Transmission Techniques for Digital Terrestrial TV Broadcasting

The authors discuss the potential of OFDM signaling, with its limitations and inherent problems, as well as another potential technique that has so far been overlooked: single-carrier transmission with frequency-domain equalization.

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Digital TV broadcasting by satellite, cable, and terrestrial networks is currently an area of intensive development and standardization activities, particularly in North America and Europe. From these applications, technically the most challenging one is terrestrial broadcasting, due to the presence of strong echoes which characterize the propagation medium. What makes the problem even more difficult is the objective in Europe of deploying single-frequency networks in order to increase the number of TV channels in the allocated frequency bandwidth. In single-frequency networks, all transmitters are synchronized to a common highly stable frequency source and simultaneously broadcast a given TV channel using the same carrier frequency and symbol timing. A TV receiver tuned on a particular channel receives a combination of useful signals incoming from different transmitters with different delays. The overall channel is then modeled as a time-dispersive channel with a long impulse response that may span several hundreds of symbol periods. (The symbol rate is in the range of 5 to 7 Mbaud.)

The common approach for digital terrestrial TV broadcasting in Europe is based on Coded-OFDM (COFDM), which has become extremely popular within the broadcasting community over the past decade. This technique was initially proposed for digital audio applications [1-4], and work on digital terrestrial TV broadcasting followed the same lines a few years later [5-10]. It is often claimed that COFDM is the only technique that makes single-frequency networks feasible. These claims can be attributed to the fact that the usual comparisons between OFDM signaling and single-carrier transmission implicitly assume that the latter technique employs an adaptive time-domain equalizer. Clearly, time-domain equalizers cannot easily handle intersymbol interference (ISI) on channels with very long impulse responses, and may fail on single-frequency networks. It is, therefore, not surprising to see these comparisons lead to the conclusion that single-carrier transmission does not offer the potential of single-frequency networks.

Unlike the situation in Europe, single-frequency networks do not seem to be an issue in the United States, and current standardization activities are focused on the 8-level vestigial sideband (8-VSB) modulation technique. This modulation is essentially equivalent to the single-carrier 64-state quadrature amplitude modulation (64-QAM) which has been standardized in Europe for digital cable television. The VSB vs. QAM issue in the United States is essentially an implementation issue, because the two modulation techniques are mathematically equivalent — provided they use the same type of channel filtering. Note that QAM is the conventional approach to bandwidth-efficient digital communications, whereas VSB is an old technique borrowed from the world of analog communications. The other important point to make on the HDTV activities in the United States is that they are centered on time-domain channel equalization, which is perhaps sufficient for average broadcast channels, but the frequency-domain equalization proposed in this article may open up new perspectives for handling more difficult channels with longer impulse responses.

This article discusses potential transmission techniques for digital terrestrial TV broadcasting. After reviewing OFDM signaling, we point out a strong analogy between this technique and frequency-domain channel equalization in single-carrier systems. It turns out that with a frequency-domain equalizer at the receiver, single-carrier systems can handle the same type of channel impulse responses as OFDM systems. In the absence of channel coding, single-carrier systems with frequency-domain equalization in fact substantially outperform OFDM signaling (which requires powerful channel coding and frequency-domain interleaving to recover its inherent performance loss). The main conclusion of our analysis is that single-carrier transmission with frequency-domain equalization offers the possibility of single-frequency networks while alleviating the nonlinear distortion and carrier synchronization problems inherent to OFDM.

In the next section we briefly recall the principle and the history of OFDM signaling, and in the third section we discuss frequency-domain equalization for OFDM systems, and highlight the fact that in the

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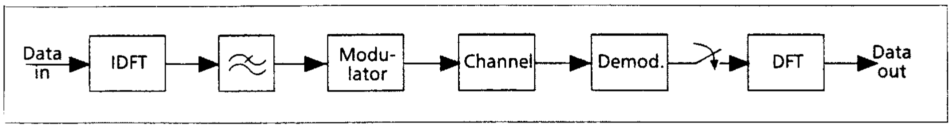


■ **Figure 1.** *Simplified block diagram of an OFDM system.*

uncoded case, OFDM suppresses the need for channel equalization only with PSK signal sets. The fourth section is devoted to frequency-domain equalization for single-carrier systems, pointing out a strong analogy between this technique and equalized OFDM systems. We next discuss the carrier synchronization issue before introducing COFDM, which makes use of channel coding and frequency-domain interleaving, followed by our computer simulation results, with our conclusions presented in the final section.

Principle and History of OFDM

OFDM is a multicarrier transmission technique based on the discrete Fourier transform (DFT). A simplified block diagram of an OFDM system is shown in Fig. 1. Serial-to-parallel and parallel-to-serial conversions inherent in this scheme are dropped for convenience. The transmitted signal is of the form

$$s(t) = Re\left\{ \sum_{n=-\infty}^{+\infty} b_n f(t-nT) e^{j(\omega_0 t+\varphi)} \right\} \quad (1)$$

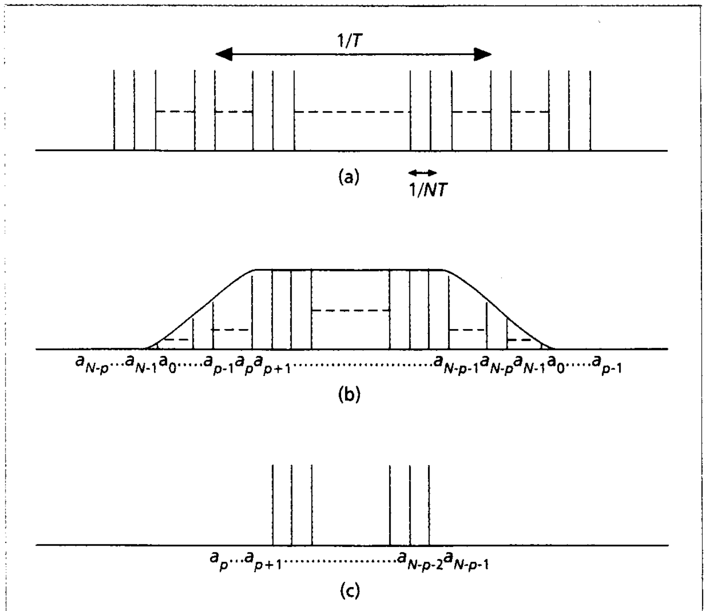
where Re(.) denotes real part, $f(t)$ designates the transmit filter impulse response, T is the symbol period, ω_0 is the carrier radian frequency, φ is the carrier phase, and the transmitted $\{b_n\}$ sequence is obtained from the input information sequence $\{a_n\}$ through an N-point inverse DFT (IDFT). In order to distinguish successive DFT blocks, we write the index n in (1) as $n = m.N + k$ with $k = 0, 1, \ldots, N-1$, and m integer. The $\{b_n\}$ sequence in (1) is then given by

$$b_k(m) = \frac{1}{N} \sum_{l=0}^{N-1} a_l(m).e^{j2\pi kl/N}, \quad k = 0,1,\ldots,N-1. \quad (2)$$

In this 2-index representation, $a_l(m)$ represents the lth input symbol of the mth IDFT block, and $b_k(m)$ is the kth output sample of the same block. After this transformation, the N parallel output samples are converted into a serial form, lowpass filtered, and passed to a quadrature modulator which shifts the signal spectrum to center it on the center frequency $f_0 = \omega_0/2\pi$.

