where $S(f)$ and $W(f)$ are the Fourier transforms of the transmitter and receiver windows, respectively. For MSK with matched filtering, (7) reduces to

$$E\{S^2\}/E\{J^2\} = \pi^2/8A^2.$$

For TFM, the signal-to-jammer ratio is given by [3]

$$E\{S^2\}/E\{J^2\} = \left\{ \begin{array}{ll} 4 \left[ \int_{-T/2}^{T/2} A(f)Z(f) df \right]^2, \\
+ 8 \sum_{k=1}^{\infty} \left[ \int_{-T/2}^{T/2} A(f)Z(f) df \right]^2 \cos 2\pi f/T d\tau \right\} \left[ A_2^2 A_2^2(0)/2 \right].$$

If (5) holds, the above reduces to

$$E\{S^2\}/E\{J^2\} = Z_2^2(0)/A^2 = 1/A^2.$$

Comparing the performance of MSK demodulated with a matched receiver window and TFM with a receiver satisfying (5), we find that MSK is about 0.9 dB less sensitive to tone jamming than TFM. This is as expected because TFM has a much smaller noise bandwidth [1] than MSK, making it more sensitive to tone jamming.

VII. CONCLUSIONS

We have evaluated the adjacent channel crosstalk in coherent TFM systems. The excellent spectral efficiency of TFM was confirmed, utilizing mean-square crosstalk as a performance measure. TFM, as expected, was found to have much better crosstalk performance than MSK, e.g., TFM is 27 dB better than MSK at a frequency separation of 1.5 $R$. The crosstalk results were then used to build an equivalent white Gaussian noise model to predict performance degradation of TFM. The crosstalk equations were used also to show that TFM is about 0.9 dB more sensitive to tone jamming than MSK.

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New Universal All-Digital CPM Modulator

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Abstract—In this paper a new modulator for a broad class of continuous phase modulations (CPM) is proposed that is more flexible and less complex than conventional quadrature modulators, while achieving the same or better performance. Being all-digital it is more suitable for one-chip VLSI implementation. Some design parameters are also discussed and experimental results are presented.

I. INTRODUCTION

Continuous phase modulation (CPM) signaling schemes have gained interest recently because of their narrow spectrum due to the smoothed phase transitions, their constant envelope which permits the use of efficient nonlinear amplifiers and their good noise immunity.

For CPM systems, the modulated signal may be written as

$$s(t, \alpha) = \sqrt{2E \over T} \cos \left\{ 2\pi f_0 t + 2\pi T \sum_{i=0}^{\infty} \alpha_i g(\tau - iT) d\tau + \phi_0 \right\}$$

where $\alpha_i = \pm 1, \pm 3, \ldots, \pm (M-1)$ are the elements of an infinite $M$-ary information symbol sequence, $T$ is the symbol time, $E$ the symbol energy, $f_0$ the carrier frequency, $\phi_0$ the arbitrary initial phase shift, $h$ the modulation index and $g(t)$ is a baseband frequency pulse which determines the modulation type. In order to have an unambiguous meaning for $h$, the following normalization has been adopted:

$$\int_0^{LT} g(t) dt = 1/2,$$

where $LT$ is the (finite) length of the frequency pulses. Different pulses $g(t)$ lead to the various CPM modulations like

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Fig. 1. Conceptual CPM modulator.

Fig. 2. Quadrature CPM modulator.

From (1) it can be seen that any CPM signal may be generated by driving a VCO with the control voltage

\[ v(t) = h \cdot \sum_{i=-\infty}^{\infty} \alpha(t-iT). \]  

However, this modulator, depicted in Fig. 1, is not practical because conventional free-running VCO's cannot achieve either acceptable frequency stability or the linearity required for low distortion and for some receiver synchronization schemes which rely on precise control of modulation index (i.e., De Buda loop [3]).

