Journal of Engineering and Development, Vol. 17, No.4, October 2013, ISSN 1813- 7822

Design of 16-QAM Transmitter Based on Software De fined Radio Technology

Lecturer-Faeza Abbas Abed
Department of Electronic Techniques, Institute of technology, 2013
Baghdad-Iraq, E-mail: dr.faezaa@yahoo.com

Abstract :

In this paper, a 16-QAM transmitter for software defined radio is designed and simulated using MATLAB-Simulink environments. The reconfigurable and reliable in the transmitter permit to do multiple functions of modulation schemes and pulse shaping in the same time. A software defined radio (SDR) technology could implement all transmitter requirements with less resources and power consumption. However, the transmitter part of SDR transceiver is investigated deeply for 70 MHz. The analysis of constellation and eye diagrams of multiple QAM schemes has been studied, examined and compared to achieve multiple system requirements. The experimental and simulation results shows an important development and the SDR transmitter style guide for current and future wireless and mobile systems. This development could support the 3G and accelerate the transformation to 4G.

Keywords-SDR, 16-QAM, Transmitter, MATLAB

تصميم مرسلة تعديل اتساعي رباعية الطور بتقنية الرادبو المعرف برمجيا

مدرس - فائزة عباس عبد
قسم التقنيات الالكترونية – معهد التكنلوجيا بغداد-2013

الخلاصة :

هذا البحث يعرض تصميم ومحاكات مرسلة ذات التعديل الاتساعي بتقنية الراديو المعرف برمجيا باستخدام برنامج ماتلاب. الموثوقية واعادة التشكيل في المرسلة تسمح للقيام بعدة مهام للتحميل وتشكيل النبضات في نفس الوقت. في تكنلوجيا الراديوات المعرفة برمجيا يمكن تنفيذ جميع متطلبات الارسال باقل الموارد واستثمار امثل للطاقة. لذلك فان قسم الارسال في مرسل مستلم الراديو المبرمج تم التحقق منه باسهاب وبحزمة التردد المتوسط 70 ميكاهيرتز. كما تم تحليل ودراسة المخططات البيانية والعينية لهذه التقنية وتمت مقارنة الاختبارات لمختلف الانظمة. نتائج التجارب والمحاكات تبنين التطور الهام في الارسال المبرمج ويمكن اعتبارها دليلا للانظمة الحالية والمستقبلية. يمكن ان يدعم هذا التطوير الجيل الثالث لانظمة الاتصالات ويعجل بالانتقال الى الجيل الرابع منه.

1. Introduction

SDR as a "radio in which some or all of the physical layer functions are software defined." This implies that the architecture is flexible such that the radio may be configured, occasionally in real time, to adapt to various air standards and waveforms, frequency bands, bandwidths, and modes of operation. That is, the SDR is a multifunctional, programmable, and easy to upgrade radio that can support a variety of services and standards while at the same time provide a low-cost power-efficient solution.

In commercial world, the need to adopt SDR principles is becoming more and more apparent due to recent developments in multi-mode multi-standard radios and the various complex applications that govern them. The flexibility of SDR is ideally suited for the various quality-of-service (QoS) requirements mandated by the numerous data, voice, and multimedia applications. Today, many base-station designs employ SDR architecture or at least some technology based on SDR principles. On the other hand, surely but slowly, chipset providers are adopting SDR principles in the design of multi-mode multi-standard radios destined for small form-fit devices such as cellophanes and laptops .The SDR architecture is a flexible, versatile architecture that utilizes general-purpose hardware that can be programmed or configured in software [1]. Compared to traditional architectures that employ quadrature sampling, SDR radios that employ intermediate frequency (IF) sampling tend to do more signal processing in the digital domain. This particular radio architecture, the advantage of SDR over traditional solutions stems from its adaptability toits environment and the number and type of application that it can support. This can be accomplished since the hardware itself and its operation is abstracted completely from the software via a middleware layer known as the hardware abstraction layer (HAL). The purpose of the HAL is to enable the portability and the reusability of software [2] .This common framework enables the development and deployment of various applications without any dependencies on the particular radio hardware. **Figure (1)** illustrates the test-bed implementation of the design process for reconfigurable computing[3].

Many attempts in this field has been presented a minimum BER for different channels have been achieved, by a multi SDR model design and implementation in FPGA form using Xilinx system generator technique [4]. Full QAM for modulation and demodulation Simulink-based system using parameters setting for random generator in an AWGN wireless channel was provided, this model was designed to show the criteria for adaptively of QAM system [5]. A comparison study of the function laities of digital modulation and demodulation along with controlling features such as symbol generation, filtering, modulating, demodulating, phase is presented by[6-7-8].

