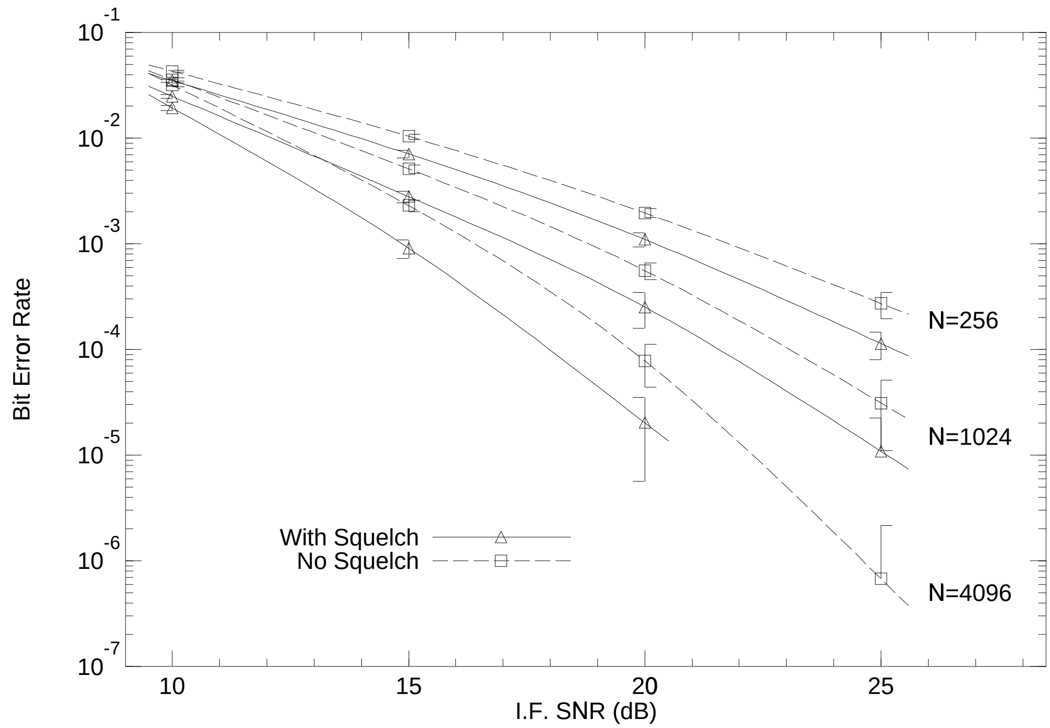

Figure 6.10: BER performance with squelch.

the block); or

- until there are no changes in the data estimate between iterations, indicating that the method has converged to an incorrect solution; or
- until a limit on the number of iterations is reached.

Unfortunately, this method does not always converge to the correct solution. The reason is that errors in the estimates of the data tend to be self-reinforcing. Errors in the data estimate produce a substitution signal that tends to produce that same erroneous estimate when demodulated. The errors corrected at each iteration must produce a new substitution signal that results in a different data estimate on the next iteration or the data estimate has converged.