On the receiver side, the received signal is coherently demodulated, sampled at the symbol rate $1/T$, and passed to a DFT operator which converts the signal back to the frequency domain. The demodulator comprises a lowpass filter which limits noise and interference from adjacent channels, without distorting the received signal.

The lowpass filter on the transmitter side needs particular attention. The IDFT output has a periodic spectrum with period $1/T$, each period being composed of N Sinc(x) = Sin(x)/x pulses with a frequency separation of $1/NT$. In order to limit the transmitted signal spectrum and assign one Sinc frequency pulse to each of the N symbols comprised in a given IDFT block, an ideal (rectangular) lowpass filter is needed. Such a filter would limit the signal spectrum to $1/T$ as would a Nyquist filter with a roll-off factor $\alpha = 0$ in single-carrier transmission. Unfortunately, such a filter is not physically realizable, and a non-ideal filter with excess bandwidth must be substituted.



■ **Figure 2.** *The transmit filtering issue in OFDM: (a) output spectrum before filtering; (b) output spectrum after raised-cosine Nyquist filtering; (c) filtered output spectrum with 2p virtual carriers.*

The transmit filtering issue is illustrated in Fig. 2. Each delta pulse in this figure represents a Sinc pulse centered on that frequency and whose zero crossings are $1/NT$ Hz apart. Fig. 2a shows the periodic (infinite-bandwidth) IDFT output spectrum. Next, assuming a raised-cosine spectral shaping, Fig. 2b shows the spectrum of the lowpass filter output. Note that a single frequency pulse is assigned to each symbol whose spectrum falls in the flat region of the filter transfer function. This holds for the $a_p, a_{p+1}, \ldots, a_{N-p-1}$ symbols in the picture at hand. As for the p edge symbols on each side of the N-point DFT block, two frequency pulses are assigned to each of them. Finally, Fig. 2c shows the transmit filter output spectrum when the p edge symbols on each side of the DFT block are set to zero. In this case, no frequency pulses appear in the roll-off region, and the respective center frequencies of two adjacent channels can be set $1/T$ Hz apart. The carriers in the roll-off region which are set to zero are often referred to as "virtual carriers," a terminology that we adopt in the sequel.

At the IDFT output, a "guard interval" is inserted (not shown in Fig. 1) between successive blocks, and the corresponding portion of the received sequence is dropped before the DFT at the receiver.

Under the acronym DMT, OFDM signaling is becoming the basis of a world standard for asymmetric digital subscriber line (ADSL) services.

The guard interval usually consists of a cyclic extension of the DFT output blocks. Provided that its length is larger than the channel impulse response, the cyclic prefix makes the linear convolution of the channel looks like circular convolution inherent to the discrete Fourier domain. On the other hand, the introduction of a guard interval of length p expands the transmitted signal bandwidth by a factor of p/N, where N designates the DFT block length. In order to limit bandwidth expansion, the length of the guard interval must be a small fraction of the DFT block length. In practical applications, the guard interval length is chosen as a function of the expected maximum channel impulse response length, and the DFT block length is then selected so as to keep the bandwidth expansion below 20 percent.

OFDM signaling was developed back in the '60s, e.g., [11-14], and used in some military HF communication systems [15-17]. It was also considered for use in high-speed modems [18, 19], but did not significantly develop in this field, and international CCITT standards for high-speed modems are based on single-carrier transmission. It was later proposed for digital mobile radio systems [20] to alleviate the channel equalization problem, increase robustness against impulse noise, and possibly make a better use of the available channel bandwidth. A summary of the development of OFDM in different application areas can be found in [21].

OFDM became popular in Europe in the mid-'80s with the Eureka 147 project on Digital Audio Broadcasting (DAB) [1]. COFDM was selected for that application, and this technique has now entered the final stages of a long standardization process. With the QPSK modulation employed in DAB applications, OFDM signaling allows to differentially equalize the transmission channel, i.e., it can cope with multipath propagation without channel identification and without involving any adaptive parameters in the receiver. At the beginning of the '90s, the focus was switched to terrestrial broadcasting of digital TV signals, and work in Europe followed very much the same lines as in the DAB project. There is today a general feeling within the broadcasting community that only COFDM can cope with difficult propagation conditions and opens up the perspectives for single-frequency networks.

Under the acronym of DMT (Discrete Multi-Tone), OFDM signaling is also becoming the basis of a world standard for asymmetric digital subscriber line (ADSL) services [22, 23]. It allows to transmit a 6 Mb/s data rate to the subscribers over the existing twisted-pair telephone network. For this application, multicarrier transmission gives the opportunity of adapting the transmitted signal to the channel in a way analogous to the pre-emphasis/de-emphasis technique in analog communication systems. More specifically, the signal constellations used for different carriers can be independently selected in accordance with the channel attenuation and interference at the corresponding frequencies. In addition, it reduces the effect of impulse noise present in that environment.

In the following section, we discuss the channel equalization issue in OFDM systems assuming for the moment that the transmitted signal is uncoded.

Channel Equalization

Let $h(t)$ designate the channel impulse response and $H(\omega)$ its Fourier transform, i.e., the channel transfer function. If the number of carriers is sufficiently large, the channel transfer function becomes virtually nonselective within the bandwidth of individual carriers. (Strictly speaking, the bandwidth of a modulated carrier is infinite, but we are referring here to the carrier spacing $1/NT$ which contains most of the energy.) Focusing on one particular carrier, the influence of multipath fading reduces to an attenuation and a phase rotation. This observation has often led in the past to the erroneous interpretation that OFDM signaling resolves the channel selectivity and the resulting ISI problem. One should not lose sight that the useful signal to be recovered is the entire multiplex, and not only one of its components. Expressed in the frequency domain, the Nyquist criterion (for ISI-free transmission) requires that the channel has flat amplitude and linear phase responses. The fact that each carrier in OFDM has a different attenuation and a different phase rotation implies that the channel is not Nyquist and still needs to be equalized.

Referring back to the channel transfer function $H(\omega)$, we let H_k designate its value within the bandwidth of the kth carrier. Equalization of the channel requires that at the DFT output in the receiver, the kth carrier signal be multiplied by a complex coefficient

$$C_k = 1/H_k. \tag{3}$$

This is the result of an optimization based on the zero-forcing (ZF) criterion [24] which aims at canceling ISI regardless of the noise level. To minimize the combined effect of ISI and additive noise, the equalizer coefficients can be optimized under the minimum mean-square error (MMSE) criterion. This optimization yields

$$C_k = \frac{H_k^*}{|H_k|^2 + \sigma_n^2/\sigma_a^2} \tag{4}$$

where σ_n^2 is the variance of additive noise, and σ_a^2 is the variance of the transmitted data symbols. Note that the MMSE solution reduces to the ZF solution for $\sigma_n^2 = 0$.

The ZF criterion does not have a solution if the channel transfer function has spectral nulls in the signal bandwidth. Inversion of the channel transfer function requires an infinite gain and leads to infinite noise enhancement at those frequencies corresponding to spectral nulls. In general, the MMSE solution is more efficient, as it makes a trade-off between residual ISI (in the form of gain and phase mismatches) and noise enhancement. This is particularly attractive for channels with spectral nulls or deep amplitude depressions.

