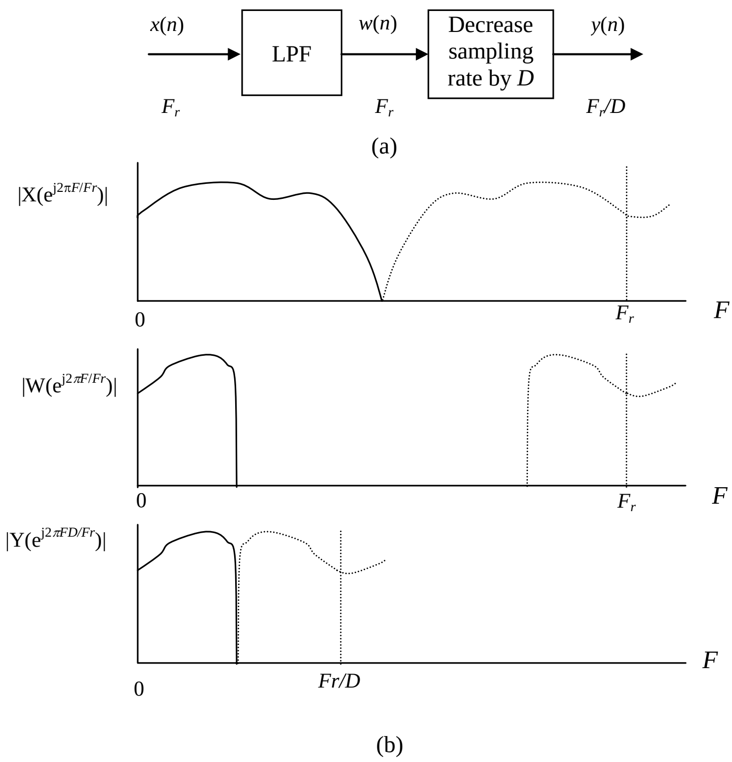
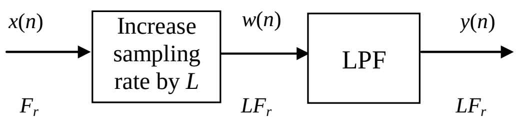


Fig. 3 (a) Block diagram of the decimation operations and (b) frequency response interpretation.

In the next section, we review some of the basic principles of decimation and interpolation established in signal processing theory. In Section 3, we introduce the digital compensation technique through a problem formulation and show the necessary steps for calculating the FIR pre-compensation filters by minimizing the MSE of the error signal. In Section 4, we present the new approach of calculating the FIR pre-compensation filters based on decimation and interpolation. In Section 5, we present some simulation results to illustrate the effectiveness of the proposed techniques. Finally, conclusions are given in Section 6.

2. Preliminary Review of the Decimation and Interpolation Processes

The process of lowering the sampling rate of a signal is called decimation, while the process of increasing the sampling rate is called interpolation [9]. The use of sampling rate reduction and sampling rate increase in signal processing applications have been examined from several viewpoints [10],[11],[12]. In [10], an efficient approach is presented to improve the MSE of the LS estimate of an impulse response based on decimation and interpolation in system identification. Crooke and Craig [11] have shown that for band limiting applications, computational efficiencies can be gained by sampling rate reduction. In this section, we will introduce some of the fundamental concepts of decimation and interpolation in the context of digital signal processing as discussed in [9].



(a)

