

# Comparison of Two Modulation Choices for Broadband Wireless Communications

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**Abstract:** Both single-carrier (SC) and multicarrier (MC) modulation schemes are currently recommended for broadband wireless communications. MC schemes are known to be very robust against severe time dispersion effects of multipath propagation, without requiring complex receiver implementations, while SC schemes are supposed to suffer from an "equalization disadvantage", when compared with their MC counterparts, under such highly dispersive conditions. This paper provides an overall comparison of these modulation choices, by considering a similar frequency-domain equalization in both cases, and then by adding the impact of channel coding and amplification issues. Our simulation results allow some relevant conclusions: there is no "equalization advantage" for the MC side; under the MC choice we can compensate for the worse uncoded transmission performance, provided that the code rate is low enough; the well-known power amplification problem of MC schemes with large number of subcarriers can be mitigated (e.g. through digital clipping), but without avoiding a significant "amplification advantage" for the SC side. The overall performance advantage of selected SC schemes is clear from our results, which are especially significant since the required implementation complexity and power consumption are not higher than with the MC choice.

## I. Introduction

Nowadays, feasibility studies and standardization activities are carried out in order to design broadband wireless communication systems, in some cases allowing user bit rates over 100 Mbit/s. This is normally regarded as a big challenge, because the required transmission bandwidth becomes much greater than the coherence bandwidth of the radio channel (due to the time dispersion effects of multipath propagation), leading to severe signal distortion levels. Two classes of modulation schemes, each one with pros and cons, are currently being considered for such broadband systems (see [1] and [2], e.g.), and the ultimate modulation choice is often dictated by arguments such as the following: on the one hand, multicarrier (MC) modulation schemes are widely recognized to be very robust against strong time dispersion effects, without the need of complex receiver implementations; on the other hand, it is well-known that, with conventional single-carrier (SC) modulation schemes, a very high receiver

complexity (and the inherent high power consumption) can be expected when trying to solve the hard "ISI problem" and ensure good BER performances through the use of conventional, time-domain equalization.

In recent years (although the concept is much older), low-complexity frequency-domain equalization (FDE) techniques have been proposed for wireless systems employing SC modulation schemes, under a burst design similar to that used for MC modulation schemes, i.e., with a "cyclic extension" in each burst preamble, which is long enough to cope with the maximum relative channel delay [3-6]. Such burst design implies that the linear convolution, inherent to the time-dispersive channel, becomes equivalent to a circular convolution in what concerns the useful part of the bursts; therefore, it can be shown that low-complexity SC/FDE transmission schemes can be derived from MC transmission schemes, by shifting an Inverse DFT from the transmitter to the receiver. Moreover, it was shown in [7] that the special choice of a "DFT matrix" as spreading matrix for a CDM/MC-based system (code division multiplexing) leads to an equivalent SC/FDE system.

In this paper, we consider both an MC option and an SC/FDE option for broadband wireless communications, possibly subjected to heavy time-dispersive conditions, using the same blockwise transmission approach. For the sake of providing a fair comparison of the two options, we evaluate transmission performances for similar bandwidth efficiencies and implementation complexity levels. In both cases, we consider transmission schemes with and without channel coding, with and without linear power amplification and also with and without space diversity at the receiver side. The paper includes several simulation results, using propagation data obtained with the help of a ray-tracing tool, for a carrier frequency close to 40 GHz and a selected indoor scenario.

## II. Frequency-Domain Equalization

In a first step of this comparative evaluation, we assume uncoded transmission under linear amplification conditions, so as to study the "equalization problem" separately.

Let us consider an  $L$ -branch "FDE receiver", as depicted in fig. 1(A), within a wireless system which uses an SC modulation and an appropriate burst design (block of  $N$  "useful" modulation symbols, plus a cyclic prefix derived from this block). It is assumed that, after being down-

