A method of compensating distortion of radio signals with amplitude-phase modulation in a quadrature driver

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Abstract. The article considers the features of spectrally effective radio signals and the principles of signal formation by differential quadrature amplitude modulation with a shift of $\pi / 4$. A method is proposed for compensating for distortion of radio signals with amplitude-phase modulation in a quadrature driver. A distortion compensation scheme is modeled, and the possibility of its application for non-linear amplification of signals with a variable envelope is considered.

1. Introduction
Currently, most modern digital radio standards use quadrature modulation signals, the principles of formation and reception of which are based on quadrature transformations. The absence of controllable reactive elements and frequency selective circuits allows QPSK quadrature phase shift keying, MSK minimum frequency shift modulation, GMSK pre-modulation Gaussian filtering, $\pi / 4$-DQPSK spectrally effective modulation, FQPSK (Feer modulation) and other varieties in a wide range of carrier frequencies digital modulation [1]. In systems with amplitude-phase modulation, during the interval of transmission of one elementary signal, the amplitude and phase can take values selected from a number of possible discrete values of amplitudes and phases [2]. For hardware implementation of quadrature modulation, signal processors of the main frequency band and vector modulators are widely used to transfer the spectrum of the modulating signal to the region of intermediate or carrier frequencies. The task of the signal processor is to form such a transfer function of the modem that would ensure smooth transitions of the digital modulating signal to reduce out-of-band emissions and increase the spectral efficiency of radio signals.

2. Features of $\pi / 4$ –DQPSK signal generation
One spectrally effective signal is a differential quadrature amplitude modulation signal with a shift of $\pi / 4$ ($\pi / 4$ –DQPSK). Figure 1 shows a simulation circuit of a radio signal former with $\pi / 4$-DQPSK modulation. The high-frequency sinusoidal oscillation by the carrier generator “V_1Tone” is fed to the input of the modulator model “PI4DQSK”, and the pseudo-random bit sequence of pulses is supplied by the voltage source “VtLFSR_DT”. The output signal of the modulator “PI4DQSK” at the intermediate frequency passes through the bandpass (smoothing) filter “BPF_RaisedCos” limiting out-of-band emissions. The signals pass through the filter in such a way that, with maximum compression, the signal distortion bands are minimal. Such a filter is a filter with a characteristic in the form of an “raised cosine” [3].
After passing through the smoothing filter (figure 1), the signal is added with the noise component, the result can be visualized and the quality of the modulated signal can be monitored using the “signal constellation”.

Figure 2a shows the ideal “signal constellation” of the π / 4-DQPSK radio signal at α = 1, and figure 2b shows the “signal constellation” with jitter at α = 0. It can be seen how a decrease in the smoothing coefficient affects the increase in the value of intersymbol interference.

The main characteristics for the π / 4-DQPSK signal when it is transmitted over the air are the frequency spectrum and the envelope amplitude. The signal constellations shown in figure 2 (a, b) describe I / Q states that are in stable positions in time only at symbol points. Between these moments of time there are transitions. The signal constellation for the encoded signal π / 4-DQPSK shows that the trajectory does not go through the point (0,0), the minimum phase shift, after each transmitted symbol an additional phase shift of π / 4 (45 °) is performed, which limits the maximum phase shifter 135 °. This ensures that the transition between the two characters does not cross the origin and that only one carrier amplitude is always transmitted.

The complex envelope of the signal is represented by the expression $A(t) = A(t)e^{j\phi(t)}$. The instantaneous value of the amplitude $A(t)$ can be represented as a vector emerging from the origin, and the phase $\phi(t)$ as the angle between the vector and the positive direction of the abscissa axis. Thus, in the process of modulation, the vector performs rotational-vibrational motion around the origin. The trajectory of the end of the vector, when modulated by a random sequence of dibits, describes the envelope of significant amplitude modulation, which is a fundamental element of the π / 4-DQPSK signal.