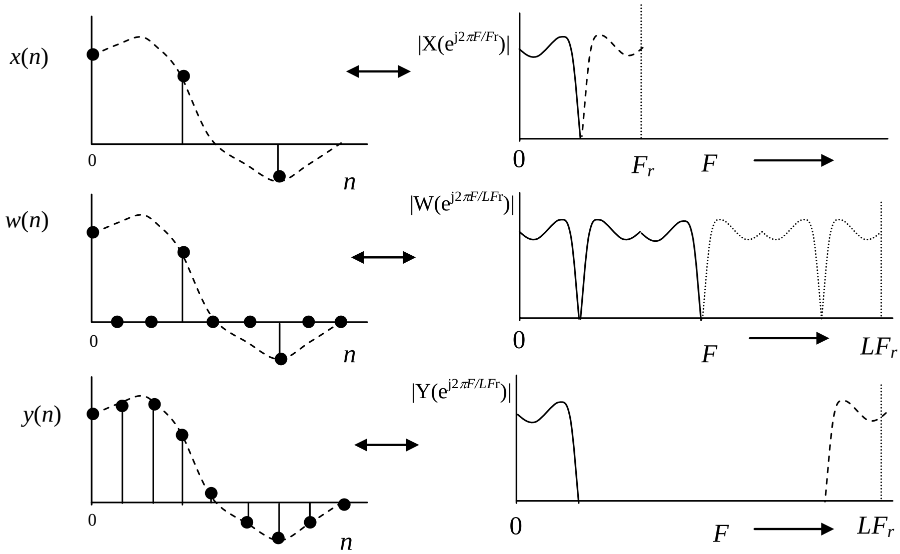


Fig. 4 (a) Block diagram of the interpolation operations and (b) frequency response and waveform interpretation.

The process of decimating a signal $x(n)$ by an integer factor D is shown in Fig. 3. The original sampling rate is denoted as F_r and the new sampling rate is then F_r/D. To avoid aliasing at the lower sampling rate, F_r/D, it is necessary to filter the signal $x(n)$ with a low-pass filter giving the signal $w(n)$. The sampling rate reduction is then achieved by forming a new sequence $y(n)$ by extracting every D^{th} sample of $w(n)$. A block diagram of these processes is given in Fig. 3(a). Fig. 3(b) shows typical spectra (magnitude of Discrete Fourier transforms) of the signals $x(n)$, $w(n)$ and $y(n)$.

The process of interpolating a signal $x(n)$ by an integer ratio L is depicted in Fig. 4. In this case, the sampling rate of a signal $x(n)$ is increased by a factor L by inserting L-1 zero-valued samples between each sample of $x(n)$. This creates a signal $w(n)$ (with sampling rate LF_r) whose frequency components are periodic with period equal to the original sampling frequency F_r as shown in Fig. 4(b). To eliminate these periodic components and retain only the base-band frequencies, it is necessary to filter the signal $w(n)$ with an appropriate low-pass filter. The resulting signal $y(n)$ with sampling rate LF_r is then the desired interpolated signals. The interpolation processes together with the corresponding frequency response interpretations are shown in Fig. 4.

3. Digital Pre-compensation

The digital compensation scheme discussed in this section is similar to that in [2] of which the structure is shown in Fig. 2. The two sets of FIR filters in Fig. 2 are chosen such that the overall discrete-time channel response (from the FIR filter input to the output of the analogue-to-digital (A/D) converter) has a transfer function that matches a desired response. This optimization structure is shown in Fig. 5 where the lower branch represents the one channel (*I* or *Q*) from the input to the FIR pre-compensation filter to the output of the A/D converter.

$D(z)$ is the nominal desired response transfer function (i.e. the response that the *I* or *Q* channels are required to have) and $H(z)$ is the equivalent discrete-time transfer function of the D/A converter, analogue low-pass filter and A/D converter in cascade. In this particular application, $D(z)$ is chosen to have the same magnitude characteristic as the nominal response of the analogue reconstruction filter but is constrained to have linear phase. Let $D(z)$ and $H(z)$ be m^{th} order FIR filters and suppose the FIR pre-compensation filter has coefficients $w(l) = [w(0), w(1),\ldots, w(N\text{-}1)]^T$, then the output $y_d(n)$ and $y(n)$ are defined by the following

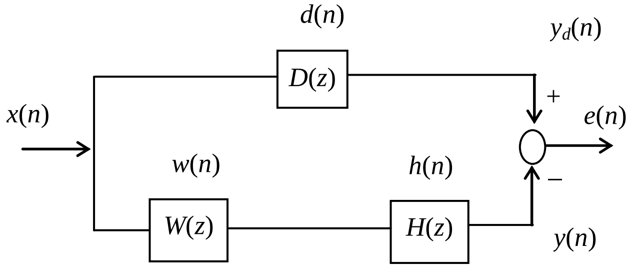


Fig. 5. *I* or *Q* channel optimization structure.