Channel equalization in OFDM systems thus takes the form of a complex multiplier bank at the DFT output in the receiver. If the modulation used is a phase-shift keying (PSK) signal format, the channel does not need amplitude equalization, because the information is entirely carried by the signal phase. In addition, phase equalization can be made differential, provided that differential encoding is used at the transmitter. These observations, made in one of the early papers on OFDM

[14], are adopted in [1] and subsequent papers which form the basis of the European DAB project.

To visualize how differential channel equalization works, we write the kth symbol at the nth DFT block output as

$$\begin{aligned} y_k(n) &= H_k(n).a_k(n) + w_k(n) \\ &= \rho_k(n).e^{j\theta_k(n)}.a_k(n) + w_k(n) \end{aligned} \tag{5}$$

where $w_k(n)$ is a noise term, and $\rho_k(n)$ and $\theta_k(n)$ respectively designate the channel amplitude and phase responses at the kth carrier frequency during the nth DFT block. This representation implicitly assumes that the channel is stationary over the period of a DFT block.

The first differential equalization technique uses the point with the same index in the previous DFT block as phase reference. The equalized signal corresponding to $y_k(n)$ is given by

$$\begin{aligned} z_k(n) &= y_k(n).y_k^*(n-1) \\ &= \rho_k(n)\rho_k(n-1)e^{j\theta_k(n)}e^{-j\theta_k(n-1)} \\ &\quad a_k(n)a_k^*(n-1) + w'_k(n) \end{aligned} \tag{6}$$

where $w'_k(n)$ is the sum of two signal x noise products and one noise x noise product. Under the assumption that the channel response does not significantly vary over the period of two consecutive DFT blocks, $z_k(n)$ can be approximated as

$$z_k(n) = r_k^2(n)ak(n)a_k^*(n-1) + w'_k(n). \tag{7}$$

Provided that the $\{a_k\}$ sequence is differentially encoded, this detector recovers the original (uncoded) sequence. At high values of the signal-to-noise ratio (SNR), the noise x noise product in $w'_k(n)$ is negligible, and the penalty incurred by this detection technique is limited to 3 dB. For carriers corresponding to deep amplitude depressions (and low SNR values), the penalty will be, however, well in excess of 3 dB.

The second differential equalization technique is based on the following operation

$$z_k(n) = y_k(n).y_{k-1}^*(n). \tag{8}$$

Here, detection of a symbol transmitted on a given carrier is made using an adjacent carrier in the same DFT block as reference. Implementation of this technique requires that the first nonzero carrier carries a predetermined sequence known from the receiver. For large N (number of carriers), there is little difference between the respective attenuations and phase rotations of two adjacent carriers, and the additional SNR degradation incurred by this detection technique remains small. It is not difficult to view that differential detector (6) works better than detector (8) on static channels or on channels with slow time variations, and that the converse is true for rapidly time-varying channels.

Now that we have described how the detection process can be performed in OFDM with PSK signal sets without explicit channel equalization, it is important to point out an ambiguous terminology that may lead to some confusion. Differential detection (or demodulation) in standard single-carrier transmission is used to avoid carrier synchronization, which is a prerequisite for coherent demodulation. Although operation of the detectors described by (6) and (8) resemble that of a differential demodulator in single-carrier systems, they do not suppress the carrier recovery function, and carrier frequency synchronization remains mandatory, as demonstrated in one of the following sections.

Frequency-Domain Equalization

Analyzing the operation principle of OFDM, the authors noticed a striking resemblance to frequency-domain channel equalization for traditional single-carrier systems, a concept proposed more than two decades ago [25]. The motivation for frequency-domain equalization was due to the ability of this technique to accelerate the initial convergence of the equalizer coefficients. It turns out that frequency-domain equalization did not establish itself as a popular technique over the years, and most, if not all, adaptive equalizers used in state-of-the-art communication systems are implemented in the time domain. Although frequency-domain adaptive filtering appears in standard signal processing textbooks (e.g., [26]) and articles, it almost completely vanished from the digital communications literature.

Frequency-domain equalization is illustrated in Fig. 3a, which shows the baseband-equivalent model of a single-carrier system employing this equalization technique. The received signal samples are passed to an N-point DFT, each output sample is multiplied by a complex coefficient C_i, and the output is passed to an IDFT to transform the signal back to the time domain. The coefficient sequence $(C_0, C_1, \ldots, C_{N-1})$ which determines the equalizer frequency response is the DFT of a sequence $(c_0, c_1, \ldots, c_{M-1}, 0, \ldots, 0)$, where $(c_0, c_1, \ldots, c_{M-1})$ represents the tap-gain vector of the equivalent M-tap time-domain equalizer. Now, if we take the system sketched in Fig. 4a and place it between an IDFT operator and a DFT operator, we obtain an OFDM system incorporating a frequency-domain equalizer. Obviously, the DFT and IDFT operators at the output end cancel each other, and the system simplifies to what we see in Fig. 3b. This is precisely the schematic diagram of the equalized OFDM system discussed in the previous section. Figs. 4a and 4b thus give evidence of the strong similarities of OFDM signaling and frequency-domain equalization in single-carrier systems. In both cases, time/frequency and frequency/time transformations are made. The difference is that in OFDM systems, both channel equalization and receiver decisions are performed in the frequency domain, whereas in single-carrier systems the receiver decisions are made in the time domain, although channel equalization is performed in the frequency domain.

From a purely channel equalization capability standpoint, both systems are equivalent, assuming they use the same DFT block length. They have, however, an essential difference that should not be underestimated. Since the receiver decisions in uncoded OFDM are independently made on different carriers, those corresponding to carriers located in a region with a deep amplitude depression will be unreliable. If we assume that one hundredth of the N modulated carriers are affected by a spectral null or a deep spectral notch, the system will have a residual error rate on the order of 10^{-2}. (This is due to the fact that the error probability of decisions corresponding to carriers located in a deep

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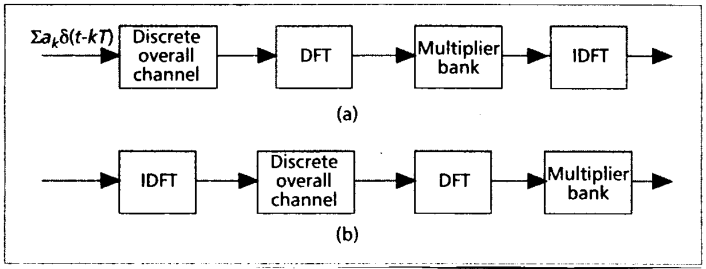


■ **Figure 3.** *Frequency-domain channel equalization; (a) baseband-equivalent model of a single-carrier system with a frequency-domain equalizer; (b) baseband-equivalent model of an equalized OFDM system.*

spectral notch will be close to 0.5.)

On a static channel, one can get out of this problem by assigning "virtual carriers" to the frequency bandwidth a spectral notch, but this technique (which is exploited in ADSL systems) does not lend itself to radio channels affected by rapidly time-varying multipath fading events. In that situation, the frequent errors randomly occurring in different parts of the signal spectrum need to be corrected using error correction coding and frequency-domain interleaving.