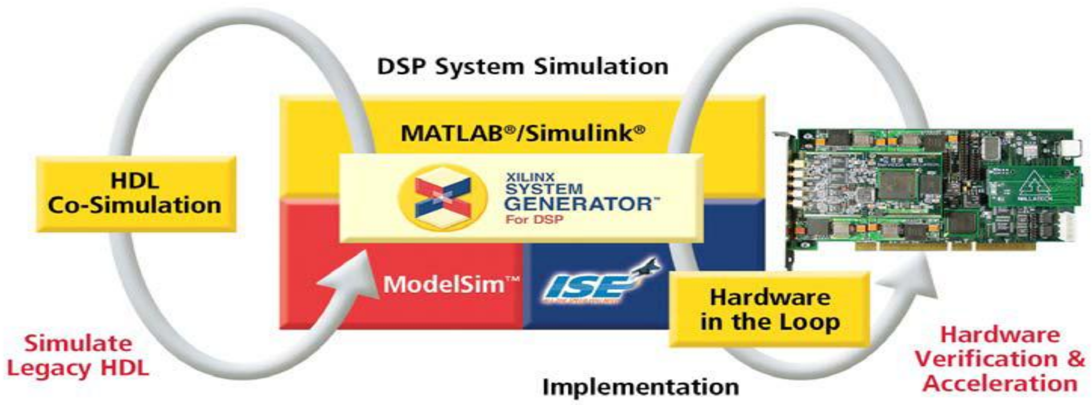


Fig.(1) Test-bed implementation process

2. SDR Transmitter Model

Simulation of the SDR system with SIMULINK tools forms is the first step of the design is process for reconfigurable computing. The complete simulink SDR transmitter model shown in **Figure (2)** and the parameters of proposal model are given in **Table (1).**

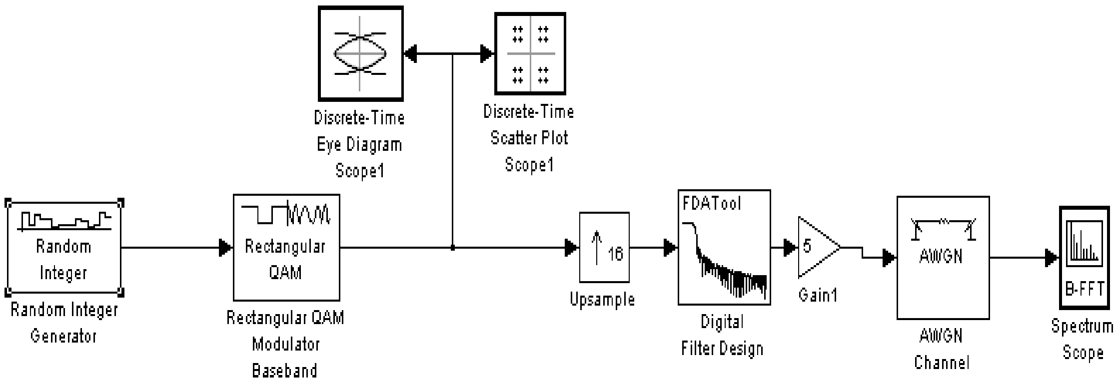


Fig.(2) QAM Transmitter Model

Table 1: Simulation Parameters of Proposed Model-

Modulation Type	16QAM
Symbol Rate	2.5Msymbol/Second
Interpolation Factor	16
RRC Filter order	64
Transmitter Sampling Rate	40 MHz
RRC Cutoff Frequency	2.5 MHz
RRC Sampling Frequency	40 MHz
RRC Rolloff Factor	0.35
RRC Beta	0.5
RRC Window Type	Kaiser Window

2.1 Modulation Technique

With the fast development of modern communication techniques, the demand for reliable high date rate transmission is increased significantly, which stimulate much interest in modulation techniques. Different modulation techniques allow you to send different bits per symbol and thus achieve different throughputs or efficiencies. QAM is one of widely used modulation techniques because of its efficiency in power and bandwidth.
The constellation points are usually arranged in a square grid with equal vertical and horizontal spacing, although other configurations are possible (e.g. Cross-QAM). Since the data is often binary, the number of points in the grid is usually a power of 2 (2, 4, 8...) and correspondingly, for QAM the most common forms are 16QAM, 64-QAM, 128-QAM and 256-QAM. By moving to a higher-order constellation, it is possible to transmit more bits per symbol. The qudrature amplitude modulation (QAM) is a better option, since it achieves a greater distance between adjacent points in the I-Q plane by distributing the points evenly QAM relies on two mechanisms to encode bits. One is the pulse amplitude, which can assume both positive and negative values, and the other is the use of two simultaneous pulses. The latter requires two independent baseband channels, one for each pulse. One channel is referred to as the I, or in-phase, channel and the other as the Q, or quadrature, channel. The variety of QAM comes forms depending on the number of bits encoded into each pair of pulses. For example, 16 QAM uses a 4-bit symbol to represent 16 possible symbol values; 64 QAM uses a 6-bit symbol to represent 64 possible symbol values; and 256 QAM uses an 8-bit symbol to represent 256 possible symbol values. Generally, QAM symbols encode an even number of bits (4, 6, 8 and so on), but odd bit schemes, though uncommon, exist as well.

For a more detailed look at the QAM encoding scheme, consider 16 QAM. Like the other QAM versions, 16 QAM uses two simultaneous pulses to encode a symbol. Therefore, for 16 QAM, each pulse must be able to assume one of four levels, because two pulses with four possible levels per pulse yields 16 possible combinations. The four possible pulse levels are: +1/3 AMAX, −1/3 AMAX, +AMAX, and −AMAX (AMAX denotes the maximum pulse amplitude). An example of the I and Q pulses for 16 QAM is shown in **Figure 3** (included are the symbol values taken from **Table(2)**.