converted and filtered, the signals from all diversity branches are sampled and A/D converted. Then, after S/P conversion and guard removal, each signal is separately transformed into the frequency domain by a DFT processor (e.g. using an  $N$ -point FFT algorithm). To minimize the combined effect of ISI and additive Gaussian noise, the set of multiplying coefficients  $\{c_k^{(l)}; l=1,2,\dots,L; k=0,1,\dots,N-1\}$  can be optimized under the minimum mean-squared error (MMSE) criterium. Let  $\sigma_d^2$  designate the variance of the time-domain data symbols and  $(\sigma_n^{(l)})^2$  the variance of the input noise samples at branch  $l$  (the noise at the several branches is assumed to be uncorrelated), and let  $H_k$  and  $\hat{H}_k$ , respectively, denote the real and the estimated channel frequency response (CFR) at the  $k$ th frequency in the RF band (frequency spacing  $1/NT_s$ , where  $T_s$  is the symbol duration). To minimize the MSE after returning to the time domain, we simply need to minimize  $E\{Y_k - D_k\}^2$ , for each  $k \in \{0,1,\dots,N-1\}$  separately, where  $Y_k = \sum_{l=1}^L c_k^{(l)} Y_k^{(l)}$ , the block  $\{Y_k^{(l)}; k=0,1,\dots,N-1\}$  denotes the DFT of the  $l$ th received block of samples  $\{y_n^{(l)}; n=0,1,\dots,N-1\}$  and  $\{D_k; k=0,1,\dots,N-1\}$  denotes the DFT of the transmitted block of time-domain symbol coefficients  $\{d_n; n=0,1,\dots,N-1\}$ .

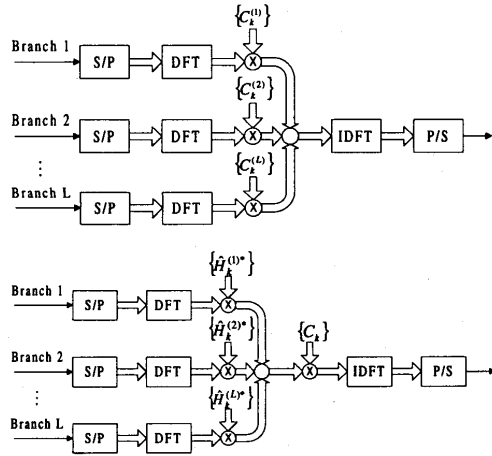


Fig. 1: General SC/FDE receiver structure (A) and equivalent structure when  $\sigma_n^{(1)} = \sigma_n^{(2)} = \dots = \sigma_n^{(L)}$  (B).

For each  $k$ , the MMSE criterium easily leads to the set  $\{c_k^{(i)}; i=1,2,\dots,L\}$ , where

$$c_k^{(i)} = (\sigma_d / \sigma_n^{(i)}) \hat{H}_k^{(i)*} / \left[ 1 + \sum_{l=1}^L (\sigma_d / \sigma_n^{(l)})^2 |\hat{H}_k^{(l)}|^2 \right], \quad (1)$$

$i=1,2,\dots,L$  [8]. Therefore, whenever  $\sigma_n^{(1)} = \dots = \sigma_n^{(L)} \triangleq \sigma_n$ , the FDE parameters are as follows:

$$c_k^{(i)} = \hat{H}_k^{(i)*} / \left[ \alpha + \sum_{l=1}^L |\hat{H}_k^{(l)}|^2 \right], \quad (2)$$

where  $\alpha = \sigma_n^2 / \sigma_d^2$ . Under these “equal noise level” conditions (which we will assume in the following), the receiver structure of Fig. 1(B) can be employed, with

$$C_k = 1 / \left[ \alpha + \sum_{l=1}^L |\hat{H}_k^{(l)}|^2 \right].$$

When adopting  $\alpha=0$  in  $C_k$ , with  $\hat{H}_k^{(l)} = H_k^{(l)}$  (perfect channel estimation) we should get

$$Y_k = \sum_{l=1}^L c_k^{(l)} Y_k^{(l)} = D_k + \sum_{l=1}^L \frac{H_k^{(l)*}}{\sum_{m=1}^L |H_k^{(m)}|^2} \cdot N_k^{(l)}, \quad (3)$$

since  $Y_k^{(l)} = H_k^{(l)} D_k + N_k^{(l)}$ , which corresponds to adopting the zero-forcing (ZF) criterium (cancellation of ISI regardless of the noise level). Clearly, the receiver structure of Fig. 1(B) actually implements a maximal-ratio combining (MRC) diversity scheme, with either  $\alpha = \sigma_n^2 / \sigma_d^2$  (MMSE criterium) or  $\alpha=0$  (ZF criterium) for frequency-domain equalization. When  $L=1$ , these results correspond to those mentioned in [3].