$$y_d(n) = \sum_{k=0}^{m} d(k)x(n-k) \tag{1}$$

$$y(n) = \sum_{k=0}^{N+m-1}\left[\sum_{l=0}^{N-1} w(l)h(k-l)\right]x(n-k) \tag{2}$$

where $d(n)$ is the desired channel impulse response, and $h(n)$ is the discrete-time equivalent impulse response of the D/A, low-pass filter and A/D converter. We define the cost function for the above optimization problem as

$$J = E\left\{e^2(n)\right\} = E\left\{\left[y_d(n) - y(n)\right]^2\right\}$$

$$= E\left\{\left[\sum_{k=0}^{m} d(k)x(n-k) - \sum_{k=0}^{N+m-1}\left[\sum_{l=0}^{N-1} w(l)h(k-l)\right]x(n-k)\right]^2\right\} \tag{3}$$

Then the coefficients $w(l)$ that minimizes the MSE in (3) can be found by setting the partial derivatives of J with respect to $w(l)$ equal to zero, as follows

$$\frac{\partial J}{\partial w(l)} = E\left\{-2y_d(n)y_h(n-l) + 2\left[\sum_{j=0}^{N-1} w(j)y_h(n-j)\right]y_h(n-l)\right\} = 0;$$

$$l = 0, 1, \ldots, N\text{-}1 \tag{4}$$

where

$$y_h(n) = \sum_{k=0}^{m} h(k)x(n-k) \tag{5}$$

or equivalently,

$$\sum_{j=0}^{N-1} w(j)E\left\{y_h(n-j)y_h(n-l)\right\} = E\left\{y_d(n)y_h(n-l)\right\};$$

$$l = 0, 1, \ldots, N\text{-}1 \tag{6}$$

In order to simplify (6) notationally, define

$$r_y(j,l) = E\left\{y_h(n-j)y_h(n-l)\right\} \tag{7}$$

and

$$r_{dh}(l) = E\left\{y_d(n)y_h(n-l)\right\} \tag{8}$$

which are the sample autocorrelation and cross-correlation sequences respectively. It can be shown that each term in (7) depends only on the difference between the indices j and l (See [8], Chapter 4). Consequently, we define $r_y(|j - l|)$ to replace $r_y(j, l)$

$$r_y(|j-l|) \equiv r_y(j,l) = E\{y_h(n-j)y_h(n-l)\} \tag{9}$$

Then the system of linear equations in (6), when expressed in matrix form, becomes

$$\begin{bmatrix} r_y(0) & r_y(1) & \cdots & r_y(N-1) \\ r_y(1) & r_y(0) & \cdots & r_y(N-2) \\ \vdots & \vdots & & \vdots \\ r_y(N-1) & r_y(N-2) & \cdots & r_y(0) \end{bmatrix} \begin{bmatrix} w(0) \\ w(1) \\ \vdots \\ w(N-1) \end{bmatrix} = \begin{bmatrix} r_{dh}(0) \\ r_{dh}(1) \\ \vdots \\ r_{dh}(N-1) \end{bmatrix} \tag{10}$$

or

$$\mathbf{R}_y \mathbf{w} = \mathbf{r}_{dh} \tag{11}$$ [1]

where R_y is the N x N Hermitian matrix of autocorrelations, w is the vector of compensation filter coefficients, and r_{dh} is the column vector of cross-correlations between $y_d(n)$ and $y_h(n)$

$$\mathbf{r}_{dh} = \begin{bmatrix} r_{dh}(0) & r_{dh}(1) & \cdots & r_{dh}(N-1) \end{bmatrix}^T \tag{12}$$

If Ry is nonsingular then the coefficients $w(l)$ that minimize J are

$$\mathbf{w} = \mathbf{R}_y^{-1} \mathbf{r}_{dh} \tag{13}$$

The above algorithm requires the measurement of $y_h(n)$ to calculate the coefficients of the FIR pre-compensation filters. This can be done by setting $w(l)$ to be a single pulse of length N as follows

$$w(l) = \begin{cases} 1 & ; \quad l = 0 \\ 0 & ; \quad l = 1, 2, \ldots, N-1 \end{cases} \tag{14}$$

so that $y_h(n)$ is in the form of (5). We will now summarize the preceding results in the following algorithm:

Algorithm 1: Computation of FIR pre-compensation filter.

(i) Measure the output sequence $y_d(n)$ and $y_h(n)$ by passing the input x(n) through filter $D(z)$ and $H(z)$ with $w(l)$ initially set as in (14). In this paper, we consider white noise sequence as input $x(n)$.

(ii) Form R_y and r_{dh} for I and Q channel using time averages.

(iii) Compute the coefficients of the FIR pre-compensation filters in each channel using (13).