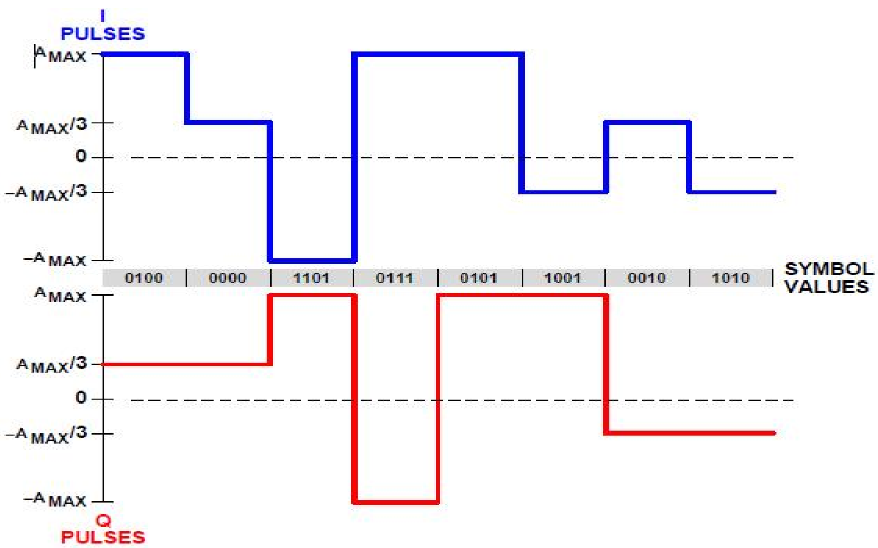


Fig.(3) I and Q pulses for 16 QAM

The amplitude values displayed on the vertical axes of **Figure (3)** provide equal steps in amplitude from the lowest to the highest value. With this arrangement, the levels of the two simultaneous pulses identify a unique 4-bit symbol. For example, let the amplitude of the I and Q pulses be represented by a coordinate pair,(Ik, Qm), where the index values (k and m) rang from 0 to 3 and the amplitudes (Ik and Qm) tak on values of $\pm$ 1/3 AMAX or $\pm$AMAX. Assigning amplitude pairs to symbol values, as shown in **Table (2),** yields the constellation diagram shown in **Figure(4).**

Table 2: The 16-QAM symbol values

Symbol (Binary)	k	m	I_k	Q_m
0000	0	0	1/3 AMAX	1/3 AMAX
0001	0	1	1/3 AMAX	AMAX
0010	0	2	+1/3 AMAX	−1/3 AMAX
0011	0	3	+1/3 AMAX	−AMAX
0100	1	0	AMAX	1/3 AMAX
0101	1	1	AMAX	AMAX
0110	1	2	+AMAX	−1/3 AMAX
0111	1	3	+AMAX	−AMAX
1000	2	0	−1/3 AMAX	+1/3 AMAX
1001	2	1	−1/3 AMAX	+AMAX
1010	2	2	−1/3 AMAX	−1/3 AMAX
1011	2	3	−1/3 AMAX	−AMAX
1100	3	0	−AMAX	+1/3 AMAX
1101	3	1	−AMAX	+AMAX
1110	3	2	−AMAX	−1/3 AMAX
1111	3	3	−AMAX	−AMAX

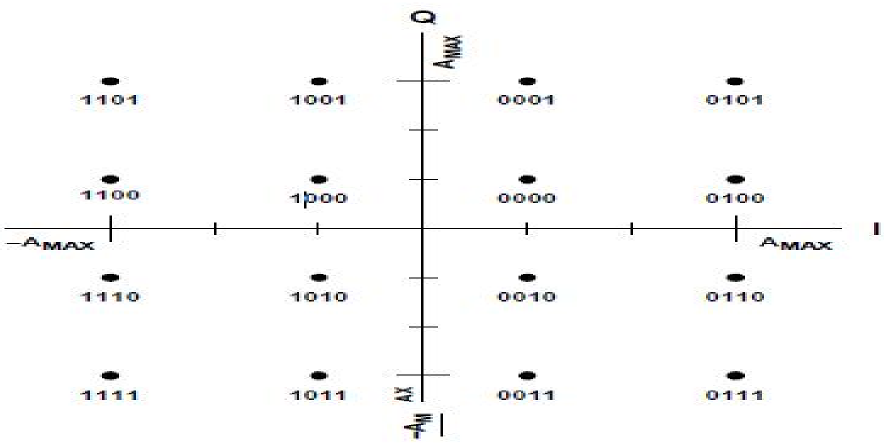


Fig.(4) Constellation diagram of 16QAM

In QPSK, four points on the constellation diagram, equi-spaced around a circle, as shown in **Figure (5).** With four phases, QPSK can encode two bits per symbol. Although QPSK can be viewed as a quaternary modulation, it can be seen as two independently modulated quadrature carriers. With this interpretation, the even (or odd) bits are used to modulate the in-phase component of the carrier, while the odd (or even) bits are used to modulate the quadrature-phase component of the carrier.