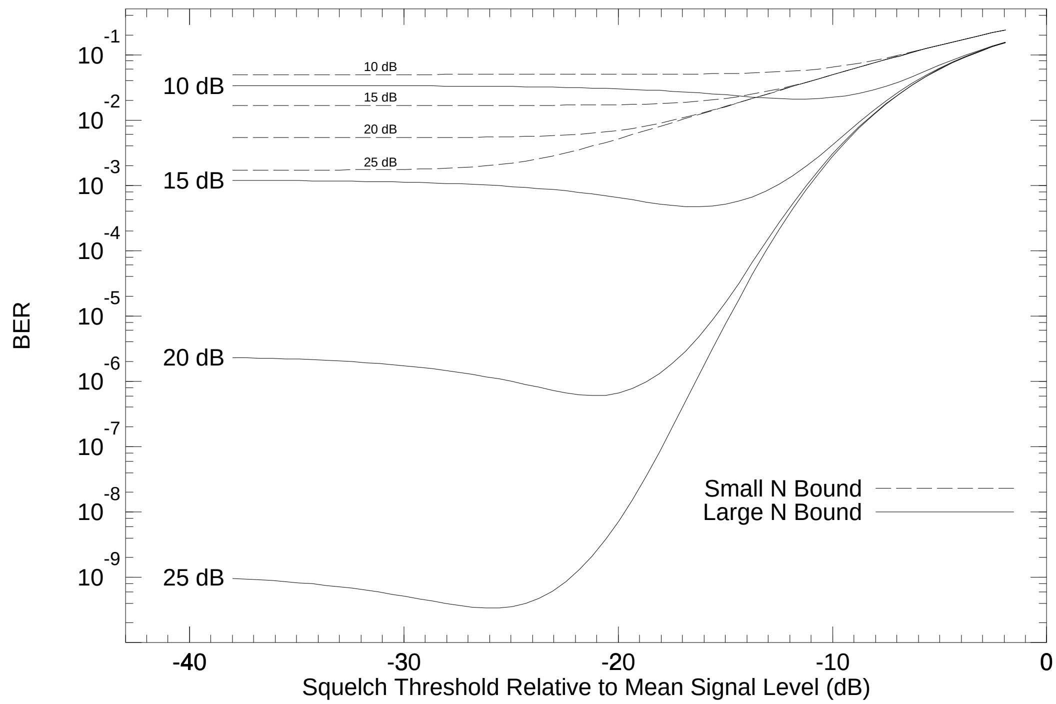

Figure 6.9: BER bounds as a function of squelch threshold.

(5) using the faded received signal (4). This estimate of the transmitted data may contain errors due to the additive noise and the fading-induced crosstalk interference between the subchannels. An OFDM modulator (inverse DFT) can then be used to produce an estimate of the transmitted signal (6) from the initial data estimate (5). Portions of this estimate of the transmitted signal (7) can then be substituted into the faded portions of the received signal to form a composite signal (8). The new composite signal can be used to make another estimate of the data. Since the composite signal has fewer fades, there should be less crosstalk interference and this should result in a second data estimate with fewer errors.

If this second data estimate has fewer errors, then a more accurate estimate of the lost (faded) signal samples can be made. The preceding steps are then repeated until

- there are no more errors (determined by an error detecting code embedded within

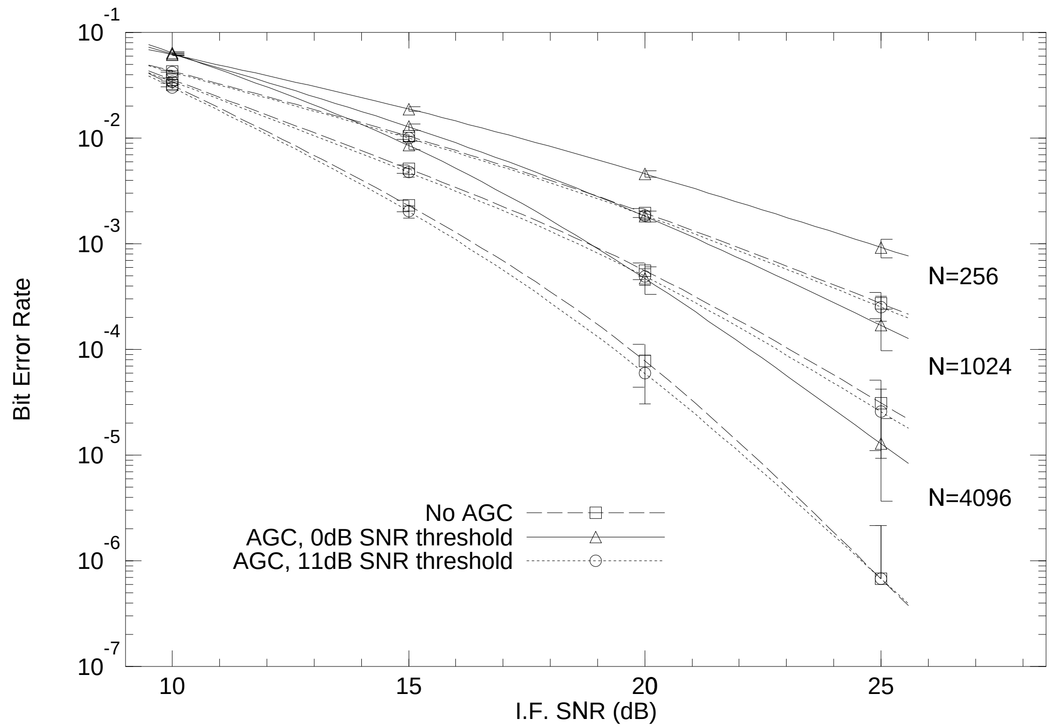

Figure 6.8: BER performance with AGC.

6.6 Decision Feedback Correction

As explained in Chapter 3, fading causes interference between the OFDM subchannels. A new method has been developed to reduce this interference. The method uses a correction signal that is generated from the received data – a process similar to decision feedback equalization – and is therefore called *decision feedback correction* (DFC).