---

[1] Solving (11) requires that the autocorrelation $r_y(|j\text{-}l|)$ and the cross-correlation $r_{dh}(l)$ be known. Since these ensemble averages are unknown, time averages are used to estimate these ensemble averages (See [8], Chapter 4).

4. Design based on Decimation and Interpolation

In this section, we present an efficient approach of calculating the digital FIR pre-compensation filters using the concepts of signal decimation and interpolation developed in Section 2. In Section 3, it is shown that the optimum FIR filters can be found by solving (13), which is a function of the auto-correlation and cross-correlation sequence. Therefore, it is possible to apply signal decimation and interpolation in this context to design FIR filters with fewer coefficients in order to increase the computational efficiency of the digital compensation scheme.

The proposed algorithm consists of three steps: the first step is the decimation of the data, the second step is the calculation of the FIR pre-compensation filters' coefficients using the decimated input/output data, and the third step is the recovery of the original FIR filters impulse response of high sampling rate by interpolation. In the first step, for reducing the sampling rate of the output signals $y_d(n)$ and $y_h(n)$ by D, the signals are only saved at every D instant in time[2]. The decimated output sequence is denoted by $y'_d(D)$ and $y'_h(D)$, i.e. $y'_d(D)=[y_d(0), y_d(D), y_d(2D),\ldots]^T$ and $y'^h(D)=[y_h(0), y_h(D), y_h(2D),\ldots]^T$ respectively. From (6), the system of linear equations becomes

$$\sum_{j=0}^{\lceil N/D \rceil -1} w(Dj) r_y \left(D \,|j-l|\right) = r_{dh}(Dl) \qquad l = 0, 1, \ldots, \lceil N/D \rceil -1 \tag{15}$$

where $\lceil N/D \rceil$ denotes the smallest integer greater or equal to N/D. Denote K as $\lceil N/D \rceil$, then in matrix form, (15) can be expressed as

$$\begin{bmatrix} r_y(0) & r_y(D) & \cdots & r_y(D(K-1)) \\ r_y(D) & r_y(0) & \cdots & r_y(D(K-2)) \\ \vdots & \vdots & & \vdots \\ r_y(D(K-1)) & r_y(D(K-2)) & \cdots & r_y(0) \end{bmatrix} \begin{bmatrix} w(0) \\ w(D) \\ \vdots \\ w(D(K-1)) \end{bmatrix} = \begin{bmatrix} r_{dh}(0) \\ r_{dh}(D) \\ \vdots \\ r_{dh}(D(K-1)) \end{bmatrix} \tag{16}$$

or

$$\mathbf{R}_y(D)\mathbf{w}(D) = \mathbf{r}_{dh}(D) \tag{17}$$

where $R_y(D)$ is the K x K autocorrelation matrix calculated using the decimated data, $w(D) = [w(0), w(D), \ldots.w(D(K\text{-}1))]^T$ and $r_{dh}(D) = [r_{dh}(0), r_{dh}(D), \ldots, r_{dh}(D(K\text{-}1))]^T$. Note that in the case of $D = 1$, (15) corresponds to the ordinary system of linear equations in (6).

The second step is to calculate the FIR filter coefficients from the decimated data by:

---

[2] In the decimation process, the signals are usually filtered through a low-pass filter in order to avoid aliasing. However, in this case, this step is not required since both $y_d(n)$ and $y_h(n)$ are signals passing through $D(z)$ and $H(z)$ that have low-pass characteristics. $H(z)$ is the equivalent discrete-time low-pass filter and $D(z)$ is the desired response which is chosen to have the same magnitude characteristic as the analog low-pass filter (See Section III).

$$\mathbf{w}(D) = \mathbf{R}_y(D)^{-1}\mathbf{r}_{dh}(D) \tag{18}$$

The FIR filter's impulse response, $w(D)$ in (18) is evaluated only at every D^{th} sampling instant. Therefore, it is necessary to increase the sampling rate by an integer factor D by inserting D-1 zero-value samples of impulse response between the decimated sampling instant [9] as follows:

$$\tilde{w} = \big[w(0), \underbrace{0,\ldots,0}_{D\text{-}1}, w(D), \underbrace{0,\ldots,0}_{D\text{-}1},\ldots, w(D(K-1)), \underbrace{0,\ldots,0}_{D\text{-}1}\big]^T \equiv [\{\tilde{w}_k\}] \tag{19}$$

This is followed by the filtering with low-pass filter to eliminate the images produced by the interpolation process. Since the FIR filter is followed by a low-pass filter in the modulator, the resulting output signal is the desired interpolated signal with the original high sampling rate.