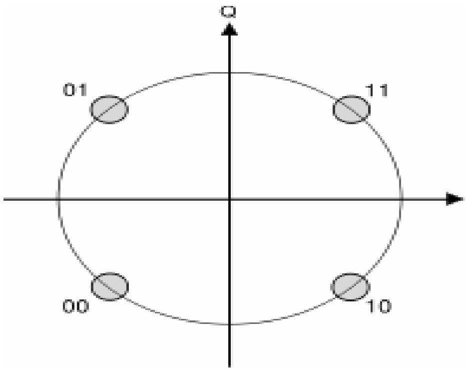


Fig.(5) Constellation diagram of a QPSK modulated carrier

2.2 Pulse Shaping Filter

Finite Impulse Response (FIR) digital filters can have a truly linear phase response and very precise performance. The ability of easily programming the hardware to accommodate different data rates, modulation formats and filter specifications also makes the FIR digital filter very flexible. Furthermore, the performance of digital filter does not vary with environmental changes. FIR filters are realized no recursively, which is by direct evaluation of **Equation (1).**

$$Y(n) = \sum_{k=0}^{N-1} x(n-k).h(k) \qquad (1)$$

Where N is the total tap length of the filter and y(n) is the current output samples that is the function of present and past input values. h(k) is the coefficient of the filter and x(n-k) is the input samples. An alternative representation for FIR in z-domain is given by Equation (2). H (z) is the transfer function of the filter

$$H(z) = \sum_{k=0}^{N-1} h(k).z^{-k} \qquad (2)$$

From the SDR system point of view, it is important to pay attention to the output spectrum those results from applying a shaping function to a modulated signal and its consequences on the radio design as a whole and to the filtering and modulation and demodulation requirements in particular. Pulse shaping has been employed by both analog and digital modulation. Nyquist stated in his landmark paper [1] that the Inter symbol Interference (ISI-free) property can be preserved if the sharp cutoff frequency of the brick-wall filter is modified by an amplitude characteristic having odd symmetry about the Nyquist cutoff frequency. One such filter is the raised cosine filter represented in the time domain as shown in both time and frequency domains for two rolloff factors α=1 and α =0.25 in **Figure (6 , 7)** respectively.

$$g(t) = \frac{\cos\left(\propto \frac{\pi t}{T_s}\right)}{1-(2 \propto t/T_s)^2} \text{ sinc } (t/T_s)$$

$$= \frac{\pi}{4}\{\text{sinc } (\propto t/T_s + 1/2) + \text{sinc } (\propto t/T_s - 1/2)\} \text{ sinc } (t/T_s) \qquad 0 \leq \propto \leq 1 \tag{3}$$

The frequency domain representation of **Equation (3)** is

$$G(f) = \begin{cases} T_s & |f| \leq (1-\propto)\frac{F_s}{2} \\ T_s \frac{1-\sin\left(\pi\left(f-\frac{F_s}{2}\right)\propto F_s\right)}{2} & (1-\propto)\frac{F_s}{2} \leq |f| \leq (1+\propto)\frac{F_s}{2} \\ 0 & otherwise \end{cases} \tag{4}$$

The symbol rate is related to the IF bandwidth as

$$R_{T_s} = \frac{B_{IF}}{1+\propto} \tag{5}$$

Note that for both roll-off factor $\alpha = 1$ and $\alpha = 0.25$, the shaping functions are unity at $t = 0$ and zero at the integer multiples of the sampling period. Furthermore, note that for smaller

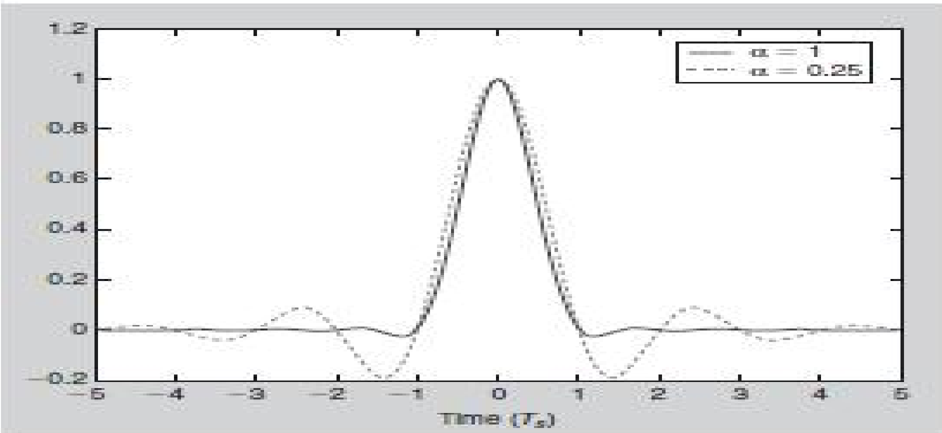


Fig.(6) Normalized time-domain raised cosine response for α =1 and α =0.25

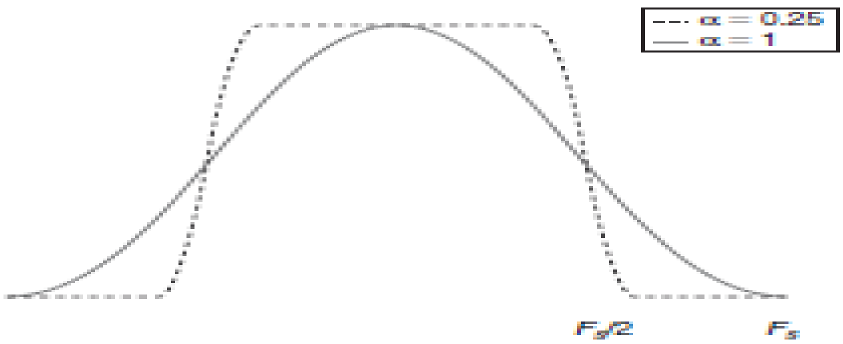


Fig.(7)Frequency domain representation of the raised cosine pulse for α =1 and α =0.25