Within a wireless system based on MC modulations, the recommendable FDE receiver structure has much in common with the structure of Fig. 1(B). Obviously, the IDFT block is not employed: such block is used at the transmitter side, since the data contents of a given burst are transmitted in the frequency domain. A decision on the symbol transmitted through “subcarrier  $k$ ” is taken just after the multiplication by  $C_k (k=0,1,\dots,N-1)$ , independently of the decisions concerning the other symbols (we are assuming here an uncoded transmission). In this case, it is appropriate to adopt  $\alpha=0$  in  $C_k$ .

When employing an FDE receiver with an SC modulation, e.g. a QPSK scheme (Quaternary Phase Shift Keying), the achieved performances can be significantly dependent on the criterium for the computation of the FDE parameters. In order to provide an analytical basis for simple comparisons and also to provide some insight into the FDE behaviour, we can resort to a conventional Chernoff bound [8, 9] in both cases, when assuming static (but possibly time-dispersive) channel conditions and a perfect channel estimation. When the MMSE criterium is selected and a perfect estimation of the signal-to-noise ratio is also assumed,

$$BER \leq \exp \left\{ -\sigma_d^2 \left[ 1 - (\epsilon_{MMSE} / \sigma_d^2) \right] / (2\epsilon_{MMSE}) \right\}, \quad (4)$$

where  $\epsilon_{MMSE}$  denotes the resulting MSE (the minimum one), given by

$$\varepsilon_{MMSE} = \frac{\sigma_n^2}{N} \sum_{k=0}^{N-1} 1 / \left[ \frac{\sigma_n^2}{\sigma_d^2} + \sum_{l=1}^L |H_k^{(l)}|^2 \right]. \quad (5)$$

In the ZF case,

$$\begin{aligned} BER &\leq \exp \left\{ -\frac{\sigma_d^2}{2\varepsilon_{ZF}} \right\} = \\ &= \exp \left\{ -\sigma_d^2 / \left[ \frac{2\sigma_n^2}{N} \sum_{l=1}^{N-1} 1 / \sum_{l=1}^L |H_k^{(l)}|^2 \right] \right\}. \end{aligned} \quad (6)$$

It is also important to compare the FDE performances with SC and MC modulations (e.g. with a QPSK constellation on each subcarrier). In the MC case,

$$\begin{aligned} BER &\leq \frac{1}{N} \sum_{k=0}^{N-1} \exp \left\{ -\frac{\sigma_d^2}{2\varepsilon(k)} \right\} = \\ &= \frac{1}{N} \sum_{k=0}^{N-1} \exp \left\{ -\sigma_d^2 / \left[ 2\sigma_n^2 / \sum_{l=1}^L |H_k^{(l)}|^2 \right] \right\}. \end{aligned} \quad (7)$$

Of course, eqns (4), (6) and (7) become identical whenever  $H_k^{(l)}$  does not depend on  $k$ . These eqns also show the following: on the one hand, a similar BER performance can be expected with the SC/FDE (MMSE), the SC/FDE (ZF) and the MC options when the CFR is practically constant across the transmission band; on the other hand, significant performance differences can be expected when the CFR is strongly variable, in spite of being approximately constant in each subchannel. The differences on BER performance, without diversity, become especially clear whenever there are deep notches in the CFR, which is a typical situation with broadband wireless transmission under the impact of multipath propagation. A close observation of the above-mentioned performance bounds shows that MC schemes are strongly affected by such situations, clearly as a consequence of the unreliable receiver decisions which are made on the subchannels that correspond to those notches. With SC schemes, a better BER performance can be expected, since the receiver decisions are based on the signal energy which is transmitted over the entire channel bandwidth and the percentage of "very bad" subchannels typically is quite small (say, less than 10%). However, as eqn (6) suggests, under the ZF criterium even a small percentage of "very bad" channels can lead to a very poor performance, provided they give major contributions to the MSE; this does not happen when employing the MMSE criterium, since such major contributions are avoided by the noise-dependent term that compensates for very low values of the local CFR (see eqn (5)). The potential benefits of space diversity, even with  $L=2$  (the usual choice in practice), are obvious from the observation of eqns (4) to (7), once we can assume that there is a very low probability of having, for any  $k$ , a "very bad" subchannel at all receiver branches simultaneously.