![Figure 1. π / 4-DQPSK Modulated Radio Shaper Modeling Scheme.](image)

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![Figure 2. Modulation π / 4-DQPSK: a) the ideal "signal constellation" of the radio signal π / 4-DQPSK; b) π / 4-DQPSK signal constellation; c) π/4-DQPSK signal spectrum.](image)

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Figure 2c shows the spectrum of the π / 4-DQPSK signal, modulated by a sequence of rectangular pulses, having a spectrum in the form of the function $\sin(x)/x$, but this spectrum is shifted to the carrier frequency. It can be seen from the figure that the π / 4-DQPSK signal spectrum is compact. When
implementing the device, hardware errors associated with the imperfection of individual links in the path are inevitable, therefore, the real spectrum is somewhat different from the theoretical one.

Figure 3 is a block diagram of the \( \pi / 4 \)-DQPSK base modulator. The data stream nonreturn-to-zero (NRZ) arriving at the modulator is divided into two separate NRZ streams using a series-parallel converter.

![Figure 3. Block diagram of the \( \pi / 4 \)-DQPSK modulator.](image)

From the original digital stream \( C_m \), the sequentially parallel encoder forms two parallel streams of dibits (symbols) \( A_k \) and \( B_k \), while \( [A_k, B_k = C_{2k-1}, C_{2k}] \). Symbol period \( T_s = 2T_b \), where \( T_b \) is the bit period.

The coding device from the flows of dibits \( A_k, B_k \) forms modulating sequences of in-phase (I) and quadrature (Q) channels, separately fed to the multipliers after low-pass filters. A carrier signal \( \cos \omega t \) is supplied to the first input of the channel multiplier I, and to the second input of the channel multiplier Q is the quadrature carrier, i.e. 90° phase-shifted signal (\( \sin \omega t \)).

The in-phase and quadrature components \( i_k, q_k \) which arrive at the balanced modulators, are as follows:

\[
\begin{align*}
i_k &= i_{(k-1)} \cos [\Delta \varphi_k(A_k, B_k)] - q_{(k-1)} \sin [\Delta \varphi_k(A_k, B_k)], \\
q_k &= i_{(k-1)} \sin [\Delta \varphi_k(A_k, B_k)] + q_{(k-1)} \cos [\Delta \varphi_k(A_k, B_k)],
\end{align*}
\]

(1)

(2)

the values of \( \Delta \varphi_k \) are defined for the corresponding values of \( A_k, B_k \) as:

\[
\Delta \varphi = \begin{cases} 
-3\pi/4 & \text{for } A_k = 1, B_k = 1 \\
3\pi/4 & \text{for } A_k = 0, B_k = 1 \\
\pi/4 & \text{for } A_k = 0, B_k = 0 \\
-\pi/4 & \text{for } A_k = 1, B_k = 0 
\end{cases}
\]

Then, the output is summed to obtain a four-phase signal. Thus, for each \( T_s \), the carrier phase is updated using the following relation for the signal when its power is \( S_0 \):

\[
S_0(t) = \sqrt{2S_0} \cos(\omega t + \varphi_k) = \sqrt{2S_0} \cos(\omega t + \varphi_{k+1}) = \sqrt{2S_0} \cos(I_k \cos \omega t + Q_k \sin \omega t),
\]

(3)

where \( I_k = \cos \varphi_k \), a \( Q_k = \sin \varphi_k \).

The phase transition is independent of the current phase of the carrier and has a maximum value of \( \pi / 4 \), which is less than the maximum phase shift created by OQPSK (QPSK with offset). The carrier phase transitions created with \( \pi / 4 \)-DQPSK are shown in figure 4.