Although the operation of a single-carrier system with a frequency-domain equalizer looks similar to that of OFDM, closer inspection reveals that this technique substantially alleviates the detection problem. Once the channel is equalized in the frequency domain, the signal is transformed back to the time domain, and the receiver decisions are based on the signal energy transmitted over the entire channel bandwidth. In other words, the SNR value that dictates performance (assuming that residual ISI is negligible) corresponds to the average SNR of the channel. Consequently, the performance degradation due to a deep notch in the signal spectrum remains small with respect to that suffered by OFDM.

The foregoing analysis indicates that with frequency-domain equalization, single-carrier transmission is substantially superior to OFDM signaling. Without channel coding, OFDM is in fact not usable on fading channels, as deep notches in the transmitted signal spectrum lead to an irreducible BER. In order to work satisfactorily, OFDM requires error correction coding with frequency-domain interleaving so as to scatter the signal samples falling in a spectral notch. With OFDM signaling, multipath fading is analogous to Rayleigh fading in single-carrier transmission. Time-domain interleaving, commonly used on Rayleigh fading channels, must then be naturally replaced by frequency-domain interleaving.

In summary, error-correction coding and frequency-domain interleaving play a key role in OFDM systems on multipath fading channels, and this forms the essence of the COFDM system employed in the European digital audio and video broadcasting projects. In contrast, single-carrier transmission with frequency-domain equalization can work without error-correction coding.

It should be noted that the circular convolution problem is encountered in frequency-domain equalization whether the transmission system is single-carrier or OFDM. This is easily understood by referring to the respective operations of the equalizer and the channel. Clearly, the channel performs a linear convolution between the incoming data stream and its impulse response, and the equalizer which attempts to invert the channel transfer function must also perform a linear convolution. If no caution is exercised, a frequency-domain equalizer performs a circular convolution instead of the desired linear convolution.

As mentioned previously, the conventional solution to this problem in OFDM systems is to use a guard interval consisting of a circular prefix. This technique is also applicable to frequency-domain equalization in single-carrier systems, but there are two other techniques that can be used in the latter case. The first is the overlap & save technique and the second is the overlap & add technique. We will not describe here these techniques which can be found in standard textbooks, e.g,. [27], but we will only point out that they can be viewed as the dual of the circular prefix technique in the sense that they make the circular convolution of the equalizer looks like linear convolution, whereas the circular prefix technique makes the linear convolution of the channel looks like circular convolution. The overlap & add and overlap & save techniques do not expand the transmitted signal bandwidth, but increase the receiver complexity.

Carrier Synchronization

When the receiver is perfectly synchronized with the transmitter, and the discrete channel is ISI-free, the IDFT and DFT operators in OFDM systems (Fig. 1) appear cascaded and give the identity operator. In the presence of carrier asynchronism, orthogonality of the multiplexed signals is destroyed and interference is created between the data symbols in a DFT block. This phenomenon is easily visualized as follows: let us denote the DFT operator by an NxN dimensional matrix D. The effect of a radian frequency error Ω between transmitter and receiver is a block transformation that can be represented using the diagonal matrix

$$E=\begin{vmatrix} 1 & 0 & 0 \cdots\cdots 0 \\ 0 & e^{j\Omega T} & 0 \cdots\cdots 0 \\ 0 & 0 & e^{j\Omega T} \ddots \vdots \\ \vdots & \vdots \ddots & \ddots \ 0 \\ 0 & 0 \cdots\cdots 0 & e^{j(N-1)\Omega T} \end{vmatrix} \quad (9)$$

For the kth block, the receiver output is related to the transmitter input through the matrix transformation

$$D' = e^{jkN\Omega T} D^{-1} E D \quad (10)$$

which is obviously not proportional to the identity matrix as long as $\Omega \neq 0$. The implication of this is that carrier frequency offsets in OFDM systems lead to ISI between the symbols of a DFT block. (Many authors refer to the DFT block period as symbol period, and to interference between the symbols of a block as intercarrier interference. We do not use this terminology borrowed from traditional FDM, and do not make any basic distinction between ISI within a DFT block and interference between consecutive blocks.) The nth output sample of the kth DFT block is given by

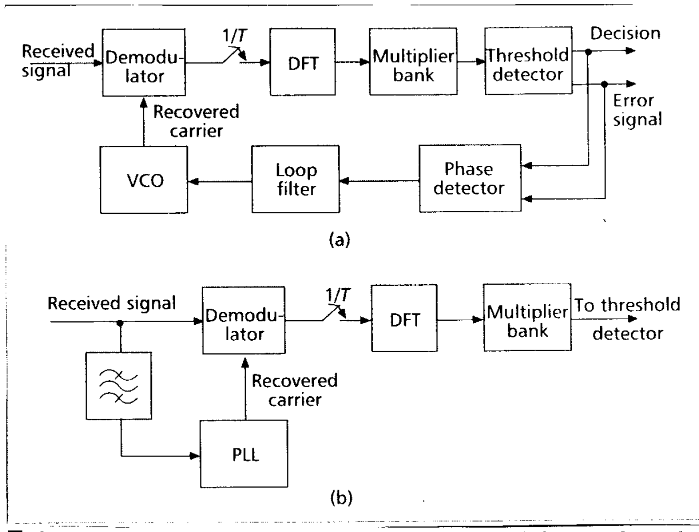
$$y_n(k) = \frac{1}{T} e^{jkN\Omega T} \sum_{m=0}^{N-1} a_m(k) \sum_{l=0}^{N-1} e^{jl\Omega T} e^{j2\pi l(m-n)/N} \qquad (11)$$

and this expression clearly shows that not only the data symbols are rotated, but they also interfere with each other. All terms corresponding to an index $m \neq n$ in the first sum on the right-hand side of (11) are indeed ISI terms. Note that this phenomenon is non-existent in single-carrier systems. Furthermore, substituting (11) into (6) and (8), it is easily verified that neither of these differential detectors suppresses the ISI created by the frequency offset.

To cancel this source of interference, carrier frequency offsets must be compensated before the DFT at the receiver. A first possibility is to use a decision-directed carrier recovery technique as shown in Fig. 4a. The phase detector employs the instantaneous receiver decision and the corresponding error signal. After passing through the loop filter, its output drives a voltage-controlled oscillator (VCO) which delivers the recovered carrier to the demodulator. The problem associated to this technique is that the excessive delay due to the presence of an N-point DFT in the loop will seriously affect its stability, increase its steady-state phase jitter, and reduce its acquisition range [28, 29].

In practical applications, this difficulty is circumvented by resorting to a "pilot tone" carrier synchronization technique, such as the one sketched in Fig. 4b. This technique is easily implemented in OFDM systems, by transmitting an unmodulated carrier at the channel center, with a number of virtual (zero-valued) carriers on both sides. On the receiver side, the received signal is passed to a bandpass filter whose nominal center frequency is equal to that of the pilot carrier, and the filter output drives a phase-lock loop (PLL) that supplies the recovered carrier to the demodulator. This solution alleviates carrier synchronization, but reduces the spectral efficiency and the power efficiency of the system. The loss in spectral efficiency is proportional to the number of virtual carriers, a parameter directly related to the frequency uncertainty of the oscillators used.

