A Simple Signal Shaper for GMSK/GFSK and MSK Modulator Based on Sigma-Delta Look-up Table

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Abstract. *Due to wide power spectrums of rectangular data streams, it is important for base-band signals to be heavily band limited before modulation. That can be achieved by pulse shaping of rectangular bits. Some of the most common are a half-sine pulse shaper and a Gaussian pulse shaper which are used in Minimum Shift Keying (MSK), Gaussian Minimum Shift Keying (GMSK) and Gaussian Frequency Shift Keying (GFSK) modulations, respectively. The most common solutions of such shapers use PCM based look-up-table (LUT), which requires an n-bit D/A converter. We proposed the use of a 1-bit Sigma Delta Modulation (SDM) LUT, which results in smaller ROM capacity, a 1-bit wide output word, and a simple1-bit D/A converter realized as an out-of-chip first-order low-pass RC filter, or an in-chip charge pump. This article describes a simple, but efficient SDM LUT-based half-sine and Gaussian shaper that can be used for generation of MSK and GMSK/GFSK modulated signals. Oscillograms and power spectrums are measured on SDM LUT realized in FLEX Altera^{TM} PLD, for a 10-bit pseudo-noise sequence test input signal.*

Keywords

GMSK/GFSK, MSK, sigma-delta, Gaussian filter.

1. Introduction

In today's modern digital high speed communication systems primary consideration is to achieve modulation with power spectrum of acceptable bandwidth and constant amplitude of the modulated signal. Some of the most efficient modulation techniques are MSK and GMSK/GFSK [1-3]. They are part of the Constant Phase Frequency Shift Keying (CPFSK) modulation family with a constant envelope. Since the modulated signal has constant amplitude, an efficient RF amplifier of class C can be used to minimize power consumption, an important consideration for battery-powered units. In MSK two generic techniques for modulation and demodulation of MSK are referred to as parallel and serial (direct) methods. In direct synthesis method of realization, MSK is derived as ordinary FSK with the modulation index set to 0.5. The modulation index of 0.5 corresponds to minimum frequency spacing that allows two FSK signals to be coherently orthogonal, and the name *minimum shift keying* implies the minimum frequency separation that allows orthogonal detection.

An MSK signal with I-Q components is formed by passing the modulation signal through a half-sine shaper before modulation. GMSK/GFSK modulation uses the same technique as MSK but instead of the half-sine pulse shape, input bits have the shape of the Gaussian bell curve. Such shapers are usually realized with different analog (8th order Bessel filter for the Gaussian shaper) or digital PCM LUT [4], [8] based circuits.

This paper proposes an efficient pulse shaping method based on the SDM [5], [6] LUT and its implementation for direct MSK and GMSK synthesis. Section 2 explains basics of MSK and direct GMSK/GFSK modulation, while Section 3 describes circuit implementation of the SDM LUT-based shapers. Section 4 gives experimental results, and the conclusions are given in Section 5.

2. MSK and GMSK/GFSK Modulation Basics

Both modulations, MSK and GMSK/GFSK, are derived from the ordinary Frequency Shift Keying (FSK) modulation scheme, which is a digital version of frequency modulation (FM). An FM signal is defined as:

$$u_{FM}(t) = U_m \cos\left[\omega_C t + \varphi(t)\right], \tag{1}$$

where U_m is the amplitude, ω_c is the carrier frequency, and $\varphi(t)$ is the phase of FM signal, which is for FSK equal to:

$$\varphi_{FSK}(t) = 2\pi \frac{m}{T_b} \int_0^t \sum_{i=0}^{\infty} b_i g_{RECT}(\tau - iT_b) d\tau, \tag{2}$$

where m is the modulation index, T_b a symbol interval and $g_{RECT}(t)$ the pulse shape function. In ordinary FSK the digital signal that modulates an FM modulator is a rectangular bipolar Non Return to Zero (NRZ) bit sequence with symbol values $b_i \in \{-1, 1\}$. Definition of a rectangular pulse shape is given by (3) and it is shown in Fig. 1, while the spectrum is shown in Fig. 2.

$$g_{RECT}(t)=\begin{cases} U_m, & -\dfrac{T_b}{2}\le t\le\dfrac{T_b}{2} \\ 0 & otherwise. \end{cases} \qquad (3)$$

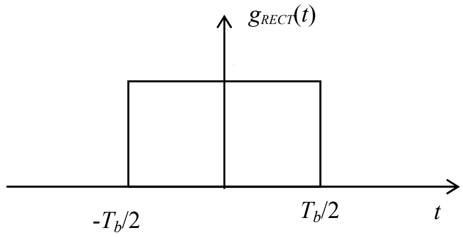


Fig. 1. Rectangular pulse shape.