Ensuring acceptable energy and spectral efficiencies is the most important requirement for digitally modulated mobile radio systems. In most cases, manufacturers of electronic equipment to amplify digital radio signals use non-linear amplifiers, which leads to the expansion of the output spectrum and the appearance of spurious interference on an adjacent channel. In this case, it is necessary to increase the channel spacing, which leads to a decrease in spectral efficiency. The developers of electronic equipment are constantly trying to find a solution to this problem, however, there is no ideal compromise between increasing spectral efficiency and reducing energy efficiency. To use non-linear amplifiers with high
efficiency at their input, the presence of a constant envelope is required, which for the above types of radio signals is variable (variable). In addition, one of the drawbacks of circuits using quadrature modulators is the possibility of out-of-synchronization of the quadrature components, in connection with which a method for compensating distortion of radio signals with amplitude-phase modulation in the quadrature driver was proposed.

3. Modeling of a method for compensating distortion of radio signals in a quadrature driver

To consider the method of compensating for distortion of radio signals in a quadrature driver with amplitude-phase modulation in the Advanced Design System, the $\pi / 4$-DQPSK signal conversion path was modeled.

Figure 5 shows a simulation diagram. The “Data” element is a source of a random NRZ data stream with a speed of 32 $kBit/s$ and a bit duration $T_b$, which enters the serial-parallel converter “SymbolSplitter” and is divided into two streams $I$ and $Q$. Four different phase states that form the serial dibit-to-symbol conversion circuit is stored during the signal interval $T_s$ which is equal to the duration of two bits $T_s = 2T_b$ [1]. The generated sequences with a symbol rate of half the speed of the input bit stream are sent to the “EncoderIQ” element. EncoderIQ performs differential encoding of data in accordance with the $\pi / 4$-DQPSK format. Under the initial conditions, $I_0 = 1$ $Q_0 = 0$, $A_k$ and $B_k$ denote the logical states of the input data $I$ and $Q$. In addition, the differential nature of the circuit means that the first transmitted symbol is a reference and does not transmit any information. Differentially encoded data pass through a smoothing filter “LPF_RaisedCosine” with a characteristic in the form of a raised cosine [3]. The high-frequency generator “N_Tones” generates a sinusoidal carrier signal $f_c$ at a frequency of 1 MHz, and the phase shifter “PhaseShiftRF” performs a shift by $\pi/2$ and converts it into a cosine signal.

$$S_1(t) = S_1 \cos \omega t,$$
$$S_2(t) = -S_2 \sin \omega t,$$  \hspace{1cm} (4)

We represent the smoothed quadrature components as follows.

$$I(t) = A_C(t) \cos \varphi,$$
$$Q(t) = A_S(t) \sin \varphi,$$  \hspace{1cm} (5)

The standard signal $\pi / 4$-DQPSK will be obtained by modulating the filtered in-phase $I(t)$ and quadrature $Q(t)$ components in the quadrature modulator “QAM_Mod”.

To bring the $\pi / 4$-DQPSK waveform to a constant envelope, it is necessary to generate narrow-band phase modulation signals with a modulation index of $m \leq 0.5$ [6]. Due to the fact that the index of phase modulation generated by quadrature modulators will be determined by the instantaneous values of the amplitudes of the signals $I(t)$ and $Q(t)$, the “Gain” elements weaken them. An increase in the phase modulation index more than $m \geq 0.5$ leads to the appearance of spurious amplitude modulation.
The ideal sine and cosine functional converters “Cos” and “Sin” receive filtered signals with a nonlinear structure I (t) and Q (t). The output signals of the functional converters $Q_{cos}(t)$, $Q_{sin}(t)$, as well as $I_{cos}(t)$, $I_{sin}(t)$ are correlated.

\[ Q_{cos}(t) = U_q cos \{ Q(t) \}, \quad Q_{sin}(t) = U_q sin \{ Q(t) \}, \quad I_{cos}(t) = U_i \arctg[I(t)], \quad I_{sin}(t) = U_i \sin[I(t)]. \tag{6} \]