The example given below concerns the uncoded transmission performance within the selected indoor scenario, for a gross bit rate (GBR) of 150 Mbit/s and  $N=512$  as block size. Perfect synchronization and channel estimation have been assumed, under time-

invariant channel conditions during each received burst (In practice, a very accurate channel estimation can be achieved through the use of appropriate reference (training) bursts [4], and a simple LMS algorithm, as proposed in [6], can be employed for tracking the channel variations and adjusting the FDE parameters between adjacent reference bursts.) The simulation results are shown in Fig. 2, both without diversity ( $L=1$ ) and with a double-branch diversity ( $L=2$ ), for the following options: MC, SC/FDE (ZF) and SC/FDE (MMSE) (a QPSK constellation was employed with both modulation choices).

From these results, and also from other results obtained so far under similar, highly time-dispersive conditions, we can conclude that the SC choice gives better performance than the MC choice, with either  $L=1$  or  $L=2$ , provided that the MMSE criterium is adopted. When using diversity, the BER performance improves dramatically in all cases, and, when using  $L=2$  with both modulations, the SC choice typically exhibits a better performance even under the ZF criterium. It should be noted that the BER performances for the MC case, with both  $L=1$  and  $L=2$ , are very close to those corresponding to flat Rayleigh fading conditions: these results can be easily explained, since the Rayleigh fading, in spite of being frequency-selective across the transmission bandwidth, is "approximately flat" at the subchannel level where the decisions are made. The main overall conclusion here is that there is no "equalization advantage" for the MC side.

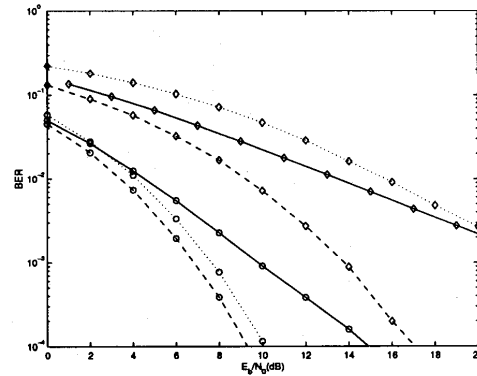


Fig. 2: Raw BER without diversity ( $\diamond$ ) and with a two-branch diversity ( $\circ$ ): solid line for MC, dotted line for the SC/FDE (ZF) and dashed line for the SC/FDE (MMSE).

### III. Impact of Channel Coding

Next, we add the impact of channel coding: a 64-state, rate-1/2 convolutional code is employed, on a burst-by-burst basis, with both modulation choices. A 6-bit tail, for trellis termination, and a 24-bit CRC overhead, for error detection, are appended to each data block. We also consider punctured versions of this code [10], so as to increase the user bit rate while keeping the same gross bit

rate. With the MC choice, an appropriate intraburst interleaving is employed (such interleaving is not required in the SC case). In Figs. 3, 4 and 5 we present the average BER, after the soft-decision Viterbi decoder, for the convolutional codes with rate 1/2, 2/3 and 3/4, respectively ( $E_b$  denotes the 'Average Channel Bit Energy', as in Fig. 2, and a perfect channel estimation for decoding purposes is assumed.).

These results indicate that the benefits of channel coding are much more significant with the MC choice. They also show that SC/FDE (MMSE) and MC performances become very similar at the channel decoder output, with small advantages for the MC choice with the low-rate code and for the SC choice with the high-rate code. This means that the higher coding gain with the MC choice can compensate for the worse uncoded transmission performance, provided that the code rate is low enough.

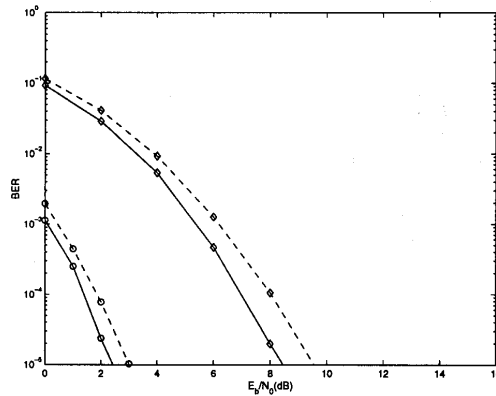


Fig. 3: BER with the rate-1/2 convolutional code, without diversity ( $\diamond$ ) and with diversity ( $\circ$ ): solid line for the MC choice and dashed line for the SC choice.