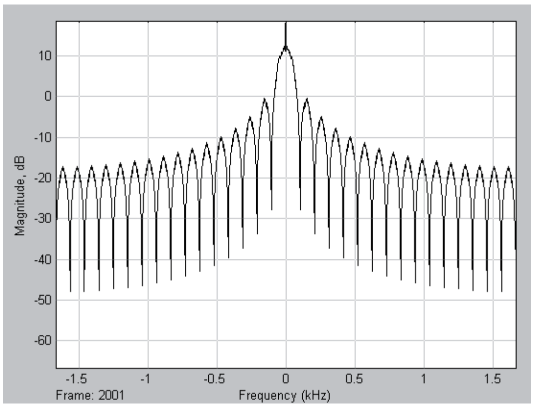


Fig. 2. Power spectrum of the rectangular non-shaped data sequence.

2.1 MSK Modulation Basics

MSK [7] is a continuous phase modulation scheme. The modulated carrier does not contain phase discontinuities and frequency changes at carrier zero crossings. It is typical for MSK that the difference between the frequency of logical 0's (f_0) and 1's (f_1) is equal to half the data rate. MSK modulation makes the phase change linear and limited to $\pm(\pi/2)$ over the symbol interval. Due to the linear phase change effect, better spectral efficiency is achieved. That means that MSK is ordinary FSK with the modulation index set to 0.5, and it is defined as:

$$m=\Delta f T_b \qquad (4)$$

where peak frequency deviation Δf is given by

$$\Delta f = | f_1 - f_0 | . \qquad (5)$$

The MSK modulator can be realized by using a direct MSK approach or the I-Q based concept. In both types of modulators the straightforward means of reducing the Out Of Band (OOB) energy is pre-modulation filtering or pulse shaping. Direct MSK modulation can be realized by direct injection of NRZ data into the frequency modulator with the modulation index set to 0.5 (Fig. 3a).

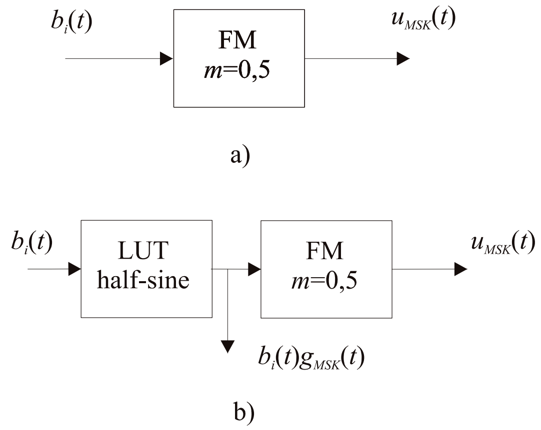


Fig. 3. Direct MSK modulator a) without and b) with pulse shaping.

The spectrum of the direct MSK modulator output is not compact enough to realize common data rates for the RF channel bandwidth (B). Because of that, pulse shaping is of particular interest (Fig. 3b). Data input sequence is forwarded to a shaping filter whose output pulse shape is given by (6) and is shown in Fig. 4.

$$g_{RECT}(t)=\begin{cases} U_m \cos\left(\dfrac{\pi t}{T_b}\right), & -\dfrac{T_b}{2}\le t\le\dfrac{T_b}{2} \\ 0 & otherwise. \end{cases} \qquad (6)$$

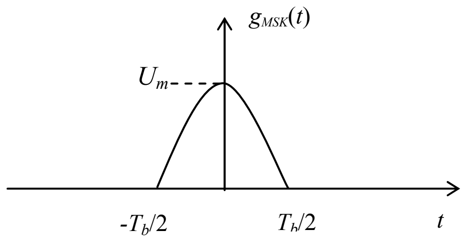


Fig. 4. Half-sine pulse shape.

The resulting pulse shaped sequence (Fig. 5) is then applied to the FM modulator whose output is a constant amplitude continuous phase FM signal (MSK signal $u_{MSK}(t)$). The resulting MSK signal can be written as (1).

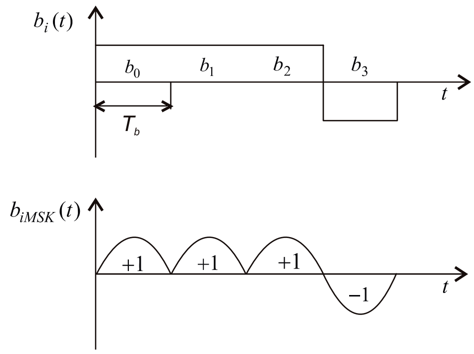


Fig. 5. Direct MSK bit streams $2T_b$.