The carrier synchronization problem can be solved in a more elegant and efficient manner in single-carrier systems, no matter whether the equalizer is implemented in the time domain or in the frequency domain [30]. A receiver structure comprising a frequency-domain adaptive equalizer and a decision-directed carrier recovery loop is depicted in Fig. 5. The received signal is first downconverted to baseband using a free-running local oscillator, and after lowpass filtering and symbol-rate sampling, the resulting signal is entered to a frequency-domain equalizer. The equalized signal is then passed to a digital demodulator [31], which compensates for frequency offsets between transmitter and receiver. In this structure, the DFT and IDFT (which are part of the equalizer) are outside of the carrier recovery loop and do not affect its stability, acquisition range, and steady-state jitter performance.



■ **Figure 4.** *Carrier synchronization in OFDM systems; (a) decision-directed carrier recovery; (b) pilot-aided carrier recovery.*

Coded OFDM

In the previous sections, the focus was on uncoded-OFDM in which the receiver decisions are made independently on each carrier. Our analysis suggested that without channel coding, OFDM is virtually unusable on multipath fading channels with deep notches occurring in the signal spectrum. It was indicated that performance of uncoded-OFDM is essentially dictated by the lowest SNR value in the signal bandwidth, whereas the IDFT operator preceeding the decision device in single-carrier systems performs perfect SNR averaging over the entire channel bandwidth.

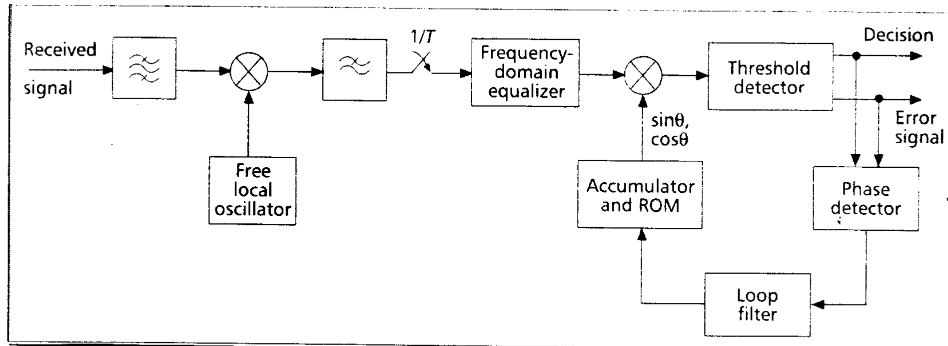
From these considerations, it is clear that the OFDM technique requires channel coding to protect the system from transmission errors which would be frequently occurring in the absence of coding. In a block coding and "hard-decision" decoding approach, a natural choice is to select the code block length equal to the DFT block length of the OFDM system. An OFDM system with such a coding and decoding scheme will be efficient if the code used can correct the errors per block with a high probability.

A better coding approach consists of convolutional coding with frequency-domain interleaving and "soft-decision" decoding. This allows SNR averaging, and the resulting system approaches the performance of single-carrier transmission with frequency-domain equalization. What we need here is an interleaver that uniformly distributes the low-SNR samples over the channel bandwidth and a convolutional code with a large hamming distance.

The receiver performs maximum-likelihood sequence decoding using the well-known Viterbi algorithm, which searches for the most likely path (the path with the smallest metric, or Euclidian distance, from the received noisy and distorted signal) in the code trellis. In an equalized OFDM system, the branch metrics over the nth DFT block period are of the form

$$D(n) = \sum_k \left| y_k(n) - a_k(n) \right|^2 \qquad (12)$$

where $\{y_k(n)\}$ represents the equalized signal



■ **Figure 5.** *Block diagram of a receiver comprising a frequency-domain equalizer and a fully digital demodulator.*

sequence, and $\{a_k(n)\}$ is the data sequence associated to a particular path in the trellis.

A still better approach in COFDM systems is to use channel state information (CSI) in computing the branch metrics:

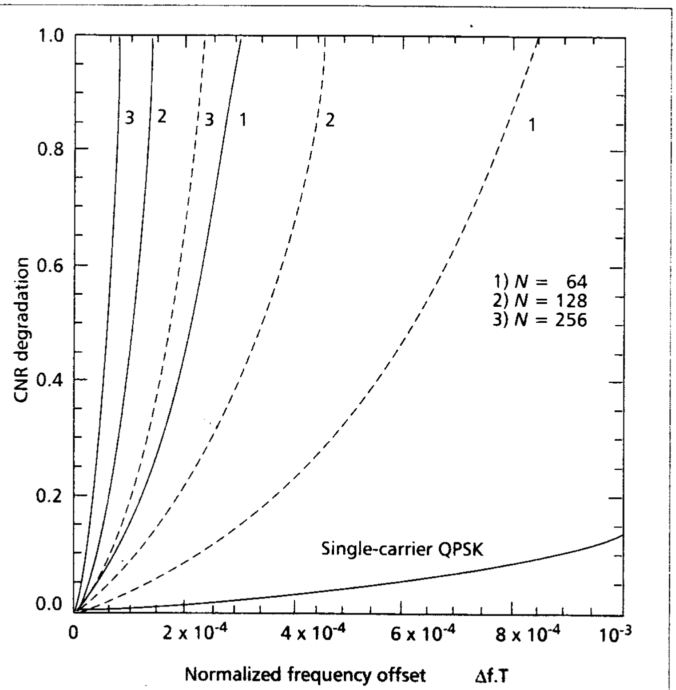
$$D'(n) = \sum_k \left| x_k(n) - \rho_k(n) a_k(n) \right|^2 \qquad (13)$$

where $\{x_k(n)\}$ represents the unequalized signal sequence, and $\{\rho_k(n)\}$ is the sequence of channel attenuation parameters during the nth DFT block. The metric $D'(n)$ can also be expressed as

$$D'(n) = \sum_k \rho_k^2(n) \left| y_k(n) - a_k(n) \right|^2, \qquad (14)$$

a form closely related to the branch metrics of (12). This expression shows that the branch metrics computed using CSI can also be interpreted as the metrics computed using the equalized signal samples and weighting each local metric by the corresponding squared channel attenuation factor. In other words, a small weighting factor is associated to local metrics with low reliability, and a large weighting factor is associated to local metrics with high reliability.

Weighting can be interpreted as the dual of equalizing the channel in the sense that equalization consists of amplifying an attenuated received signal to match it to the nominal decision levels, whereas weighting consists of matching the decision levels to the received signal attenuation. Weighting clearly avoids the noise enhancement inherent to equalized OFDM systems and appears as the best strategy in branch metric computations.



■ **Figure 6.** *CNR degradation vs. normalized frequency offset in OFDM systems (the* N *parameter is the number of carriers). The solid-line curves in OFDM correspond to differential detector (6), and the dashed curves correspond to detector (8).*

Computer Simulation Results

A large number of computer simulations were carried out to compare OFDM signaling and single-carrier transmission in different conditions. For the sake of simplicity, the QPSK signal format was employed in all simulations. Neither this modulation scheme nor the channel models used are representative of terrestrial TV broadcasting systems, but they are still sufficient to make a general comparison and indicate the relative performances of the two transmission techniques. Some of the results were previously reported in [32-34].