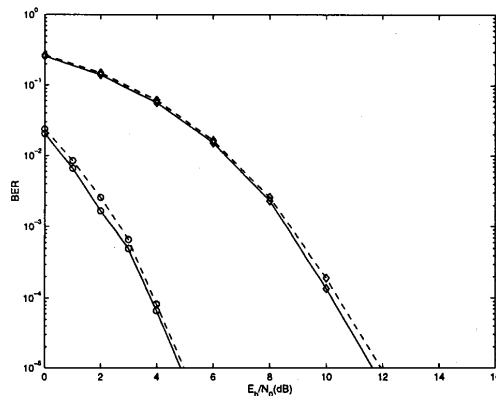


Fig. 4: As with fig. 3 but for the rate-2/3 convolutional code.

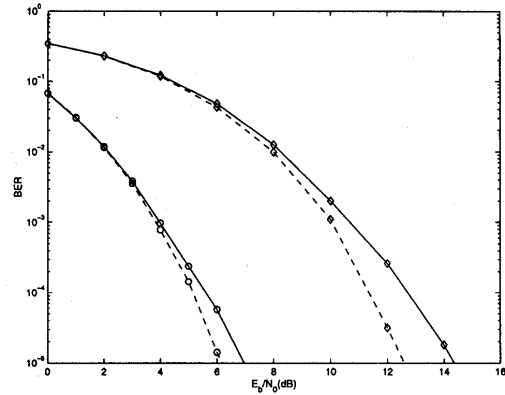


Fig. 5: As with fig. 4 but for the rate-3/4 convolutional code.

#### IV. Power Amplification Issues

From the numerical results of the preceding sections (and other similar results which are not reported for lack of space), it may seem that, when employing a moderate channel coding redundancy, the SC/FDE (MMSE) and the MC modulation choices are nearly equivalent since they provide a given average BER, at a given user bit rate, under similar equalization/decoding complexity levels and average power requirements. However, we still need to consider the power amplification issues with both types of modulations. On the one hand, an SC scheme can be chosen to have a low envelope fluctuation together with a compact spectrum (e.g. a scheme belonging to the PSK class), thus allowing an efficient power amplification. On the other hand, conventional MC schemes are known to have strong envelope fluctuations (an approximately Rayleigh-distributed envelope arises when  $N$  is high), which leads to power amplification difficulties: the I-Q diagram of Fig. 6 can be taken as an example of this undesirable characteristic of MC schemes. A simple and effective way for solving this problem is to adopt an appropriate digital clipping technique, capable of reducing the envelope fluctuations and the peak-to-mean envelope power ratio (PMEPR) of the signals at the amplifier input, while keeping the inherent nonlinear distortion at an acceptable level. When adopting such digital clipping, as proposed in [11] and [12], it is also recommendable to employ a selected oversampling and a complementary filtering, so as to jointly reduce the in-band and out-of-band distortion effects. In the following, we consider an envelope clipping technique for MC transmission which is similar to that described in [12], together with an oversampling factor  $M=2$  and a square-root Nyquist filtering.

In order to give an idea of the envelope fluctuations with both realistic SC and MC schemes, in this context, we show in Fig. 7 the corresponding I-Q diagrams. From the MC side, we consider a QPSK constellation on each

subcarrier, again with  $N=512$  (A); from the SC side, we consider standard QPSK (B) and also Offset QPSK (C). A square-root Nyquist filtering with the same roll-off factor 0.5 is employed in all cases. The “clipping level” for the MC case was selected through a criterium similar to that used in [12]. Clearly, the OQPSK-SC choice has the smallest envelope fluctuations (the envelope never drops down to zero) and the MC choice has the greatest envelope fluctuations, even with the help of the digital clipping effects. The corresponding PMEPR values, when assuming random data, are as follows: 5.7dB (A), 2.8dB (B) and 2.6dB (C).

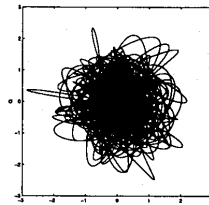


Fig. 6: I-Q diagram for the MC scheme reported in Sec. II ( $N=512$ ).