Reduction of OOB energy obtained by the half-sine pulse shaping can be seen in Fig. 6.

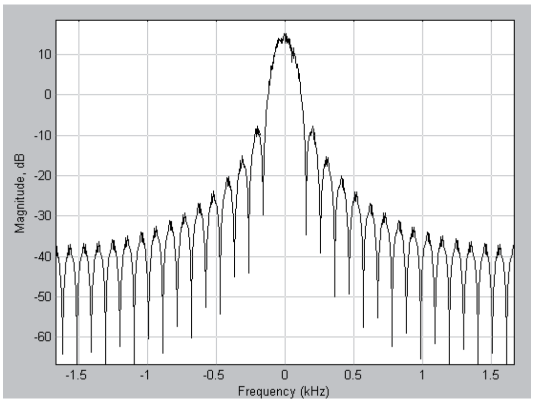


Fig. 6. Power spectrum of half-sine pulse shaped sequence.

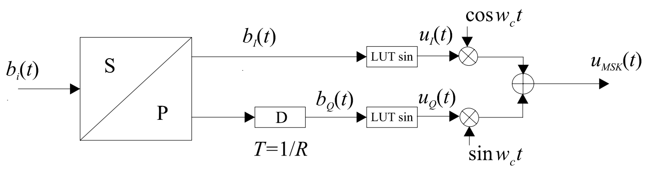


Fig. 7. OQPSK LUT based I-Q MSK modulator.

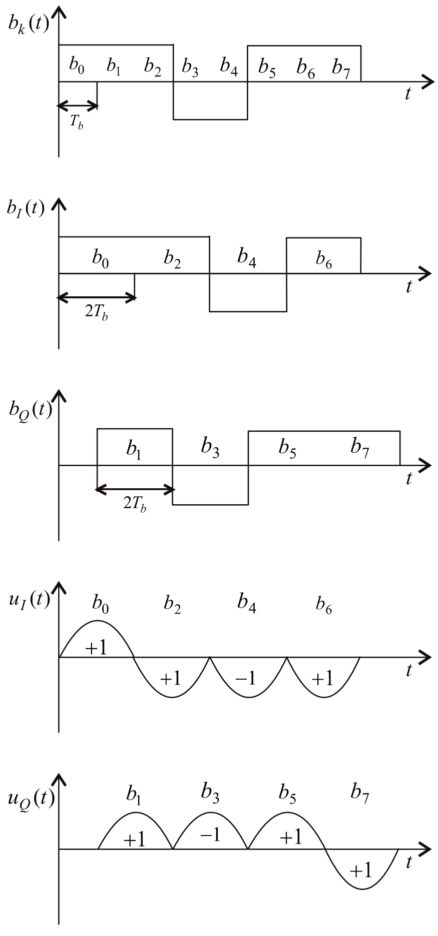


Fig. 8. I-Q based MSK bit streams.

The phase of the MSK signal is given by (7):

$$\varphi_{MSK}(t) = 2\pi \frac{m}{T_b} \int_0^t \sum_{i=0}^{\infty} b_i g_{MSK}(\tau - iT_b) d\tau . \qquad (7)$$

Fig. 7 shows an LUT based concept of I-Q MSK realization derived from Offset Quadrature Phase Shift Keying (OQPSK) [7]. The input data stream, which arrives to the modulator at the rate of $R_s = 1/T_b$ bits/sec, separates into two data streams $b_I(t)$ and $b_Q(t)$, containing odd and even bits respectively, with the rate $R_p = 1/(2T_b)$. OQPSK is obtained by delaying the odd bit stream by a symbol interval T_b with respect to the even bit stream (I and Q streams). If these two streams are offset by one symbol interval, amplitude fluctuations become minimized since the phase always changes by ±90°.

The MSK signal is derived by replacing the OQPSK rectangular data streams pulses used in QBPSK with half-sine pulses. In that way I and Q components of the MSK signal $u_I(t)$ and $u_Q(t)$ become:

$$u_I(t) = b_I(t) g_{MSK}(t) ,$$

$$u_Q(t) = b_Q(t) g_{MSK}(t - T_b), \qquad (8)$$

and the MSK signal is defined as:

$$u_{MSK}(t) = b_i(t) g_{MSK}(t) \cos(\omega_c t) + b_Q(t) g_{MSK}(t) \sin(\omega_c t). \quad (9)$$

2.2 GMSK/GFSK Modulation Basics

GMSK/GFSK modulation can be realized by both parallel [8] and serial synthesis [9]. It differs from the ordinary MSK by using the Gaussian LP filter or Gaussian shaper on the input of the I-Q or FM modulator. Fig. 9 shows a basic GMSK/GFSK modulator.

