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N92-24168

Considerations of Digital Phase Modulation for Narrowband Satellite Mobile Communication

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ABSTRACT

The Inmarsat-M system[1] for mobile satellite communication is specified as a FDMA system, applying Offset QPSK for transmitting 8 kbit/sec in 10 kHz user channel bandwidth.

In this paper we consider Digital Phase Modulation, DPM[2], as an alternative modulation format for Inmarsat-M. DPM is similar to Continuous Phase Modulation, CPM[3], except that DPM has a finite memory in the premodulator filter with a continuous varying modulation index.

It is shown that DPM with 64 states in the VA obtains a lower Bit Error Rate, BER.

Results for a 5 kHz system, with the same 8 kbit/sec transmitted bitstream, is also presented.

Introduction

The specifications for the Inmarsat-M system is based upon the use of OQPSK as modulation format, combined with convolutional codes to obtain the wanted BER of the speech and data services. OQPSK is selected due to its spectral properties when exposed to a nonlinear transmitting chain.

As an alternative to OQPSK, ELAB-RUNIT has been considering DPM for the same transmission specifications, as those defined for OQPSK. This is a required adjacent interference level, under specific relative power levels between the centre channel and the two neighbours.

In this paper, we first review DPM. The performance of DPM compared to OQPSK for the 10 kHz channel is shown along with results for the imaginary case of lowering the bandwidth to 5 kHz and maintaining the transmitted bitrate.

Digital Phase Modulation

Here we present DPM and some of its properties. Reference 1 has further informations on DPM.

DPM Transmitter

The DPM signal is equal to

$$Re(s(t)) = Re(\exp(j2\pi h \sum_{l=1}^{\infty} a_l h_p(t-lT_s)) \qquad (1)$$

where

a_l - transmitted symbols, $a_l \in \{(M-1),\ldots,1,-1,\ldots,-(M-1)\}$

$h_p(t)$ - phase shaping filter

$$h_p(t) \begin{cases} 0, & 0 \le t \le T_L \\ 0, & \text{otherwise} \end{cases}$$

T_s - symbol duration

h - modulation index

$Re(x)$ - real part of x

Observe that (1) has constant envelope and is of the same form as CPM, but with one important difference. The duration of $h_p(t)$, which for CPM is infinite, is for DMP finite of length T_L.

In a practical realization, s(t) is digital, that is s(nD). The oversampling is defined by

$$\eta = \frac{T_s}{D} \qquad (2)$$

where a typical value of η is 4 to 12, depending on the filter $h_p(t)$ and h. D gives the sampling clock relative to the symbol rate.

The value of the modulation index, h, influences the phase excursions. We have

$$(M-1)\pi h = \max\{\phi(t+T_s) - \phi(t)\} \qquad (3)$$

When discussing h, we should emphasize one important practical difference between DPM and CPM. In CPM, h must have a value[2],

$$h_{CPM} = \frac{2k}{q} \qquad (4)$$

due to the receiver implementation, while in DPM

$$h_{DPM} \in R^{+} \qquad (5)$$

where R^{+} denotes the set of non-negative real numbers.

This valuable freedom which DPM have, proves to be very useful in designing systems under various constraints like tuning to spectral requirements and/or BER performance curves.

The spectrum of s(nD) has a continuous part and a discrete part[1]. In Fig. 1 we show an example of the continuous part. The modulation parameters are

$$M = 8, \quad T_L = 3, \quad h = 0.56, \quad \eta = 8$$

and the filter is a raised cosine (RC) with

$$h_p(t) = \frac{1}{4T_L}\left(1-\cos\left(\frac{2\pi t}{T_L}\right)\right), \quad 0 \le t \le T_L \qquad (6)$$

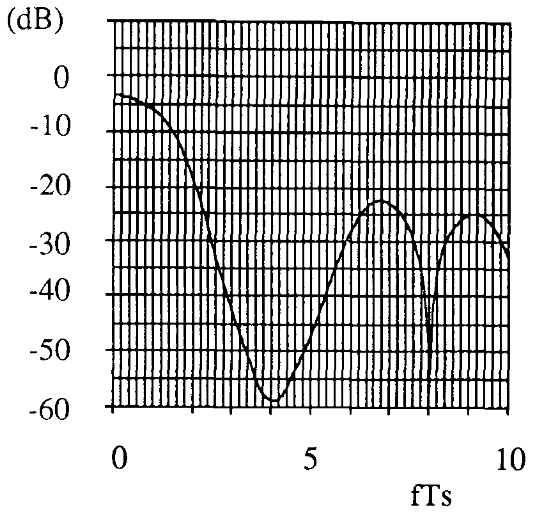

Fig. 1. Spectrum of DPM, continuous part. M=8, $T_L=3 \cdot T_s$, η=8, h=0.56.