The shaping function tends to have more contained spectrum at the expense of more oscillations in its response thus requiring a more robust timing recovery algorithm. The raised cosine pulse oscillations in general, however, decay at the rate of 1/ t 3 for t Tsas compared with the normal sinc function of the brick-wall filter. This rapid decay in magnitude response allows the designer to truncate the pulse response without any significant degradation to the performance. Another important observation from Figure 7 is excess bandwidth with respect to the Nyquist frequency. For roll off factor $\alpha = 1$ the excess bandwidth is 100% or twice the Nyquist frequency, and for $\alpha = 0.25$ the excess bandwidth is 25%. Hence, α dictates the bandwidth of the filter. The null itself is placed at $(1 + \alpha)Fs/2$ as suggested in (4) and is dictated mainly by the amount of adjacent channel interference that can be tolerated by the system. In many practical systems, a sin(x)/ x correction 3 is applied to equalize for the sinc like behavior of the square pulse, thus modifying the relation presented in **Equation (4)** to:

$$G_m(f) = \begin{cases} \dfrac{\pi f T_s}{\sin(\pi f T_s)} & |f| \le (1-\alpha)\dfrac{F_s}{2} \\ \dfrac{\pi f T_s}{\sin(\pi f T_s)} \cos^2\left[\dfrac{\pi T_s}{2\alpha}\left(f - \dfrac{1-\alpha}{2T_s}\right)\right] & (1-\alpha)\dfrac{F_s}{2} \le |f| \le (1+\alpha)\dfrac{F_s}{2} \\ 0 & otherwise \end{cases} \qquad (6)$$

Furthermore, the raised cosine pulse may be realized as a combination of transmit root raised cosine filter in the transmitter and an identical root raised cosine filter in the receiver where the cosine squared term in (6) is replaced with its square root to realize the root raised cosine pulse

3. Simulation Results

In the transmitter, the generated signal from the random integer generator is modulated by the 16-QAM modulator, which has a symbol rate of 2.5 M symbol/s. The ideal eye figure of the generated 16-QAM baseband signal is shown in **Figure (8),** whereas the constellation diagram is shown in **Figure (9).** The modulated signal is subsequently up-converted by a factor of 16 and pulse-shaped by the Root Raised Cosine (RRC) filter.