Carrier Frequency Offsets

It was pointed out earlier that carrier frequency offsets between transmitter and receiver in OFDM systems lead to ISI and, consequently, to a significant degradation of the carrier-to-noise ratio (CNR). To demonstrate the fact that differential channel equalization does not solve the carrier synchronization problems, we have simulated an OFDM/QPSK system in which demodulation is performed using a free-running local oscillator. After the DFT operator which converts the signal back to the frequency domain, differential channel equalization was performed using either detector (6) or detector (8). The number of virtual carriers was $N/8$, where N denotes the number of points in the DFT. With detector (8), the first nonzero element of an input block was differentially encoded and detected with respect to the last nonzero element of the previous block. The simulated channel was a distortion-free additive white Gaussian noise (AWGN) channel.

The influence of carrier asynchronism on system performance was evaluated by computing the equivalent CNR degradation at the bit error rate (BER) of 10^{-3} and plotting it as a function of the frequency offset, Δf. The results are given in Fig. 6. The CNR degradations displayed in this figure do not take into account the inherent degradation of differential detection that is common to OFDM and single-carrier QPSK systems. The results clearly show that an OFDM/QPSK system does not tolerate a frequency offset $\Delta f = 10^{-3}/T$, whereas the CNR degradation of a single-carrier QPSK system caused by such an offset is only 0.1 dB. A second observation is that in OFDM systems, the CNR degradation is higher with detector (6) which detects the phase of each point in the current DFT block using the same point in the previous block as reference. Note that the phase rotation between these two points due to a frequency offset Δf is $2\pi.N.\Delta f.T$, and the resulting CNR degradation rapidly increases with N. An OFDM/QPSK system with $N = 512$ (not shown in Fig. 6) does not even tolerate a frequency offset $f = 10^{-4}/T$, whereas the degradation caused by such an offset in a single-carrier system is virtually zero.

To give an idea of carrier synchronization problems of OFDM in terrestrial broadcast applications, a frequency offset $\Delta f = 10^{-4}/T$ corresponds to 600 Hz if we assume a transmission speed of 6 Mbaud. Now, even if we assume crystal oscillators with a precision of 10^{-5}, typical frequency offsets in the UHF range amount to several kHz, and these figures (together with the results of Fig. 6) indicate that perfect carrier synchronization is compulsory in OFDM systems.

Nonlinear Distortion

Nonlinear distortion in digital communication systems is primarily due to the transmit power amplifier which must be driven as close to its saturation point as possible, in order to make the best possible use of its output power. Low sensitivity to nonlinear distortion is a particularly stringent requirement in satellite-based systems due to the limited power at the output of the satellite transponder, which comprises a traveling-wave tube (TWT) power amplifier. In the signal bandwidth, the TWT amplifier has a frequency-independent response which can be characterized using its AM/AM and AM/PM curves [35], respectively, given by

$$A(r) = \frac{2r}{1+r^2} \tag{15}$$

and

$$\phi(r) = \frac{\pi}{3} \cdot \frac{r^2}{1+r^2} \tag{16}$$

where r designates the signal amplitude at the amplifier input.

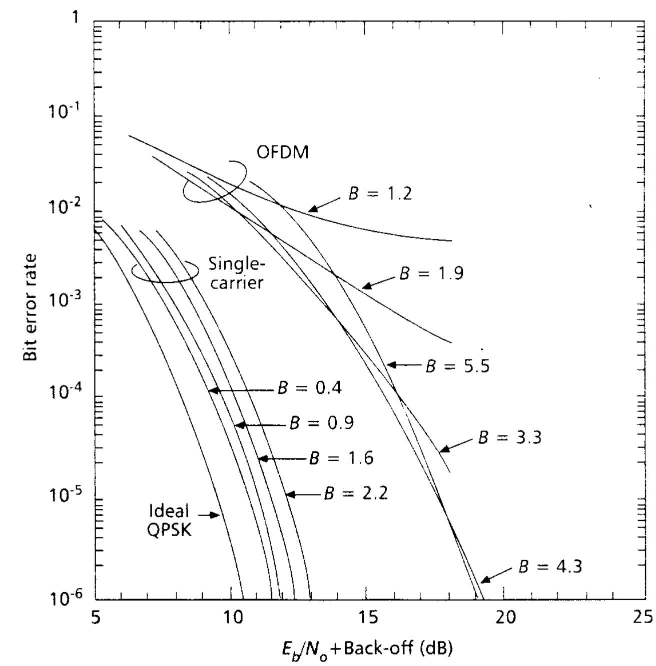
With a nonlinear transmit amplifier, the output power must be backed off from its saturation value in order to reduce the equivalent CNR degradation of the nonlinearity. On the other hand, backing off the amplifier reduces the output power and degrades the link budget. The operating point of the amplifier is, therefore, optimized in practice so as to make the best trade-off between these conflicting two requirements. Our definition of the back-off in this article is the difference between the amplifier's output saturation power and the actually transmitted average signal power.

Using the TWT characteristics given by (15) and (16), the influence of nonlinearity was evaluated by computing the BER on an AWGN channel as a function of E_b/N_0 (energy per bit to noise spectral density ratio) for various values of the back-off (denoted B in the sequel) and plotting it vs. $E_b/N_0 + B$. In this way, we obtain a set of curves that reflect the combined effect of the amplifier back-off B and of the equivalent CNR degradation due to the nonlinearity. The simulated OFDM system has $N = 512$ carriers 64 of which are virtual and a guard interval of 128 symbols. The single-carrier system employed a Nyquist roll-off factor $\alpha = 0.4$, which is consistent with the OFDM system parameters given above.

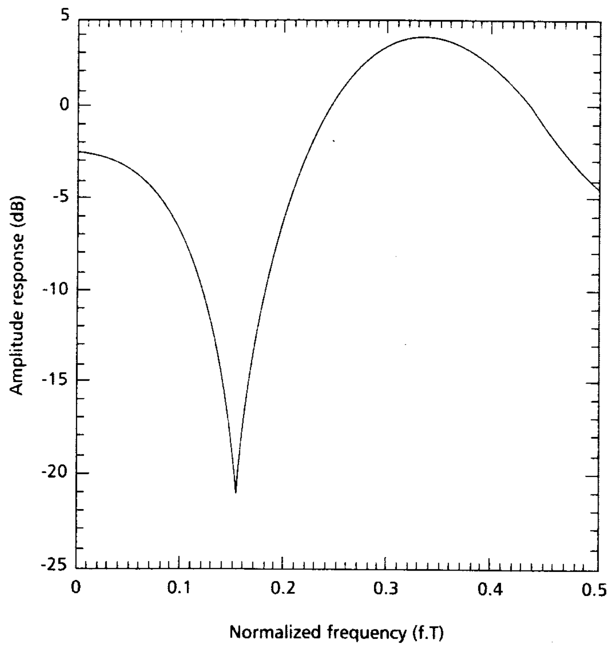
The simulation results are given in Fig. 7. For the single-carrier QPSK system, the best results are obtained with the back-off value $B = 0.4$ dB, which gives a total degradation of 0.8 dB at the BER of 10^{-3}. Next, examining the results corresponding to OFDM, we observe that the best results at the BER of 10^{-3} are obtained with an amplifier back-off of 4.3 dB, which gives a total degradation of 6.8 dB. These results demonstrate that OFDM/QPSK loses some 6 dB with respect to equivalent single-carrier QPSK. Similar results are also obtained with other signal sets, and this implies that an OFDM-based terrestrial broadcast system will be significantly more sensitive to nonlinear distortion than the corresponding single-carrier system.