In a real system, s(nD) is lowpass filtered so that only the main lobe are transmitted. This results in minor amplitude variations.

General considerations concerning the spectral properties are:

larger h : increased bandwidth
larger T_L: more compact spectrum

DPM Demodulation

Demodulation of DPM is usually based on the application of the Viterbi algorithm (VA). This requires a coherent demodulator for being optimum in the maximum likelihood (ML) sense, since the VA does not give ML performance using a noncoherent metric.

The coherent metrics are in white gaussian noise defined as

$$\lambda_n = \mathrm{Re}(r_n \cdot s^{*}_{n,j}) \qquad (7)$$

where

- r_n - received signal sample (complex) at time n.
- $s_{n,j}$ - one of the possible phases, signals, at time n.

$$j = 1:M^r, \quad r = \left\lfloor \frac{T_L}{T_s} \right\rfloor$$

where we defined $\lfloor \ \rfloor$ as the next closest integer, e.g. $\lfloor 6.7 \rfloor = 7$.

The number of states and transition in the VA for DPM are equal to

$$N_S = M^{r-1} \qquad (8)$$
$$N_{tr} = M^r$$

The states are only correlative due to the finite memory of $h_p(t)$ making N_s and N_{tr} independent of h. CPM has in addition phase states, caused by the infinite memory.

The shape of the BER curve of Fig. 2 is typical for DPM. However, it is possible to shift the crossing point with the QPSK curve and to change the tilt

$$\frac{\partial BER}{\partial (E_b/N_0)} \qquad (9)$$

by varying the parameters involved. These are, as we know from the definition of the DPM signal, h, M and T_L.

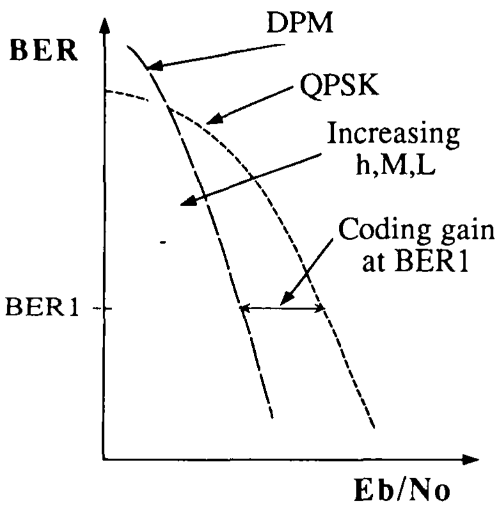

Fig. 2. The general trend of the behaviour of the DPM BER curve for variations of the parameters h, M and T_L.

One important observation is that the parameters are not independent. Therefore, Fig. 2 must be interpreted as an illustration of the fact that the BER curve is controlled by all the three parameters.

Practical systems exhibit a varying phase error between the received carrier phase and the reference in the receiver.

The VA requires a tracking error, $\delta\theta$, below a given value, β, to keep the performance degradation as low as possible. That is

$$|\delta\theta| \le \beta \qquad (10)$$

Examples of simulations with a <u>non-coherent</u> metric applied in the VA, show degradations in E_b/N_0 versus $\delta\theta$ as indicated in Table 1.

Table 1. Degradation in E_b/N_0 at BER = 10^{-2} for two values of $\delta\theta$, using the noncoherent metric in the VA algorithm. M=8, T_L=3, h=0.26.

	[dB]	
$\Delta\ E_b/N_0$	0.1	0.5
$\delta\theta[^\circ]$	5	10

The incoherent metric is defined as

$$\lambda_j = \left| \sum_{i=1}^{N} r_i \cdot s^*_{i,j} \right|, \ \forall_j \qquad (11)$$

where the absolute value is applied instead of the real part from (8).

Using VA and the incoherent metric is an approximate solution since Eq. (11) is not directly summable, but several simulations show that under phase errors it is more tolerable than the coherent metric. Under zero phase error, a minor difference between the performance of the coherent and non-coherent matric is observed.

To compensate for a phase shift, $\delta\theta(t)$, due to doppler, a wide loop bandwidth must be applied. This very often leads to a contradiation with the oscillator phase noise requirements where a narrower bandwidth may be more optimum.

One possible way of overcoming or reducing this problem is to extend the DPM trellis with phase states. The resulting trellis structure then becomes a parallel structure consisting of a number of similar trellis units. Each trellis unit is optimum for a receiver coordinate system offset with θ_i from the transmitter axis where

$i = 1:N_{PT}$

N_{PT} - number of parallel trellis structures

If we now tolerate $\pm 10^0$ for each unit, $N_{PT}=5$ covers

$$\theta_{max} - \theta_{min} = 100^0$$

Consequently, a much wider phase variation range is available as input to the VA, lifting some of the burden off the phase tracking system. At the same time we see that if $\delta\theta$ reaches a value of $20\text{-}30^0$, a PLL solution reaches its linear range limit, making a forward carrier phase estimator solution more attractive[4]. Due to the delay requirements on a satellite link, we can not adapt the same tracking methods as available in a landmobile system.