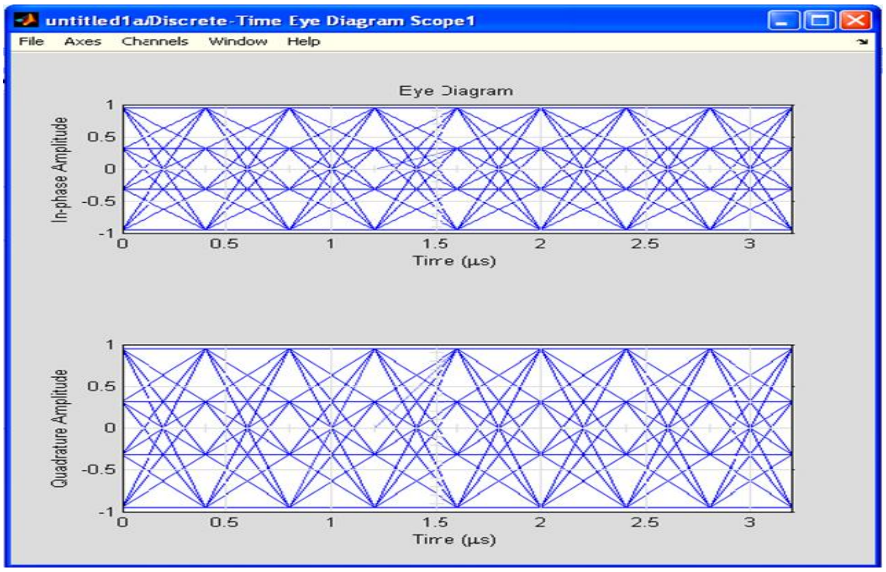


Fig.(8) Eye diagram of the generated Baseband Signal

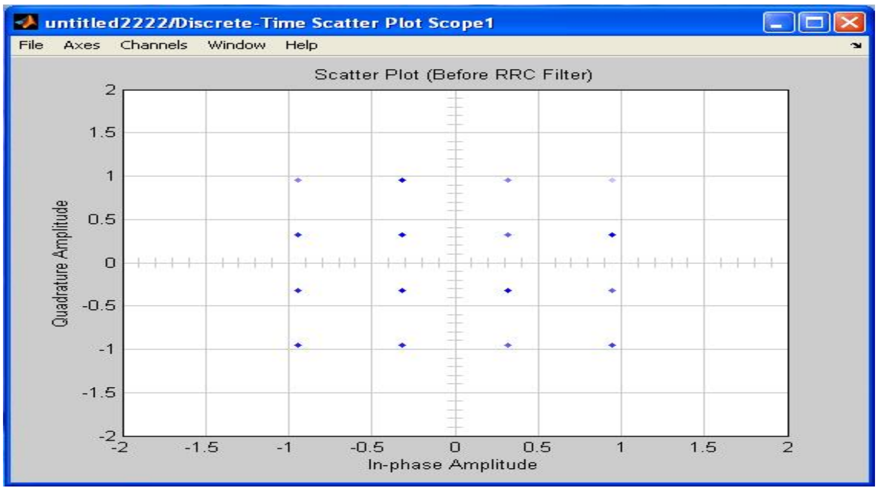


Fig.(9) Constellation diagram of the generated Baseband Signal

3.1 Interpolation (Up-Sampling)

Generally,it is useful to modify the effective variety speed of an existing sampled signal. A sample rate of a signal is increased by interpolation. Conceptually, interpolation comprises the generation of a continuous curve passing through old samples, followed by sampling the curve at the new sample rate to obtain the interpolation sequence. To increase a given sample rate or up-sample, a factor of M, M-1 intermediate values between each samples in the old signal, has to be calculated. The interpolation increases the sampling frequency (Fs) to:

$$f_s = f_{new} = M f_{old} \qquad (7)$$

For the proposed system, the 16-QAM baseband signal is up-sampled by a factor of 16 to produce a new sample rate of 40 M symbol/s. The interpolation process produces inherent amplitude loss factor of M. This loss factor is compensated by adding gain stage in the system to achieve unity gain between the old and the new sequences.

3.2 Pulse Shaping (Root Raised Cosine Filter)

The core of any wireless communication system is the use of transmitter and receiver pulse shaping filters. The requirements of Root Raised Cosine filter (RRC) is shown in **Table (3).**

Table 3: Design Requirement of Root raised cosine Filter

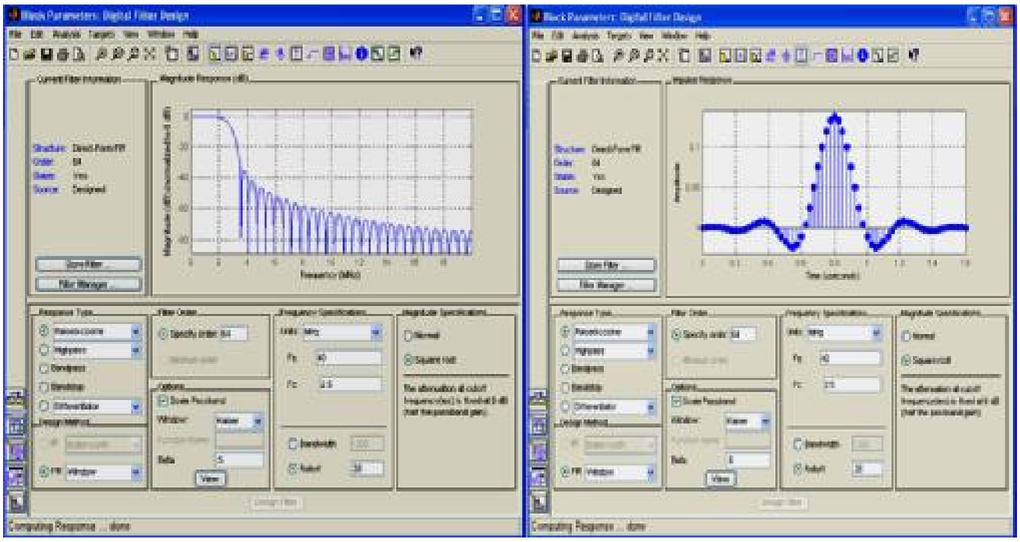
Centre frequency f_{centre}	2.5 MHz
Sampling rate	40 MHz
Stop-band attenuation	-40 dB
Roll-off factor β	0.35
Phase Response	Linear
Stability status	Stable

In order to achieve these requirements, a FIR filter using Kaiser Window is chosen to design the filter. There are three significant frequency points associated with the raised cosine response. The first is known as: Nyquist frequency, which is equal to the centre frequency. The second significant frequency point is the stop band frequency (f_{stop}), defined as the frequency at which the response first reaches zero magnitude. The third, and the final significant frequency point is the pass band frequency (f_{pass}) defined as the frequency at which the response first begins to depart from its peak magnitude.