Mutual correlation can be expressed as:

\[ Q_{cos}(t) = \sin(\arccos[Q_{sin}(t)]), \quad I_{cos}(t) = \sin(\arccos[I_{sin}(t)]) \tag{8} \]

**Figure 6.** Timing diagram of the quadrature component of the signal at the output of the functional converters.

Figure 6 on an enlarged scale in the time diagram shows the values of the quadrature and in-phase components of the signal at the output of the functional converters. A correlation of the cosine and sinusoidal transformations is observed. The maxima $Q_{cos}(t)$ and $Q_{sin}(t)$ at points 1, 3, 6, 8 correspond to the maxima $I_{cos}(t)$ and the average values of $I_{sin}(t)$. Highs $I_{cos}(t)$ and $I_{sin}(t)$ at points 2, 4, 5, 7 correspond to the maxima of $Q_{cos}(t)$ to the average values of $Q_{sin}(t)$. Thus, the synchronization of the quadrature components of the modulator. The output signals of the functional converters are modulating signals for the quadrature modulators “QAM Mod”, and the necessary condition for the formation of the correct waveforms is the multiplication of $Q_{sin}(t)$, $I_{sin}$ by the sinusoidal high-frequency component, and $Q_{cos}(t)$, $I_{cos}(t)$ by the cosine. At the output of the quadrature modulators, phase-modulated radio signals with a constant envelope $u_Q(t)$ and $u_I(t)$, are formed, which allows them to use a highly efficient non-linear mode of operation of the power amplifier.

\[ u_Q(t) = U_q \cdot S_1 (cos\omega t cos[Q(t)] - sin\omega t sin[Q(t)]) = S_Q \cos(\omega t + [Q(t)]), \tag{9} \]

\[ u_I(t) = U_i \cdot S_2 (cos\omega t cos[I(t)] - sin\omega t sin[I(t)]) = S_I \cos(\omega t + [I(t)]), \tag{10} \]

where $S_Q = U_q \cdot S_1$ and $S_I = U_i \cdot S_2$ - constant envelope.

As a result of combining two phase-modulated signals, we obtain the resulting signal $E(t)$

\[ E(t) = u_Q(t) + u_I(t) = S_Q \cos(\omega t + [Q(t)]) + S_I \cos(\omega t + [I(t)]) = \]

\[ = 2S_Q S_I \cos \frac{2\omega t + I(t) + Q(t)}{2} \cos \frac{I(t) - Q(t)}{2}. \tag{11} \]

Figure 7a shows the time diagram of the π / 4-DQPSK signal and the linearized component of the π / 4-DQPSK signal at the input of a nonlinear power amplifier with a gain of $G = 5$ and a third-order intersection point (TOI) = 30 dB. Figure b shows the timing diagrams of the signals at the output of the amplifier. One can observe the presence of pronounced distortions of the amplified signal π / 4-DQPSK.
Figure 7. Timing diagram of the $\pi/4$-DQPSK signal: a) at the input of a nonlinear amplifier and a linearized component of the signal; b) at the output of the amplifier.

Figure 8 shows the spectrum of the $\pi/4$-DQPSK signal and the linearized signal $E(t)$ at the input (a) and output (b) of a nonlinear power amplifier. It can be noted that the spectra of the signals are identical until amplification. At -60 dB, the $\pi/4$-DQPSK signal spectrum increases by 64 kHz relative to the linearized signal $E(t)$.

Figure 8. The spectrum of the $\pi/4$-DQPSK signal and the linearized signal $E(t)$: (a) at the input and (b) output of a nonlinear power amplifier.

4. Conclusion
As a result of the analysis of the method for compensating distortions of amplitude-phase-modulated radio signals in a quadrature driver using the $\pi/4$-DQPSK signal as an example, we can conclude that this method allows correlation of the quadrature components of the modulator to synchronize and minimize errors during the modulation process. The formation of signal components with phase modulation and a constant envelope allows the use of highly efficient non-linear modes of operation of a power amplifier.

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