In this approach, the order of the compensation filter is preserved, i.e. the FIR filter is still of N^{th} order. This approach results in significant savings in computation since the FIR filter impulse response samples between the sampling instant D are zero. In addition, the condition number of the R_y matrix is also greatly improved because the Euclidean distance between the adjacent vectors in forming $R_y(D)$ is increased.

5. Simulation Studies

To demonstrate the effectiveness of this digital compensation technique, we present the results of MATLAB simulation studies centered on a single channel of an ERMES modulation format transmitter, which generates a 4-level pulse amplitude modulation (PAM)/CPFSK output signal. The modulator has the basic structure shown in Fig. 2. The two low-pass reconstruction filters LPF1 and LPF2 have a nominal 6th order *Butterworth* characteristics, implemented using three *Sallen & Key* second order sections with component tolerances of 5% for resistors and 10% for capacitors. The sampling frequency is 200 kHz [13]. The optimum delay parameter, τ_o in the desired channel response is solved numerically [2] and for this example, is found to be 150 μs.

Fig. 6 shows the envelope functions obtained for the system with and without pre-compensation. These envelope signals were generated using a random sequence of 4000 inputs symbols (4-PAM/CPFSK) to the CPFSK modulator while the FIR compensation filters are found using the algorithm in Section 3.

Table 1. RMS envelope ripple for different FIR filter lengths.

FIR filter length	RMS envelope ripple	Ripple reduction factor
50-tap	0.034 mV	92
30-tap	0.068 mV	46
20-tap	0.219 mV	14
Uncompensated	3.158 mV	-

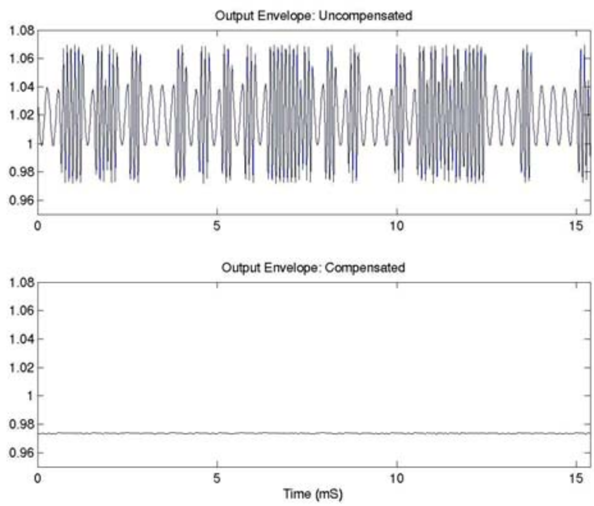


Fig. 6. Vector modulator output envelope functions
upper plot: without pre-compensation
lower plot: with pre-compensation (50-tap FIR filter)

Table 2. RMS envelope ripple using method of [2]

FIR filter length	RMS envelope ripple	Ripple reduction factor
50-tap	0.036 mV	87
30-tap	0.174 mV	18
20-tap	0.508 mV	6
Uncompensated	3.158 mV	-

The RMS values of the envelope ripple using different filter tap lengths are given in Table 1. Also shown for reference is the RMS value of the envelope ripple for the uncompensated system. From Table 1 and the envelope function plot, it can be seen that the FIR filters obtained via the new technique can effectively reduce the output envelope ripple. An identical computer simulation is carried out using the method proposed in [2] and the results are recorded in Table 2. From Table 1 and 2, it can be seen that the new technique gives better results in terms of ripple reduction factor for all filter lengths considered.

Table 3 gives the results obtained by performing algorithm outlined in Section 4 for various D values. Please note that the system with D = 1 corresponds to ordinary system that does not involve signal decimation and interpolation. In Table 3, it is shown that the condition number improves by a factor of 10^3 for every increment in D up to D = 3. This provides a solution that is less sensitive to numerical errors and therefore avoiding the need to discard eigenvalues in a regularization process as in [2].