$$f_{stot} = (1+\beta)f_{center} = (1+0.35)X\ 2.5MHz = 3.375MHz \tag{8}$$

$$f_{pass} = (1-\beta)f_{center} = (1-0.35)X\ 2.5MHz = 1.625MHz \tag{9}$$

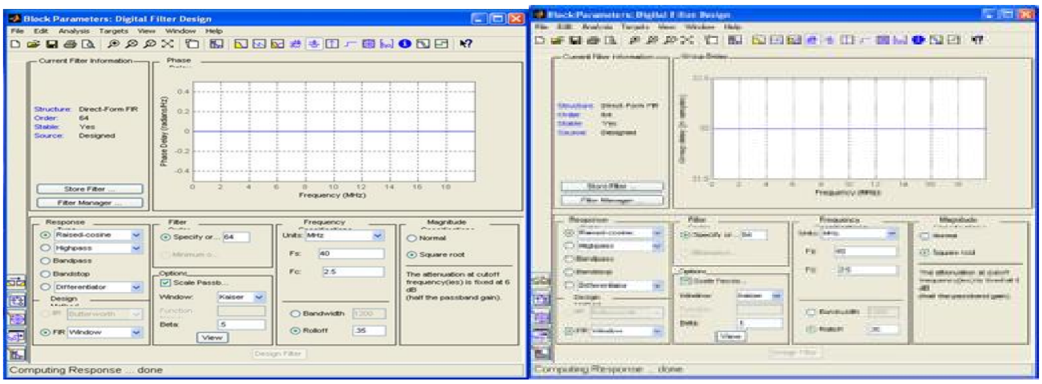
The magnitude and impulse responses of RRC filter are shown in **Figure 10.** The characteristics of the filter indicate that the filter is non-distorting.



(a) (b)

Fig.(10) (a):Magnitude response RRC filter, (b): Root Raised Cosine Filter impulse response

The phase response is linear within the bandwidth of interest and the group delay is constant within the bandwidth as shown in **Figure (11)**. Linear phase refers to the condition where the phase response of the filter is linear (straight line) function of the frequencies. This results in the delay though the filter being the same at all frequencies. Therefore, the filter does not cause phase distortion or delay distortion.



(a) (b)

Fig.(11) (a): Root Raised Cosine Filter phase delay, (b): Group delay response

The Interpolation Filter is developed as root-raised cosine (RRC) pulse shaping filter (PSF) in polyphase FIR interpolator architecture as presented. There are two multi-rate interpolation filters (MRIFs) or 2xMA FIR Multi Interpolator subsystems, each for I and Q channels. The MRIF design parameters are listed in Table (4). Note that cut-off frequency f_{co} is set to symbol rate R_{sym} but not half symbol rate $R_{sym}/2$ because it is the minimum bandwidth for 4-PAM and 16-QAM transmissions without loss of information after DAC. The FDATool block is used to set only the design parameters of the highest M-array modulation i.e. 16-QAM. The total latency of MRIF (D_{MRIF}) is 6. The outputs I_out and Q_out are converted to fixed-point: signed 16-bit, binary point: 13 for DAC process..

Table 4: Design Parameters of Multi-Rate Interpolation Filter (MRIF)

Parameter	BPSK	4-PAM, QPSK	16-QAM
Filter Taps, N	9	17	33
Filter Order, $M = N-1$	8	16	32
Interpolation Rate, L	4	8	16
Cutoff frequency, $\omega_{co} = 2\pi/L$ (rad/sample)	0.5π	0.25π	0.125π
Filter Type	RRC	RRC	RRC
Roll-off Factor, α	0.5	0.5	0.5
Window	Kaiser	Kaiser	Kaiser
Window Parameter, β	0.5	0.5	0.5
Scaled Passband to 0 dB	Yes	Yes	Yes
Normalized Coefficients	Yes	Yes	Yes

The optimized design parameters of polyphase interpolator in the MRIF are listed in Table (5). Over phase tap O is the number of extra taps per phase as compared to integer multiple of L, so $O = 1$ indicates that only first phase component ($\varnothing = 1$) has extra 1 tap and requires 1 extra MA operation. Thus, the MA operation of phase $\varnothing = 1$ is designed to work separately, in order to eliminate extra 0-value coefficients and multipliers used in the successive phase sub-filters for fully standardized MA operation. The total latency of this subsystem (D_{MRIF}) is 6. The T_s is changed from 4 to 1 (= 4/4) unit sample time. The mask parameters of this subsystem and the associated internal subsystems (including parameter settings) are described in **Table (5).**

Table 5: Design Parameters of Polyphase Interpolator of MRIF

Parameter	BPSK	4-PAM, QPSK	16-QAM
Symbol Period, T_{sym} (unit sample time)	4	8	16
Interpolation Rate, L	4	8	16
Sampling Period, $T_s = T_{sym}/L$ (unit sample time)	1	1	1
Filter Taps, N	9	17	33
Over Phase Taps, $O = N$ (mod L)	1	1	1
Optimized Filter Taps, $N_o = N-O$	8	16	32
Optimized Phase Taps, $q_o = N_o /L$	2	2	2

The original spectrum of the baseband signal is processed under the 20dB AWGN channels as shown in **Figure (12).** The power of the baseband magnitude signal is 2.5dB.