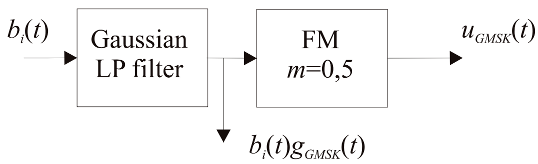


Fig. 9. Basic GMSK/GFSK modulator.

Minimization of the spectral bandwidth of the output signal $u_{GMSK}(t)$ for the NRZ input sequence can be realized by filtering with the Gaussian LP filter, whose name came from impulse response function $h_{GAUSS}(t)$ shown in Fig. 10.

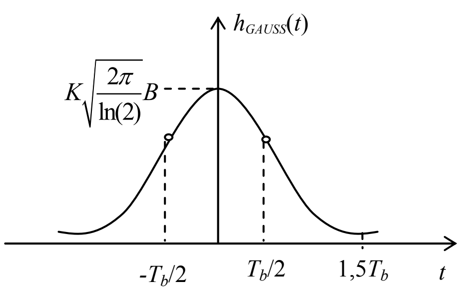


Fig. 10. Impulse response of the Gaussian pulse shaping filter.

Impulse response of the Gaussian pulse shaping filter is given by:

$$h_{GAUSS}(t) = K\sqrt{\frac{2\pi}{\ln(2)}}Be^{-2\frac{(B\pi)^2}{\ln(2)}t^2} . \quad (10)$$

Gaussian shaped bit stream $g_{GMSK}(t)$, which is equal to convolution of $g_{RECT}(t)$ and $h_{GAUSS}(t)$ becomes:

$$g_{GMSK}(t) = -\frac{K}{2\sqrt{\ln(2)}}\left[Erf(2B\pi\frac{t-\frac{T_b}{2}}{\sqrt{\ln(4)}}) - Erf(2B\pi\frac{t+\frac{T_b}{2}}{\sqrt{\ln(4)}}) \right] \quad (11)$$

where $Erf(t)$ is error function [10] given by

$$Erf(x) = \frac{2}{\sqrt{\pi}}\int_0^x e^{-t^2}dt . \quad (12)$$

From Fig. 11 it can be seen that the main characteristic of the Gaussian filter is a BT_b product, where B is a –3 dB bandwidth of the Gaussian filter, and T_b is a previously defined symbol interval. Product BT_b determines the pulse shape of the output bit stream. A lower BT_b product implies lowering amplitude and increasing the pulse width.

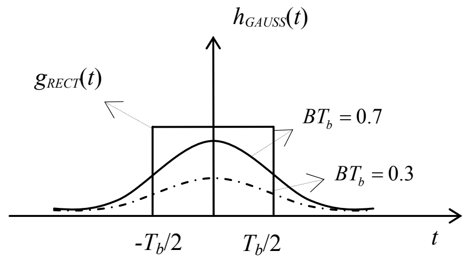


Fig. 11. Gaussian pulse shape for two BT_b products.

The power spectrum of the shaped data sequence is shown in Fig. 12. Comparison of this spectrum with spectrums in Figs 6 and 2 shows that the Gaussian shaped pulse sequence has better base-band spectrum performance (low OOB energy).

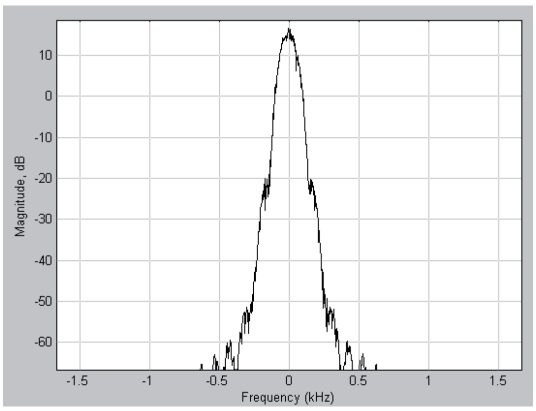


Fig. 12. Power spectrum of the Gaussian pulse shaped sequence.