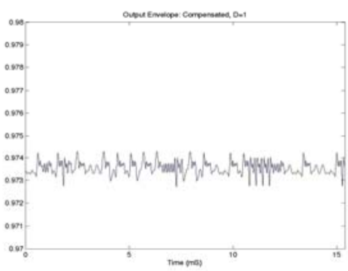
Table 3. Summary of RMS values of output envelope ripples.

Delay/tap, D	Taps	RMS ripple (mV)	Ripple reduction factor	Condition No. of $\mathbf{R}_y$ matrix
1	50	0.034	92	$\sim 10^7$
2	25	0.026	118	$\sim 10^4$
3	17	0.094	33	$\sim 10^1$
4	13	0.315	10	$\sim 10^0$

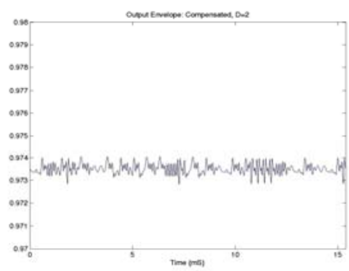
Table 4. Summary of RMS values of output envelope ripples using method of [4].

Delay/tap	Taps	RMS ripple (mV)	Ripple reduction factor
2	25	0.028	114
3	17	0.086	37
4	13	1.022	3

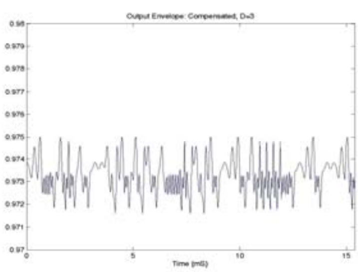
In terms of computational efficiency, the new FIR filters offers great computational savings since the filters' coefficient are only nonzero for every D sampling time instant and therefore reducing the effective number of multiplication and addition operations required per output sample. Furthermore, the size of the $\mathbf{R}_y$ matrix is reduced by approximately a factor of D and thus less computation is required in calculating the matrix inverse.

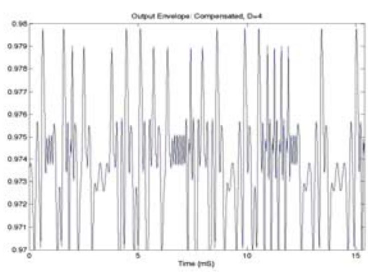


(a)



(b)





(c) (d)

Fig. 7. Output envelope ripples for various *D* values (a) $D = 1$, (b) $D = 2$, (c) $D = 3$ and (d) $D = 4$ (on expanded scale).

Fig. 7 shows the output envelope ripples of the *IQ* modulator on expanded scale for various *D* values. From Table 3 and visual inspection of the plots, the best result is obtained when $D = 2$. Table 4 shows the results obtained using the approach outlined in [4]. From Table 3 and 4, we can see that the new technique presented in Section IV performs slightly better than the method of [4] in all cases except for $D = 3$. All the above simulations are carried out using the same delay parameter, τ_o of 150 μs.

6. Conclusions

A new technique of designing the FIR pre-compensation filters is presented. Simulation results presented in Section 5 show that substantial reduction of the output envelope ripples can be achieved using the proposed technique. The major advantage of this technique is that it does not require the extensive numerical estimation of the impulse response of the analog reconstruction filters. It is a one step approach in calculating the *I* and *Q* channel optimum pre-compensation filter.

Since the pass-band of the analogue low-pass filter response to be compensated only occupies a small band of frequency, it is possible to extend the theory of signal decimation and interpolation in this application to reduce the number of non-zero filter coefficients required to achieve the same level of compensation. This time-domain approach based on decimation and interpolation offers an efficient solution to the computational load and ill-conditioned issues encountered by FIR filters in digital compensation scheme. Furthermore, this time-domain formulation also provides a neat solution which is easy to compute in MATLAB. However, further investigation into this approach is required and will be the subject of on-going research.

Acknowledgments

The authors would like to thank our colleagues Prof. A. Cantoni, Dr. J. Tuthill and Dr. Y. H. Leung for their comments in the course of writing this paper.

References

[1] S. Leyonhjelm and M. Faulkner, "*The Effect of Reconstruction Filters on Direct Upconversion in a Multichannel Environment*," *IEEE Trans. Vehic. Tech.*, vol. 44, no. 1, pp. 95-102, Feb. 1995.