Performance Comparisons for Inmarsat-M, 8 kb/s in 10 kHz.

The following[1] transmission parameters are valid

- Bitrate 8 kb/s
- Bandwidth 10 kHz
- Adjacent channel interference degradation $\Delta E_b/N_0 \leq 0.1$ dB.
- 3/4 rate convolutional code, constraint length 7.

Fig. 3 gives the result.

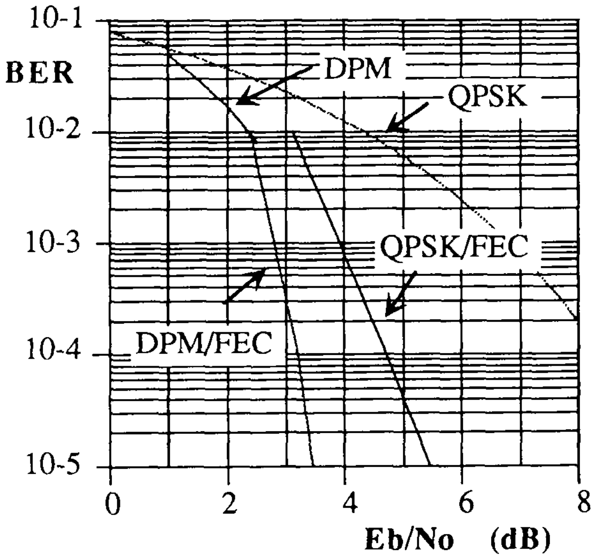

Fig. 3. 8 kbit/sec in 10 kHz user channels. Bit error rate for QPSK, QPSK/FEC and DPM/FEC. FEC: 3/4 rate, K=7, number of states = 64, DPM: M=8, Filter 3RC, h=0.56.

The results exhibit two important properties.

- We can reduce satellite power for a given BER.
- The BER curve for DPM is steeper than for OQPSK.

A consequence of this, is that for a given probability density of the received signal power, p(P), we obtain

$$T_{DPM}(BER \leq BER_1) > T_{OQPSK}(BER \leq BER_1)$$

where $T(BER \leq BER_1)$ is the percentage of time the BER is lower or equal to BER_1.

For the system this implies that the user will have access to a given quality a larger portion of time with DPM than with OQPSK or put another way the average quality obtained by DPM is expected to be higher than in a OQPSK system.

Performance of 8 kb/s in 5 kHz Bandwidth

Result for this imaginary case is shown in Fig. 4. The DPM parameters are M=8, filter 3RC and h=0.26.

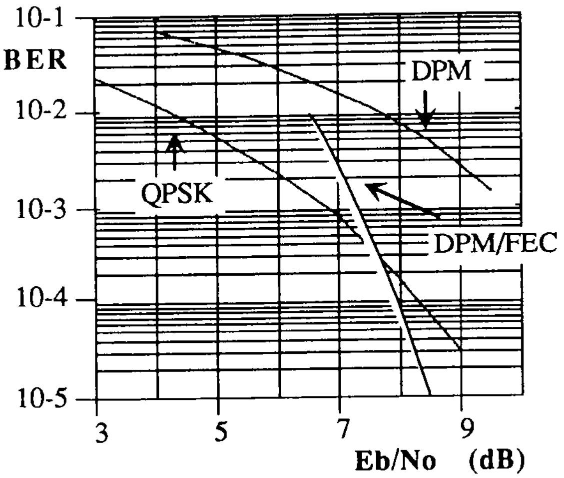

Fig. 4. BER curve for QPSK, DPM and DPM/FEC. DPM: M=8, filter= 3RC, FEC: 3/4 rate. Coherent demodulation.

If we compare Fig. 3 with Fig. 4, we see that the reduction of bandwidth from 10 kHz to 5 kHz and maintaining the 8 kb/s rate transmitted requires an increase in available power per channel of approximately 4 dB and 10^{-2} and 5 dB at 10^{-5}.

With respect to unfiltered 8PSK, DPM with h=0.26 has a negative gain of 0.3 dB at 10^{-2} and a crossing to positive gain around E_b/N_0=8.7 dB.

Conclusion

DPM is an alternative modulation method for narrowband satellite communication. We have shown that for the Inmarsat-M spesifications, it performs better than OQPSK. With respect to carrier phase estimation, the type of estimator has to be considered by taking into account the demodulation metric and the complexity of the VA unit.

Acknowledgement

The project has been financed by the Norwegian Telecommunications Administration (NTA). The author would like to thank S.R. Skaland, NTA, for many valuable discussions. Also B. Risløw at ELAB-RUNIT who has programmed the simulation software.

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