The resulting Gaussian pulse shaped sequence $g_{GMSK}(t)$ is then applied to the FM modulator, resulting by data phase $\varphi_{GMSK}(t)$:

$$\varphi_{GMSK}(t) = 2\pi\frac{m}{T_b}\int_0^t \sum_{i=0}^{\infty} b_i g_{GMSK}(\tau - iT_b)d\tau . \quad (13)$$

3. Circuit Implementation

Shapers shown previously can be efficiently realized as ROM look-up tables (LUTs). Required waveform shapes are sampled to get their n-bit PCM representation for the implementation in the LUT. Our proposition is to replace PCM with Sigma Delta Modulation (SDM). It requires a simpler 1-bit D/A converter (a low-pass *RC*-filter or a charge pump), instead of the complex (out- or in-chip) n-bit D/A converter, and lower ROM LUT capacity. In order to get SDM LUT data for each waveform shape, we have built a MATLAB model (Fig. 13) of the Gaussian shaper, followed by the first-order SDM. The same model is also used for the half-sine shaper, excluding the Gaussian LP filter.

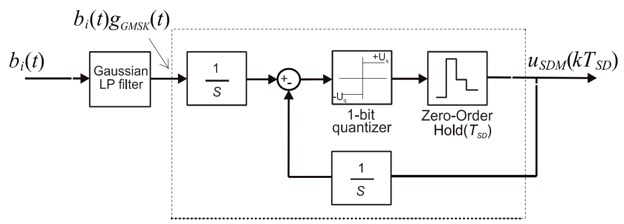


Fig. 13. MATLAB model of the Gaussian shaper .

One of the main problems in the SDM application is so-called *slope overload*. Fig. 13 shows that Gaussian shaped signal is integrated in the SDM input. Generally, slope overload occurs when the input signal slope is greater than the maximal slope of the SDM integrator. The condition for the slope overload avoiding [11] is

$$\left|g_{GMSK}(t)\right|_{\max} \le \frac{U_h}{T_{SD}} = \frac{kU_h}{T_b} = U_q , \quad (14)$$

where $|g_{GMSK}(t)|_{\max}$ is the maximum of the Gaussian pulse, U_h/T_{SD} is the maximal slope of the SDM integrator, $U_h=U_qT_{SD}$ is the integration quant, T_{SD} is the SDM pulse period (sampling time), U_q is the amplitude of the quantizer, and $k = T_b/T_{SD}$. The maximal amplitude of the Gaussian shaped pulse is achieved for $t = 0$:

$$\left|g_{GMSK}(t=0)\right|_{\max} = \left|\frac{K}{\sqrt{\ln(2)}}Erf\left[\frac{BT_b\pi}{\sqrt{2\ln(2)}}\right]\right|_{\max} = U_m , \quad (15)$$

where $U_m = K/\sqrt{\ln(2)}$ is its voltage equivalent. According to [14], the slope overload is avoided for any voltage U_m lower than U_q. For the $BT_b < 1$, the given condition holds even more:

$$U_m \le \frac{kU_h}{T_b} . \quad (16)$$

The same calculation for avoiding slope overload can be also carried out for the half-sine shaper:

$$\left|g_{MSK}(t)\right|_{\max} = \left|U_m\cos(\frac{\pi t}{T_b})\right|_{\max} = U_m \le \frac{kU_h}{T_b} . \quad (17)$$

The maximum achievable slope for both input signals is $U_m = 5$V and according to (15) and (17), slope overload is avoided. To achieve the power signal-to-noise ratio (*PSNR*) over 53 dB, k is chosen to be 128. The PSNR of the first-order SDM is defined by the following equation [14]:

$$PSNR = 6.02b + 1.76 - 5.17 + 30\log(OSR);\,[\text{dB}], \quad (18)$$

where the number of bits b is equal to 1 and the oversampling ratio *OSR* is $f_{SD}/2f_b$. In our case, *OSR* was taken to be equal to:

$$OSR = \frac{f_{SD}}{2f_b} = \frac{T_b}{2T_{SD}} = \frac{k}{2} = 64, \quad (19)$$

that gives *PSNR* = 56.79 dB, which is, according to the requirements for the pulse shaping techniques proposed in [15], equal to 8-bit PCM resolution.

3.1 MSK SDM Pulse Shaper

Digital part of the SDM LUT based MSK pulse shaper and a low-pass *RC* filter as a D/A converter are shown in Fig. 14a. A half-sine shape is written in serial ROM LUT as a sequence of 64 bits. For the i^{th} bit of the input data sequence, $b_i(t)$, those 64 bits are serially read-out from ROM LUT by clock pulses ($f_{Cp} = 64/T_b$). If the bit is negative, a ROM LUT sequence is inverted by EX-OR function. Because of the half-sine shape symmetry, the number of samples can be halved, increasing a little bit complexity of read-out circuitry, but reducing further on ROM capacity. It should also be noted that except small capacity ROM, such design does not require an n-bit D/A converter (additional reduction of design complexity!).