[2] J. Tuthill and A. Cantoni, "*Optimum Pre-compensation Filters for IQ Modulation Systems*," *IEEE Trans. Comm.*, vol. 47, no. 10, pp. 1466-1468, Oct. 1999.

[3] R. Marchesani, "*Digital Precompensation of Imperfections in Quadrature Modulators*," *IEEE Trans. Comm.*, vol. 48, no. 4, pp. 552-556, April 2000.

[4] G.C. Lee, J. Tuthill and A. Cantoni, "*Efficient Implementation of Digital Compensation in IQ Modulators*," *Proc. IEEE ICASSP 02*, Orlando, Florida, pp. 2697-2700, May 2002.

[5] M. Valkama and M. Renfors, "*Digital Filter Design for I/Q Imbalance Compensation*," *Proc. 10th*

European Signal Processing Conference (*EUSIPCO* 2000), Tampere, Finland, Sep. 2000.

[6] E. H. Soh, J. Tuthill, V. Sreeram and A. Cantoni, "*Digital Compensation in IQ Modulators Using IIR Filters*," *Proceedings TENCON-2001*, Singapore, Aug. 2001.

[7] A.G. Lim and V. Sreeram, "*Optimum Digital Pre-Compensation in IQ Modulators using State-Space Approach*," *Proc. 5th International Conference on Optimization: Techniques and Applications* (*ICOTA* 2001), vol. 1, pp. 324-331, Hong Kong, Dec. 2001.

[8] M. H. Hayes, *Statistical Digital Signal Processing and Modelling*, *John-Wiley & Sons*, New York, 1996.

[9] R.E. Crochiere and L. R. Rabiner, *Multirate Digital Signal Processing*, *Prentice Hall*, Engelwood Cliffs, N.J. 1983.

[10] J. Xin, H. Tsuji, and A. Sano, "*Optimal sampling interval for system identification based on decimation and interpolation*," *IEE Proc.-Control Theory Applications*, vol. 142, no. 1, pp. 15-22, Jan. 1995.

[11] A.W. Crooke and J. W. Craig, "*Digital filters for sample-rate-reduction*," *IEEE Trans. Audio Electroacoust*., vol. AU-20, pp.308-315, Oct. 1972.

[12] R.E. Crochiere and L.R. Rabiner, "*Optimum FIR Digital Filter Implementations for Decimation, Interpolation, and Narrow-Band Filtering*," *IEEE Trans. Acous., Speech and Sig. Processing*, vol. ASSP-23, no.5, pp. 444-456, Oct. 1975.

[13] J. Bustillo, M. Rodriguez-Palanca and J. Perez, "*4-PAM/FM Modulator with DSP: A Solution for ERMES*," *IEEE 41st Vehicular Technology Conference*, pp. 548-588, 1991.

Dr. Anthony G. Lim was born in Kuching, Sarawak, Malaysia. He received his Bachelor's degree and Ph.D degree in Electrical and Electronic Engineering from the University of Western Australia, Australia in 2000 and 2004 respectively.

Currently, he is working at Deloitte Touche Tohmatsu as a Management Consultant.

His research interests include control techniques and signal processing.

Victor Sreeram, School of Electrical, Electronic and Computer Engineering, University of Western Australia, WA, Australia. Email: sreeram@ee.uwa.edu.au

Dr. Sreeram obtained Bachelor's degree in 1981 from Bangalore University, India, Master's degree in 1983 from Madras University, India, and Ph.D degree from University of Victoria, Canada in 1989, all in Electrical Engineering.

He worked as a Project Engineer in the Indian Space Research Organisation from 1983 to 1985. He joined the Department of Electrical & Electronic Engineering, University of Western Australia in 1990 and he is now an Associate Professor.

He has held Visiting Appointments at the Department of Systems Engineering, Australian National University during 1994, 1995 and 1996 and at the Australian

Telecommunication Research Institute in Curtin University of Technology during 1997 and 1998.

His research interests are control and signal processing.

Dr. Guo-Qing Wang obtained his Bachelor's and Master's degrees from Zhejiang University, Hangzhou and University of Electronic Science and Technology, Chengdu respectively. He then obtained is Ph.D. degree from the University of Western Australia, Australia in 2001.

During 1999-2002, he was a Research Associate in the Control Systems Research Group, Department of Electrical and Electronic Engineering, University of Western Australia. Since 2003 he is an adjunct visiting fellow in the same research group.