Channel Equalization

Performance comparison of OFDM and single-carrier transmission was carried out using a number of channels. In this section, we report the results corresponding to two channels, which we refer to as channel A and channel B, respectively. Channel A corresponds to a mild amplitude distortion. Its discrete impulse response, shown in Fig. 6.4.7a of



■ **Figure 7.** *Influence of a TWT nonlinearity on the performance of single-carrier and OFDM QPSK (the B parameter is the amplifier back-off expressed in dB).*



■ **Figure 8.** *Amplitude response of channel B.*

[24], is represented by (0.04, –0.05, 0.07, –0.21, –0.5, 0.72, 0.36, 0.0, 0.21, 0.03, 0.07). Channel B represents a stronger amplitude distortion, and its transfer function has a 25-dB notch in the signal spectrum. Its discrete impulse response is (0.74, –0.42, 0.083, 0.049, –0.12, 0.01). The amplitude response of channel B is shown in Fig. 8, and that of channel A can be found in Fig. 6.4.8a on p. 573 of [24].

In these simulations, we used an OFDM system with $N = 1024$ carriers and a single-carrier system with a 1024-tap frequency-domain equalizer. In both cases, we used a circular prefix of minimum length, i.e., the prefix length was 10 for channel A and 5 for channel B. The equalizer optimization was performed under the MMSE criterion. (It was verified by means of computer simulations that this optimization leads to significantly better performance than the zero-forcing criterion.)

The first set of simulations were performed using uncoded systems. Next, we included the $K = 7$, rate-1/2 convolutional code which has become a "de facto" industry standard for channel coding. In the coded case, the receiver comprised a "soft-decision" Viterbi algorithm for maximum-likelihood decoding. The interleaver used in coded-OFDM was a block interleaver represented by a matrix of 16 columns and eight rows, where the input symbols are written by rows and read by columns. The deinterleaver simply performs the inverse operation. With this interleaver/deinterleaver pair, two symbols transmitted at two adjacent frequencies are separated by 16 symbols at the Viterbi decoder input.

The simulation results are given in Fig. 9 for channel A and in Fig. 10 for channel B. The dashed curves correspond to OFDM and the solid-line curves correspond to single-carrier transmission with frequency-domain equalization. Each figure also includes a dotted curve which corresponds to coded-OFDM with weighted decoding. On both channels, we observe that in the absence of channel coding, single-carrier transmission with frequency-domain equalization substantially outperforms OFDM signaling. The second basic observation is that the convolutional code, used in the second simulation runs, leads to a tremendous improvement, particularly with OFDM signaling. With convolutional coding, frequency-domain equalization, and maximum-likelihood decoding, performance of OFDM becomes very close to that of single-carrier transmission. Finally, with weighted maximum-likelihood decoding, COFDM leads to a slightly improved performance than single-carrier transmission.

Conclusions

The terrestrial broadcast channel is characterized by strong echoes with large delays, which are difficult to handle using a time-domain equalizer. Further, the objective of deploying single-frequency networks in order to make a better use of the available radio spectrum makes the problem even more difficult and precludes single-carrier transmission with time-domain equalization. The focus in Europe for this application has been on COFDM, which can efficiently handle multipath channels with large echo delays.

We have shown that, provided it employs a frequency-domain equalizer, single-carrier transmission can handle the same type of channels as does OFDM signaling. In the absence of channel coding, it even substantially outperforms OFDM signaling; the latter technique requires powerful channel coding and frequency-domain interleaving to recover its performance loss with respect to single-carrier transmission. With coding, interleaving, and weighted decoding, OFDM signaling eventually surpasses the performance of single-carrier transmission with frequency-domain equalization, but it suffers from strong sensitivity to nonlinear distortion and carrier synchronization difficulties. Both techniques may be regarded as strong potential techniques for digital terrestrial TV broadcasting, and they can efficiently compensate for multipath channels with long echo delays. In conclusion, single-carrier transmission with frequency-domain equalization opens up new perspectives for digital terrestrial TV broadcasting.

References

[1] M. Alard and R. Lassalle, "Principles of modulation and channel coding for digital broadcasting for mobile receivers," *EBU Review*, no. 224, pp. 3-25, Aug. 1987.

[2] B. Le Floch, R. Halbert-Lassalle, and D. Castelain, "Digital sound broadcasting to mobile receivers," *IEEE Trans. Consumer Electronics*, vol. 35, no. 3, Aug. 1989.

[3] P. Höher, "TCM on frequency-selective land-mobile fading channels," in Coded Modulation and Bandwidth-Efficient Transmission, E. Biglieri and M. Luise, eds., pp. 317-328, (Amsterdam: Elsevier Science Publishers,1992).

[4] P. Höher *et al.*, "Performance of an RCPC-coded OFDM-based digital audio broadcasting (DAB) system," GLOBECOM '91 Conf. Rec., vol. 1, Phoenix, AZ., Dec. 1991, pp. 40-46.

[5] J. F. Hélard and B. Le Floch, "Trellis coded orthogonal frequency division multiplexing for digital video transmission," GLOBECOM '91 Conf. Rec., vol. 2, Phoenix, AZ., Dec. 1991, pp. 785-791.

[6] B. Sueur *et al.*, "The DIGICAST system intended for digital terrestrial television broadcasting," (in French), *L'Onde Electrique*, vol. 73, no. 4, July/Aug. 1993, pp. 30-34.

[7] U. Reimers, "European perspectives in digital television broadcasting Conclusions of the Working Group in digital television," *EBU Technical Rev.*, no. 256, 1993, pp. 3-8.

[8] D. Wood, "European perspectives in digital television Quality objectives and prospects for commonality," *EBU Techn'l Rev.*, no. 256, 1993, pp. 9-15.

[9] P. Appelquist, "HD-DIVINE, a Scandinavian terrestrial HDTV system," *EBU Technical Rev.*, no. 256, 1993, pp. 16-19.

[10] B. Marti *et al.*, "European activities on digital television broadcasting From company to cooperative projects," *EBU Technical Rev.*, no. 256, 1993, pp. 20-29.