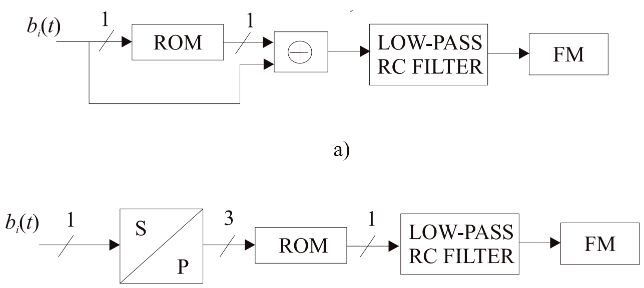


Fig. 14. Functional scheme of the SDM LUT based a) half-sine pulse shaper, and b) Gaussian pulse shaper.

3.2 GMSK SDM Pulse Shaper

The Gaussian pulse shaper can also be realized by SDM ROM LUT. Theoretically, impulse response of the Gaussian filter extends from -∞ to +∞, but in practice, the extent of the impulse response is usually considered to be limited only to the two nearest neighbor data sequence bits: $\{b_{-1}, b_0, b_{+1}\}$. Individually shaped pulses of those three consecutive bits are added together. Fig. 15 shows the resulting curve for the data sequence: 101, and Tab. 1 gives all eight possible curves for all three bit combinations.

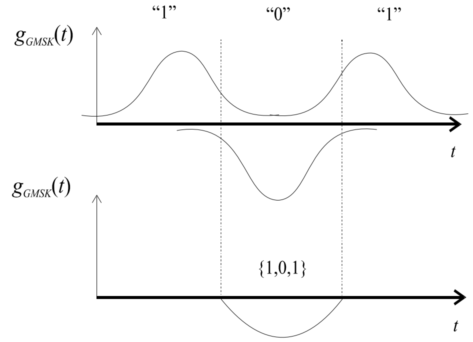
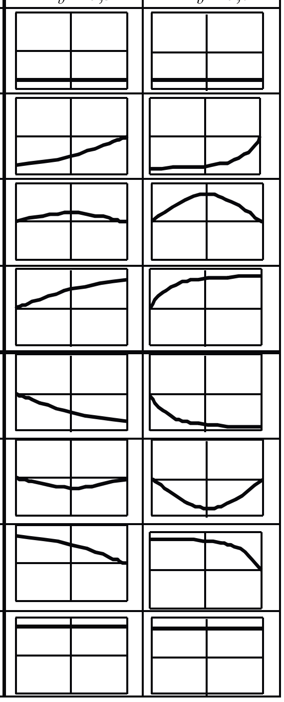


Fig. 15. Individually shaped pulses and resulting curve for three consecutive bits: 101.

A MATLAB model of the Gaussian filter, followed by the first-order SDM, is used for digitalization of curves (Tab. 1). Inputs to the Gaussian filter were bipolar NRZ pulses. For each three-bit combinations on the input of the filter, one 64-bit long sequence of a 1-bit wide SDM sampled curve is stored for the later implementation into the LUT. Two types of the Gaussian filter have been created, one for $BT_b = 0.3$ and the other for $BT_b = 0.7$.

b_{-1}, b_0, b_{+1}	$BT_b = 0{,}3$	$BT_b = 0{,}7$
000		
001		
010		
011		
100		
101		
110		
111		



Tab. 1. Eight possible curves for all combinations of three consecutive bits ($BT_b = 0.3$, $BT_b = 0.7$).

Digital part of the SDM LUT based Gaussian pulse shaper is shown in Fig. 14b. A serial-to-parallel shift register forms three-bit combinations of the input serial data stream $\{b_{-1}, b_0, b_{+1}\}$, which are then used to address ROM LUT. LUT capacity can be simply calculated as 8 curves x 64 1-bit samples = 512 bits. A similar pulse shaper is proposed in [8], but based on the 6-bit PCM code-words stored in ROM. Its implementation requires 4 curves x 64 samples x 6 bit = 1536 bits of the ROM capacity. Comparing these two approaches, the SDM based LUT proposed in this paper gives 8-bit PCM resolution, requires 3-times lower ROM capacity and does not require any complex (6-bit or higher resolution) D/A converter at the output of LUT. Only the first-order low-pass RC filter is enough to obtain the analog signal for further FM modulation.

Since complementary three-bit combinations result in complementary curves (see Tab. 1), LUT capacity can be reduced further on. Thus, with a little bit more complex read-out circuitry we can get lower ROM capacity (256 bits), or keep it the same (512 bits), but with 9 dB higher *PSNR,* increasing the OSR from 64 to 128.

4. Experimental Results

Experimental results were obtained by the measuring set shown in Fig. 16. Output of the half-sine or Gaussian pulse shaper feeds FM modulator - signal generator Agilent 33250A. Test results are captured by oscilloscope Agilent 54622D and spectrum analyzer Agilent E4402B.