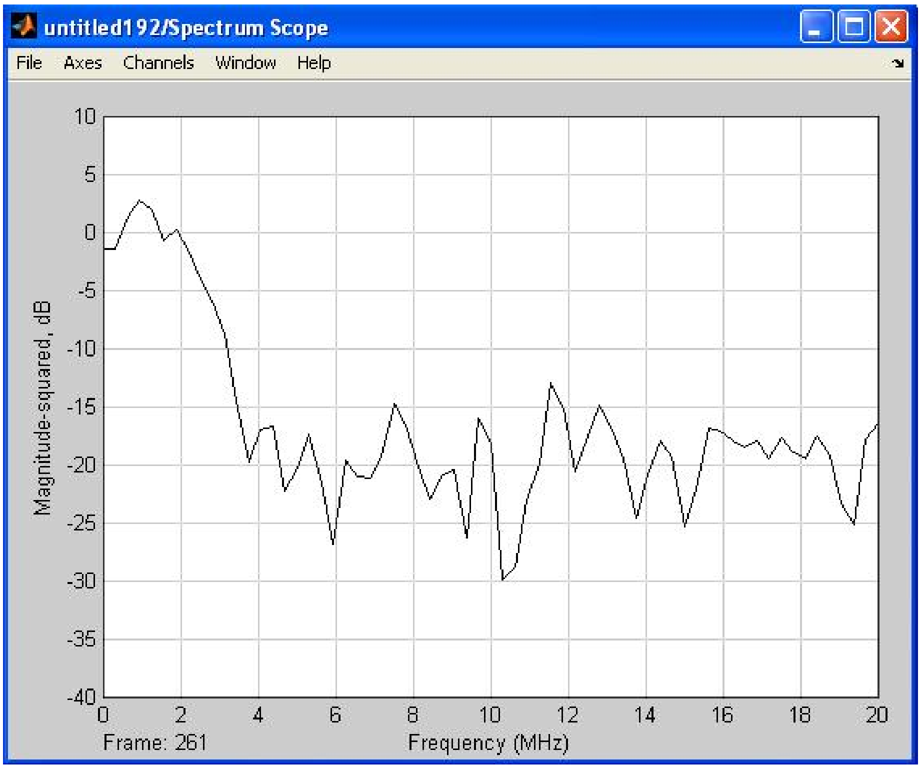


Figure 12: Original Baseband Signal

4. Conclusion

In this paper the design of Software Defined Radio (SDR) transmitter model is presented. For the design of SDR model in Matlab/Simulink, the transmitted signal with 2.5 MHz bandwidth undergoes 20 dB SNR (Signal-to-Noise Ratio) in Additive White Gaussian Noise (AWGN) channel. Therefore this develop on 3G of wireless and mobile systems more economical, effective and accelerate the transition from 3G to 4G of wireless systems. The 16-QAM scheme uses 4-bits per symbol to represent 16 possible symbol values and two simultaneous pulses to encode the symbol. Therefore, for 16QAM, each pulse must be able to assume one of four levels because, two pulses with four possible levels per pulse give 16 possible combinations. These four possible pulses levels are 1/3,-1/3, A max and -Amax. A quadrature amplitude modulation is a commonly used scheme in multi-bit- pulse-encoding which relies on two mechanisms for encoding bits, the pulse amplitude and two simultaneous pulses. The pulse amplitude could be assumed to has both positive and negative values, while the two simultaneous pulses require two independent baseband channels, one referred to in-phase (I), and the other for quadrature (Q) channel. Qudrature Amplitude Modulation (QAM) comes in a variety of forms depending on the number of bits encoded into each pair of pulses.

The software defined radio transmitter presented in this paper, is not fully complete. Only the IF and baseband stage of the wireless communication system has been implemented.

Therefore, further research could focus on an end-to-end implementation of a wireless communication system in software defined radio. The channel is considered to have only additive white Gaussian noise. Doppler shift, phase error, multipath fading, etc., can be added to the channel, in order to closely simulate real life systems.

References

1. **Jeffry. H. Reed, "*Software Radio a modern approach to radio engineering*", Prentice Hall PTR, A division of person education Inc, ISBN 0-13-081158-0, pp.33-12, 2002**
2. **Munro A. "*Mobile middleware for the reconfigurable software radio*" . IEEE Communications Magazine August 2008;38(8):152-61.**
3. **Pallavi M. Mannan "*Frame Work for the Design and Implementation of Software Defined Radio based Wireless Communication System",* M.Sc. thesis, University of Akron: http://etd.ohiolink.edu/view.cgi/Mannar%20Manna**
4. **M.S. Naghmash ,M.F. Ain, C.Y.Hui,"*FPGA Implementation of SDR Model based 16 QAM,"* Eurojournais publishing ,Inc,vol.35.no.2,2009.**
5. **A.Zafor and Z.Farooq,"I*mplementation and Analysis of Qpsk & 16 QAM Modulator & Demodulator,*" vol.2,P.64 -68, Norember ,2008.**
6. **Xiaolong Li, *"Simulink-based Simulation of Quadrature Amplitude Modulation (QAM) system ,"* Proceeding of the International Conference ,2008.**
7. **Thomos, C.; Kalivas, G. , "*FPGA-based architecture of a DS-UWB Channel Estimator and RAKE Receiver employing a hybrid selection scheme,*" 2010 IEEE 17th International Conference on Telecommunications (ICT), pp.903-909, April 2010.**
8. **Himanshu Shekhar, C.B.Mahto and N.Vasudevan, "*FPGA Implementation of Tunable FFT For SDR Receiver",* IJCSNS International Journal of Computer Science and Network Security, Vol.9 No.5, May 2009.**