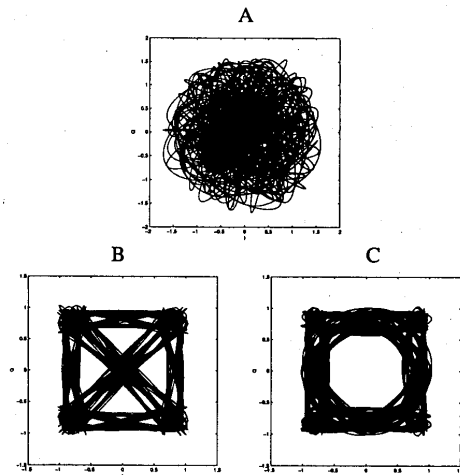


Fig. 7: I-Q diagrams for the clipped MC scheme (A), and also for SC-QPSK (B) and SC-OQPSK (C) schemes.

All signals considered in Fig. 7 have some envelope fluctuation; therefore, a nonlinear distortion effect is unavoidable when submitting each of them to a practical, somewhat nonlinear, power amplification. This distortion implies some degree of spectral spreading, depending on the envelope spread at the input to the amplifier.

Fig. 8 shows some results on fractional out-of-band power (FOBP) for the three schemes of Fig. 7, either under a linear (A) or a grossly nonlinear (B) power amplification, the two extreme cases. It should be noted that, in the MC case, there is some nonlinear distortion even when the amplifier has a linear characteristic: this is a consequence of the digital clipping reported above, appropriately combined with a selected oversampling (factor two).

In Fig. 9 we present several BER performance results, when employing the rate-2/3 convolutional code and two-branch space diversity, for both the MC scheme of Fig. 7(A) and, with FDE (MMSE), the SC scheme of Fig. 7(C). These results have been obtained for the two extreme cases considered above, i.e., linear and (only for the SC scheme) grossly nonlinear power amplification.

It should be noted that the performance results of Figs 8 and 9 concerning SC-OQPSK transmission with grossly nonlinear power amplification are not the “optimum” ones: for example, by replacing the square-root Nyquist pulse by an appropriate “filtered MSK” pulse, it is possible to achieve a FOBP value below  $-50$ dB at  $|f - f_c|T_{bit} = 1$ , together with a slightly improved BER performance. It should also be noted that the results concerning this SC scheme under a linear amplification are practically identical to those obtained with the conventional SC-QPSK scheme (dashed line of Fig. 4, when assuming diversity).

The results reported in this section (PMEPR, FOBP, BER) show that the SC choice has a significant power/bandwidth advantage over the MC choice, when either a linear power amplification (for the range of variation of the input envelope) or a grossly nonlinear power amplification are intended. When the amplifier is grossly nonlinear, the FOBP results of Fig. 8 are enough to conclude that the SC-OQPSK scheme is the only acceptable choice, and the BER results of Fig. 9 show that the small performance degradation, due to the nonlinear distortion effects with this SC scheme, is also acceptable. An appropriate design of the modulation pulse within the SC-OQPSK modulation format, so as to get “optimum” results for specific spectral efficiency requirements and nonlinear amplifying characteristics, can be carried out easily (e.g. using the transmission modelling approach presented in [13]), but is obviously beyond the scope of this paper.

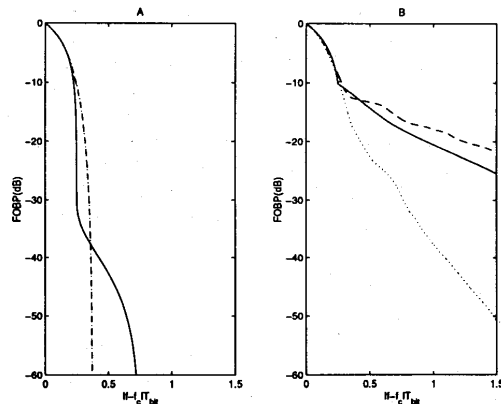


Fig. 8: FOBP with a linear amplifier (solid line, for MC, and dot-dashed line, for SC, in (A)) and with a grossly nonlinear amplifier (solid line, dashed line and dotted line for MC, SC-QPSK and SC-OQPSK, respectively, in (B)).