Digital circuits and ROM LUT implemented in Altera[TM] EPF10K20RC240-4 [12], 1-bit DAC and the first-order low-pass *RC* filter, represent the Gaussian shaper or the half-sine shaper. Output of the RC filter feeds the FM modulator for generation of the GMSK or MSK modulated signal. Analog switches (1-bit DAC) are inserted between digital output of the shaper and the RC filter to convert a 1-bit unipolar into the 1-bit bipolar signal. Two sets of Gaussian curves (for BT_b=0.3 and BT_b=0.7) are stored in the ROM, which then consumes in total of 2 x (8 x 64) = 1024 bits.

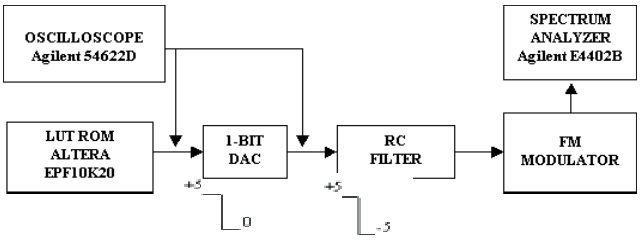


Fig. 16. Test set.

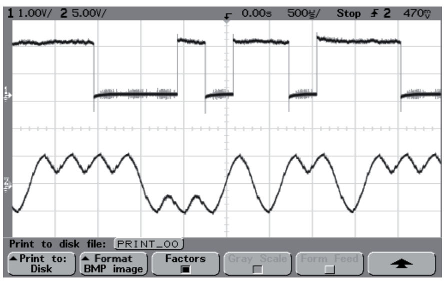


Fig. 17. 10-bit pseudo-noise test sequence (upper trace) and half-sine shaped pulses (lower trace).

Upper trace oscillogram in Fig. 17 shows part of a rectangular test data sequence (10-bit pseudo-noise sequence internally generated in PLD). This data sequence represents data input stream $b_i(t)$, to the half-sine pulse shaper and it is generated at data frequency of $f_b = 1/T_b$ = 3 kHz. Lower trace shows half-sine shaped pulses at the output of the first-order low-pass *RC*-filter.

Oscillogram in Fig. 18 shows the same rectangular test sequence and Gaussian shaped pulses at the output of the LP *RC*-filter for the case of $BT_b = 0.7$.

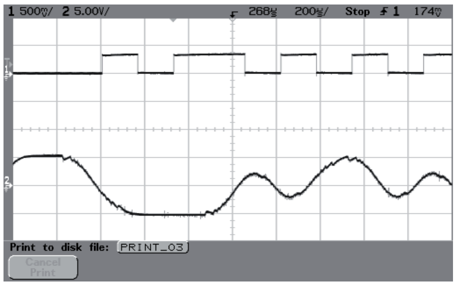


Fig. 18. 10-bit pseudo-noise test sequence (upper trace) and Gaussian shaped pulse (lower trace).

Fig. 19 shows an eye diagram of the MSK pulse shaper, which basically consists of two half-sine curves. As can be seen from the eye diagram in Fig. 19, there is no Inter Symbol Interference (ISI) between two consecutive bits.

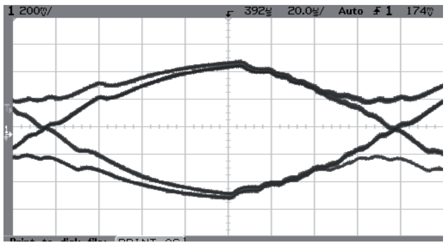
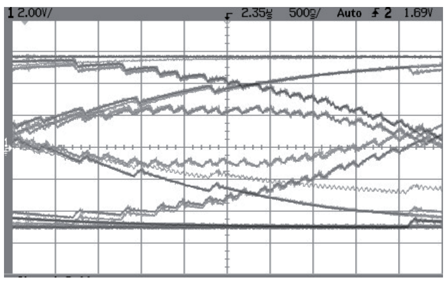
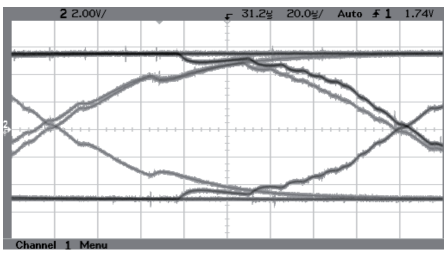


Fig. 19. Eye diagram.