As a practical solution for the CPM generation problem a quadrature modulator has been proposed [8], with baseband quadrature signals \( I(t) \) and \( Q(t) \), in the interval \( nT \leq t \leq (n+1)T \), defined as the cosine and sine of the value \( \theta(t, \alpha) + \theta_n \), where \( \theta(t, \alpha) \) depends on the last \( L \) symbols and \( \theta_n \) represents the modulator state

\[ \theta_n = \frac{\pi}{h} \sum_{i=-\infty}^{-1} \alpha_i. \]  

For rational modulation indexes \( h \) of the form \( p/q \) (\( p \) and \( q \) integers), there is a finite number of possible "phase states" \( \theta_n \), so that \( J \) samples of \( I(t) \) and \( Q(t) \) per symbol interval \( T \) may be generated from sine and cosine ROM's addressed by \( \theta_n \), and the last \( L \) symbols \( \alpha_i \), as in Fig. 2. Such a quadrature modulator has thus far been successfully built for TFM [6], while in the case of GMSK a special VLSI chip has been designed for generating the signals \( I(t) \) and \( Q(t) \) [9].

Although a quadrature modulator may be quite easily adapted for various modulation schemes through changing the ROM's contents, the mixture of digital and RF techniques does not easily lead to single-chip implementations. In order to reduce the carrier-leak, image and other spurious components to an acceptable level, careful mixer design and precise amplitude and phase balances are required.

Other reported CPM modulators include the "serial" MSK modulator [10], [11], where the MSK signal is obtained by passing BPSK through a properly designed transversal SAW or stripline filter, and an all-digital CPFSK modulator with just one D/A converter for full response (\( L = 1 \)) signaling [12] (as a CPFSK signal during one symbol interval is just one of \( M \) sinusoids, samples of all these sinusoids are stored in a ROM, and a special sequencer addresses the appropriate memory locations, sequencing them in such a manner that no phase jumps arise after the symbols change). The possibility of generating GTFM signals with one DAC has also been mentioned in [7].

II. DIRECT DIGITAL MODULATOR

The modulator which we propose here may be regarded as a direct digital realization of the scheme in Fig. 1, with \( G(jw) \)
replaced by a digital FIR filter $G(z)$ and the VCO by a numerically-controlled-oscillator (NCO).

The NCO, which can be identified as one block in Fig. 3, is just a digital synthesizer as proposed by Tierney, Rader, and Gold in 1971 [13]. It consists of an accumulator with output

\[ n^{(j)} = \{k + n^{(j-1)}\} \mod N, \]

where $k$ is the value of the (binary) number on its input, and one ROM with the samples of just one sinusoid so that the content of the $i$th memory location is

\[ f(i) = \text{int}\left(2^{Nq-1} \cos (2\pi i/N)\right) \] (6)

In (6) int ( ) means the nearest integer, and $N_q$ is the number of bits per location in the ROM, i.e., the D/A converter resolution.

It is not difficult to check that the frequency of the sinusoidal signal at the output of DAC, after appropriate filtering, is given by

\[ f_{\text{NCO}} = k \cdot K_{\text{NCO}} \leq N/2 \]

where

\[ K_{\text{NCO}} = f_s/N, \]

and the constraint in (7) comes from the sampling theorem.

Apart from being extremely stable, with very low phase noise, this synthesizer has the following features which are important in CPM modulator applications:

1) After changing $k$, the NCO naturally preserves the phase continuity;

2) the NCO's "sensitivity" $K_{\text{NCO}}$ depends only on $f_s$ and $N$, so that it can be ultrasensitive;

3) by augmenting $N$, the NCO's resolution may be increased without limit (since the sine ROM has capacity $N \times N_q$, this may cause an enormous increase of its capacity, but a simple modification described in [14] allows us to cover all cases of practical importance with 2-4 K address locations, and with some additional logic this may be further reduced by a factor of 2 or 4 [13]);

4) if the NCO's frequency does not vary considerably with respect to $f_s$, the signal envelope is practically constant.

The block marked NCO in Fig. 3 is, therefore, by itself a generator of a full-response ($L = 1$) CPFSK signal, without any sequencer and with just one memorized sinusoid.