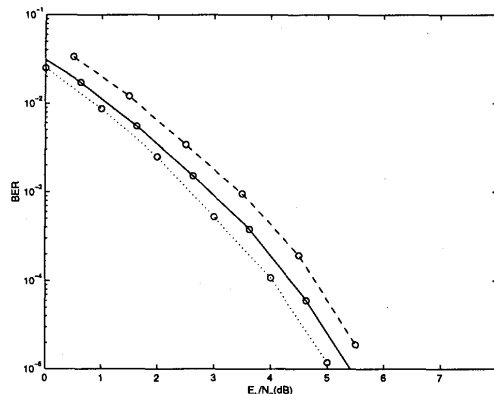


Fig. 9: BER for the following cases: MC scheme with digital clipping and linear amplification (solid line); SC-OQPSK scheme with linear amplification (dotted line); SC-OQPSK scheme with grossly nonlinear amplification (dashed line).

## V. Final Remarks

The comparative performance results of this paper are especially significant, since the overall signal processing complexity (transmitter plus receiver) with the SC choice is not higher than with the MC choice. Actually, the number of required FFT processors is  $L+1$  in both cases (one in the transmitter and  $L$  in the receiver, in the MC case, and  $L+1$  in the SC receiver). However, the MC transmitter requires an additional signal processing effort, due to the digital clipping needs; moreover, the increased amplifier nonlinearity which is allowed with selected SC schemes also means that a more power-efficient amplifier implementation is available through the SC choice.

When the wireless network includes fixed terminals (e.g. within base stations) a realistic solution is to choose an SC/FDE scheme, exhibiting low envelope fluctuations, for the uplink, and an MC scheme for the downlink. This means an implementation advantage for the mobile terminals, where simple SC transmission and MC reception functions are then carried out. The "implementation charge" is concentrated in the fixed terminals (where increased power consumption and cost are not so critical), concerning both the signal processing effort and the power amplification difficulties.

## REFERENCES

- [1] - U.Dettmar et al, "Modulation for HIPERLAN Type 2", VTC'99 (Spring), Houston, USA, May 1999.
- [2] - H.Zou et al, "Equalized GMSK, Equalized QPSK and OFDM, a Comparative Study for High-speed Wireless Indoor Data Communications", VTC'99 (Spring), Houston, USA, May 1999.

- [3] - H.Sari et al, "An Analysis of Orthogonal Frequency-division Multiplexing for Mobile Radio Applications", VTC'94, Stockholm, Sweden, June 1994.
- [4] - G.Kadel, "Diversity and Equalization in Frequency Domain-A Robust and Flexible Receiver Technology for Broadband Mobile Communication Systems", VTC'97, Phoenix, USA, May 1997.
- [5] - A.Czylwik, "Synchronization for Single Carrier Modulation with Frequency Domain Equalization", VTC'98, Ottawa, Canada, May 1998.
- [6] - M.Clark, "Adaptive Frequency-domain Equalization and Diversity Combining for Broadband Wireless Communications", IEEE Journal on Sel. Areas in Comm., Vol. 16, N° 8, October 1998.
- [7] - K.Brüninghaus and H.Rohling, "Multi-carrier Spread Spectrum and its Relationship to Single-carrier Transmission", VTC'98, Ottawa, Canada, May 1998.
- [8] - A.Gusmão and J.Conceição, "On Frequency-domain Equalization and Space Diversity for Broadband Wireless Communications", IEEE/IEE ISCTA'99, Ambleside, UK, July 1999.
- [9] - J.Saltzberg, "Intersymbol Interference Error Bounds with Applications to Ideal Bandlimited Signalling", IEEE Trans., IT-14, July 1968.
- [10] - J.Cain et al, "Punctured Convolutional Codes of Rate  $(n-1)/n$  and Simplified Maximum Likelihood Decoding", IEEE Trans., IT-25 Jan. 1979.
- [11] - X.Li and L.Cimini, Jr., "Effects of Clipping and Filtering on the Performance of OFDM", IEEE Comm. Let., may 1998.
- [12] - R.Dinis and A.Gusmão, "On the Performance Evaluation of OFDM Transmission Using Clipping Techniques", VTC'99 (Fall), Amsterdam, September 1999.
- [13] - A.Gusmão, V.Gonçalves and N.Esteves, "A Novel Approach to Modelling of OQPSK-type Digital Transmission over Nonlinear Radio Channels", IEEE Journal on Sel. Areas in Comm., SAC-15, May 1997.