[11] B. R. Saltzberg, "Performance of an efficient parallel data transmission

system," *IEEE Trans. Commun. Technol.*, vol. COM-15, no. 6, Dec. 1967, pp. 805-811.
[12] R. W. Chang, "Orthogonal frequency division multiplexing," U.S. Patent 3 488 445, filed Nov. 14, 1966, issued Jan. 6, 1970.
[13] R. W. Chang and R. A. Gibbey, "A theoretical study of performance of an orthogonal multiplexing data transmission scheme," *IEEE Trans. Commun. Technol.*, vol. COM-16, Aug. 1968, pp. 529-540.
[14] S. B. Weinstein and P. M. Ebert, "Data transmission by frequency-division multiplexing using the discrete Fourier transform," *IEEE Trans. Commun. Technol.*, vol. COM-19, no. 5, Oct. 1971, pp. 628-634.
[15] M. S. Zimmerman and A. L. Kirsh, "The AN/GSC-10 (KATHRYN) variable rate data modem for HF radio," *IEEE Trans. Commun. Technol.*, vol. COM-15, April 1967, pp. 197-205.
[16] P. A. Bello, "Selective fading limitations of the Kathryn modem and some system design considerations," *IEEE Trans. Commun. Technol.*, vol. COM-13, Sep. 1965, pp. 320-333.
[17] G. C. Porter, "Error distribution and diversity performance of a frequency-differential PSK HF modem," *IEEE Trans. Commun. Technol.*, vol. COM-16, Aug. 1968, pp. 567-575.
[18] B. Hirosaki, "An orthogonally-multiplexed QAM system using the discrete Fourier transform," *IEEE Trans. Commun. Techn.*, vol. COM-29, no. 7, July 1981, pp. 982-989.
[19] B. Hirosaki, "An analysis of automatic equalizers for orthogonally multiplexed QAM systems," *IEEE Trans. Commun.*, vol. COM-28, no. 1, Jan. 1980, pp. 73-83.
[20] L. J. Cimini, Jr., "Analysis and simulation of a digital mobile channel using orthogonal frequency division multiplexing," *IEEE Trans. Commun.*, vol. COM-33, no. 7, July 1985, pp. 665-675.
[21] J. A. C. Bingham, "Multicarrier modulation for data transmission: An idea whose time has come," *IEEE Commun. Mag.*, vol. 28, no. 5, May 1990, pp. 5-14.
[22] J. S. Chow, J.-C. Tu, and J. M. Cioffi, "A discrete multitone transceiver system for HDSL applications," *IEEE JSAC*, vol. 9, Aug. 1991, pp. 895-908.
[23] P. S. Chow *et al.*, "A multicarrier E1-HDSL transceiver system with coded modulation," *European Trans. Telecomm.*, vol. 4, no. 3, May-June 1993, pp. 257-266.
[24] J. G. Proakis, "Digital Communications," (New York: McGraw-Hill, 1989).
[25] T. Walzman and M. Schwartz, "Automatic equalization using the discrete Fourier domain," *IEEE Trans. Inform. Theory*, vol. IT-19, Jan. 1973, pp. 59-68.
[26] E. R. Ferrara, Jr., "Frequency-domain adaptive filtering," chapter 6 in Adaptive Filters, C. F. N. Cowan and P. M. Grant (Editors), (Englewood Cliffs, NJ: Prentice-Hall, 1985), pp. 145-179.
[27] L. R. Rabiner and B. Gold, "Theory and Application of Digital Signal Processing," (Englewood Cliffs, NJ: Prentice-Hall, 1975).
[28] W. C. Lindsey, "Synchronization Systems in Communication and Control," (Englewood Cliffs, NJ: Prentice-Hall, 1972).
[29] H. Sari *et al.*, "Baseband equalization and carrier recovery in digital radio systems," *IEEE Trans. Commun.*, vol. COM-35, no. 3, March 1987, pp. 319-327.
[30] D. D. Falconer, "Jointly adaptive equalization and carrier recovery in two-dimensional digital communication systems," *Bell Syst. Tech. J.*, vol. 55, no. 3, March 1976, pp. 317-334.
[31] G. Karam et al., "All-digital implementation of the carrier recovery loop in digital radio systems," ICC '91 Conf. Rec., vol. 1, June 1991, Denver, CO, pp. 175-179.
[32] H. Sari, G. Karam, and I. Jeanclaude, "Channel equalization and carrier synchronization in OFDM systems," in Audio and Video Broadcasting Systems and Techniques, R. De Gaudenzi and M. Luise: Editors, (Amsterdam: Elsevier, 1994), pp. 191-202.
[33] H. Sari, G. Karam, and I. Jeanclaude, "Analysis of orthogonal frequency-division multiplexing for mobile radio applications", VTC '94 Conf. Rec., vol. 3, June 1994, Stockholm, pp. 1635-1639.
[34] H. Sari, G. Karam, I. Jeanclaude, "Frequency-domain equalization of mobile radio and terrestrial broadcast channels," GLOBECOM '94, Nov. 1994, San Francisco.
[35] A. A. M. Saleh, "Frequency-independent and frequency-dependent nonlinear models of TWT amplifiers," *IEEE Trans. Commun.*, vol. COM-29, no. 11, Nov. 1981, pp. 1715-1720.

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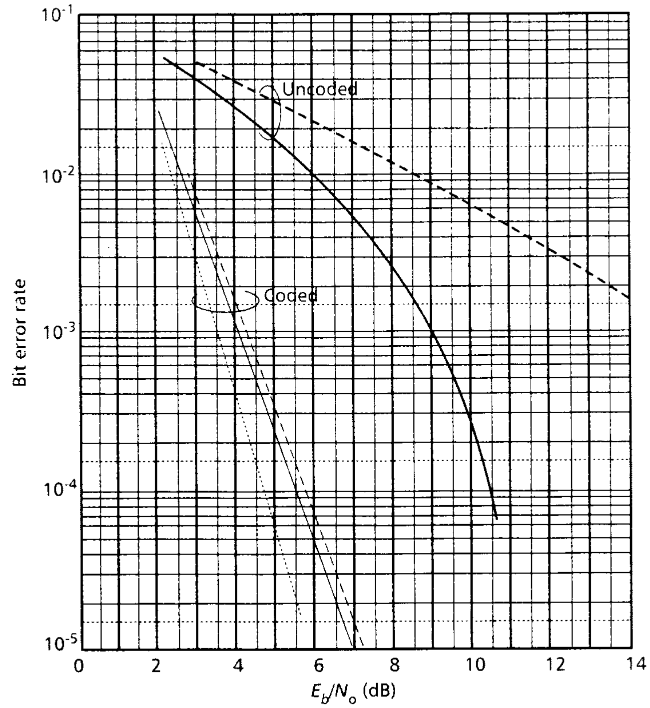
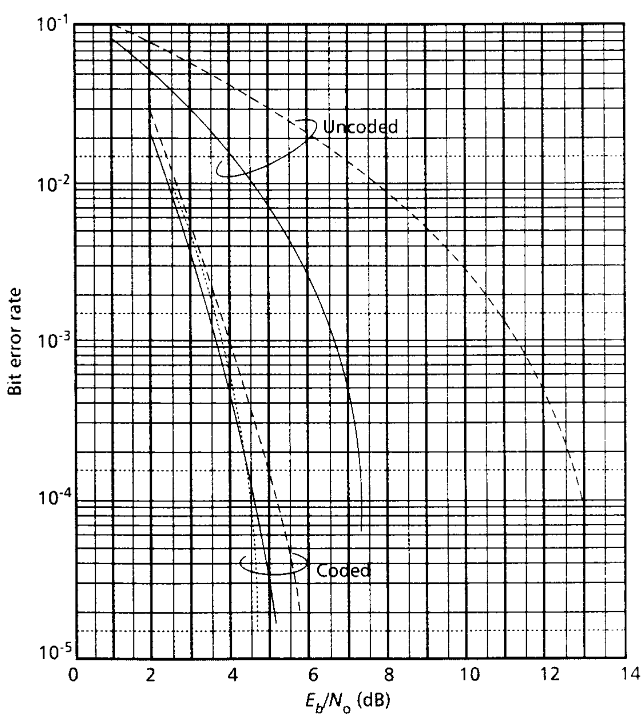


■ **Figure 9.** *Bit error rate performance of different systems on channel A.*



■ **Figure 10.** *Bit error rate performance of different systems on channel B.*