In order to generate more complex CPM waveforms, it is necessary to provide for premodulation shaping preceding the NCO. In principle, this filtering can always be accomplished by an FIR filter, but in practice it is usually much simpler to memorize the last $L$ symbols and use them as the addresses for $J$ samples of one of $M^L$ possible frequency paths (it should be noticed that the capacity of this ROM is smaller than the capacity of the sine or cosine ROM in the quadrature modulator of Fig. 2 by a factor equal to the number of states $\theta_n$, and it is not necessary to have a third ROM for recursion on $\theta_n$.

When a ROM is used as an FIR filter, it is necessary to adequately scale the frequency pulse $g(t)$ in order to control the modulation index $h$. The discrete case of (2) can be represented as

\[ h = 2K_{\text{NCO}}(T/J) \sum_{i=1}^{J} g_i, \]

which implies that $h \cdot J/2K_{\text{NCO}} T$ should be an integer. Since in practice it is necessary to use some $K$th multiple of $T/J$ for $f_s$, we have actually the condition that $hN/2K$ should be an integer, which can be satisfied for any rational value of the modulation index $h = p/q$ by choosing the appropriate $N$ and $K$. If all these requirements are fulfilled, and a correct discretization of $g(t)$ is made, then the generated modulation will have the exactly the desired modulation index and there will be no accumulated phase error.

Before proceeding to discuss the design problems associated with realization of the proposed direct digital modulator, it is worth noting that the NCO has been initially described as a synthesizer capable of generating frequencies from zero to somewhat below $f_s/2H$ with $K_{\text{NCO}}$ Hz resolution. Since the maximal deviations for CPM modulation are rather small, we can also provide the channel raster somewhat greater than double the value of maximal deviation. If the system in which the proposed modulator has to be incorporated requires just a relatively small number of channels, the modulator can act as a synthesizer as well, and if so many channels are required that $f_s$ becomes impractically high, it is always possible to incorporate this synthesizer/modulator in a hybrid PLL configuration [15] where it provides for both modulation and fine resolution, while the high reference PLL expands the covered bandwith and filters the noise away from the carrier which arises in all ROM and DAC-based realizations.

III. DESIGN CONSIDERATIONS AND EXPERIMENTAL RESULTS

In practical realizations of the proposed modulator certain compromises have to be made with respect to:

1) number of samples $J$ per symbol interval $T$;

2) number of quantization levels $N_q$ of frequency pulse $g(t)$;

3) number of symbols $L$ after which a truncation of the pulse $g(t)$ is performed in the case of SRC, GMSK,
TFM and other time-infinite frequency pulse modulations;
4) resolution $N_o$ of the output DAC.

By increasing any of these parameters better performance is achieved, but complexity, price and power consumption also increase.

Although the first three tradeoffs must be made in a quadrature modulator as well as in this realization, no general analysis has been made thus far because of the nonlinear nature of the modulation process. It is not easy to derive analytical results even with a single source of quantization, and any combined effects analysis is much more difficult. Since the parameters are influenced by the chosen modulation format as well, they are usually determined by means of simulation. A typical simulation example is given in Fig. 4, and shows the spectrum of $M = 2$, $h = 0.5$, $3RC$ modulation (which has been chosen because there are no truncation effects) with respect to the number of quantization levels $N_q$. It can be seen that $N_q = 2^6$ is sufficient to have the noise down at $-80$ dB.

Due to price and power limits, however, in practical realizations it is rather difficult to make $J$, $N_q$, $L$, and $N_o$ sufficiently large in order to keep the overall noise level below

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**Fig. 4.** Spectra of $3RC$ modulation ($M = 2$, $h = 0.5$) for various number of quantization level $N$, 1) $N = 10$, 2) $N = 20$, 3) $N = 40$, 4) $N = 60$, and 5) $N = \infty$.

**Fig. 5.** Spectra of GMSK $M = 2$, $h = 0.5$ for: 1) $B_sT = 0.5$, 2) $B_sT = 0.3$, 3) $B_sT = 0.2$ (bit rate 16 kbits/s, ver. 10 dB/div, hor. 10 kHz/div, res. 1 kHz).
of such a modulator are briefly discussed, and measurements on a model are presented which confirm the new concept.

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