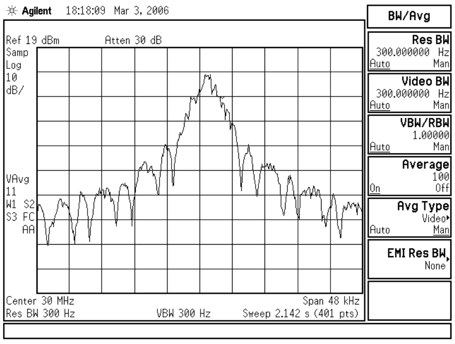
a)



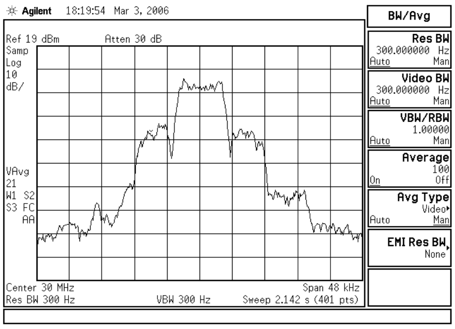
b)

Fig. 20. Eye diagrams for a) BT_b product equal to 0.3, and b) BT_b product equal to 0.7.

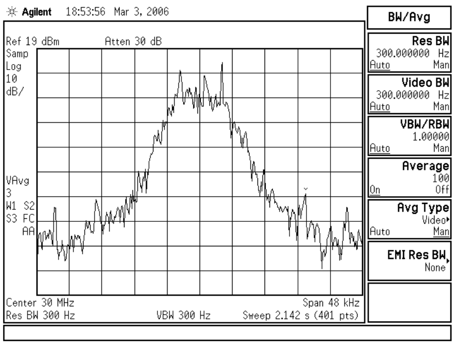
Fig. 20 shows an eye diagram of the GMSK pulse shape which consists of eight basic curves from Tab. 1. Those eye diagrams are captured for BT_b = 0.3 and BT_b = 0.7, respectively. As can be seen from Fig. 20, the eye diagram is closed more when the product BT_b is smaller, i.e. bandwidth of the Gaussian filter is narrower.



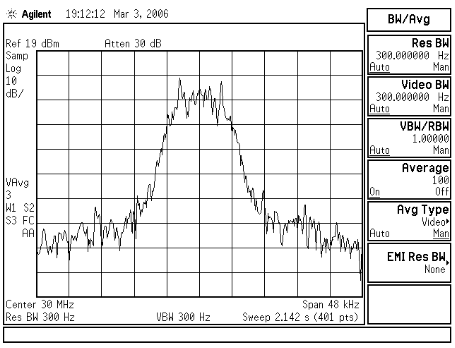
a)



b)



c)



d)

Fig. 21. Pass-band spectrums of a) direct MSK without pulse shaping b) direct MSK with pulse shaping c) direct GMSK with BT_b = 0.7, and d) GMSK with BT_b = 0.3.

Gaussian filtered pulses have narrower spectrum, i.e. a lower OOB part of the spectrum, than half-sine pulses. Better OOB is also achieved by lower product BT_b of the Gaussian filter.

Due to the small OOB it is obvious that modulation of the high frequency carrier with Gaussian shaped signals causes a narrower pass-band spectrum. To confirm that we have modulated carrier frequency of f_c = 30 MHz by a half-sine data sequence and obtained a direct MSK signal. Then, we used a Gaussian shaped data sequence and obtained direct GMSK modulation. Frequency deviation of the used FM modulator was set to Δf = 1.5 kHz, which implies a modulation index equal to m = 0.5. Results shown in Fig. 21 are matched to theoretical results [13].

5. Conclusion

MSK and GMSK/GFSK are digital modulation techniques used for digital links when bandwidth conservation and the use of efficient amplitude saturating transmitters are important requirements. Those modulation techniques are based on pulse shaping. Different pulse shapers are the base of modern high-speed digital systems. Due to its complex analog realization, the interest in digital circuit solution has gained increased popularity.

The contribution of this work is efficient hardware implementation of SDM LUT based sine and the Gaussian rectangular pulse shaper and 1-bit realization which requires less ROM LUT resources and asks for simpler D/A conversion.

In this paper we showed that the proposed pulse shaper based on the SDM LUT provides 3-times less ROM capacity than standard 6-bit PCM realization. At the same time, using symmetry of three-bit combinations of Gaussian curves, ROM LUT capacity can be reduced (halved) further on, or *PSNR* can be increased (9 dB) further on by keeping ROM LUT capacity unchanged. The proposed solution not only makes ROM capacity economical, but it also eliminates the need for a complex n-bit PCM D/A converter. Since we use a 1-bit wide SDM word, the simple out-of-chip first-order low-pass RC filter or in-chip charge pump can be used as 1-bit D/A converters.

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