6.6.1 Description of the Method

Figure 6.11 shows a flowchart of the signal processing steps while Figure 6.12 gives an example of the waveforms at the different steps of the DFC processing. The OFDM modulator converts the original data (1, Figure 6.12) into an OFDM signal (2). The envelope of the received signal (3) fades and occasionally falls below a DFC threshold. The first step in the DFC process is to make an initial estimate of the transmitted data

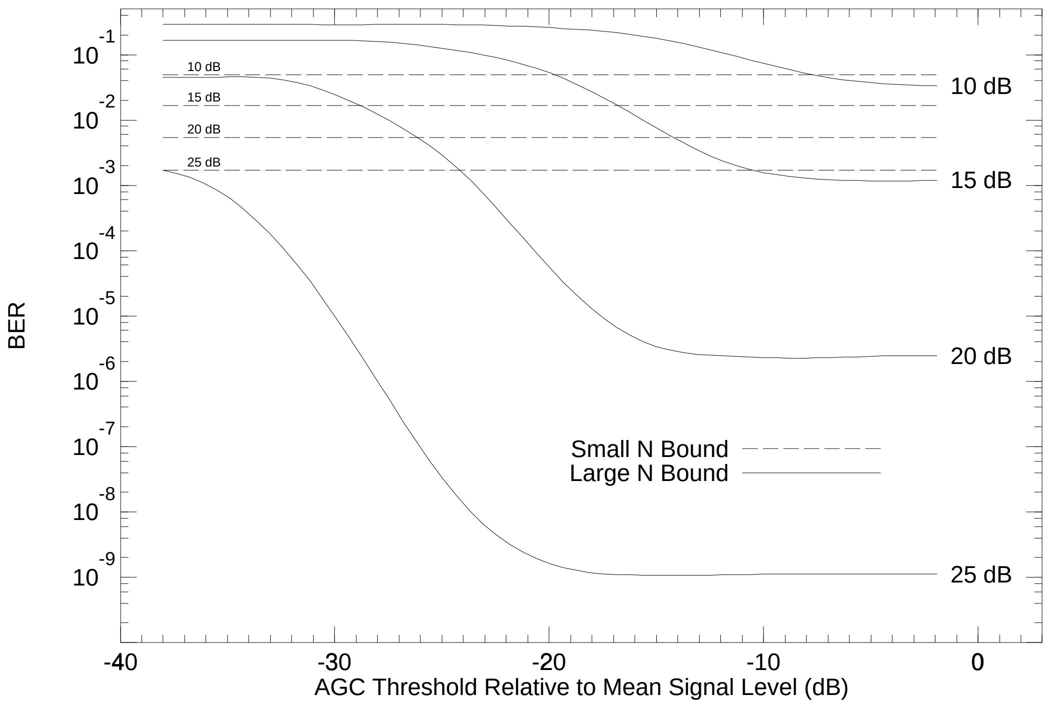

Figure 6.7: BER bounds as a function of AGC threshold.

those portions of the output that are most noisy. Unfortunately, the additional distortion due to eliminating the signal also increases the crosstalk interference between subchannels. Squelch could improve the performance of OFDM if the reduction in noise power could offset the increase in crosstalk interference.

The squelch threshold is the IF SNR below which the receiver output is set to zero. The bounds for large and small block sizes were evaluated for various squelch thresholds to investigate the effect of the threshold on the BER. The results, shown in Figure 6.9, indicate that squelch with a threshold of about 0 dB IF SNR might produce a small improvement in BER performance.

Figure 6.10 shows the BER results of a simulation using a squelch threshold at 0 dB SNR. As predicted by the bounds on the BER for large block sizes, the use of squelch at 0 dB SNR provides a small improvement.

signal [21,27].

As described in Chapter 3, amplitude variations in the received baseband OFDM signal cause crosstalk between the subchannels. The use of AGC might reduce this crosstalk.

A problem with using AGC on a fading channel is that the receiver's output noise level increases as the AGC amplifier gain increases to compensate for fading. OFDM averages the noise power received during the block and thus even a short increase in noise power can result in a significant contribution to the average noise power. The gain of the AGC amplifier must therefore be limited to prevent large noise contributions when the received signal undergoes deep fades. This gain limit reduces the baseband noise output during deep fades but increases the interference between subchannels because of limited correction of the fading. This topic has been studied for the OFDM/SSB channel [1].

The AGC threshold is the SNR below which the AGC gain is fixed. The bounds for large and small block sizes (see Chapter 4) were computed for various AGC thresholds to investigate the effect of the threshold on the BER. The results, shown in Figure 6.7, indicate that AGC would not significantly improve the BER performance of this system.

The BER performance of the OFDM/FM system with AGC was evaluated with simulations. AGC thresholds of 0 dB SNR and 11 dB SNR were tested. The results, shown in Figure 6.8, show a degradation of several dB with an AGC threshold at 0 dB SNR and negligible improvement with a threshold at 11 dB SNR.

6.5 Squelch

The receiver's squelch circuit is designed to shut off the audio output when no signal is being received. Squelch is normally used as a convenience for the user and to reduce power consumption.

Squelch reduces the average noise power output of the receiver because it removes that block interleaving increases the corrected bit error rate of the RS code by a very small amount. This is what would be expected if the errors were already approximately randomly distributed within the OFDM block. The sub-block length was chosen to be the smallest integer greater than the square root of the number of bits in the code word.

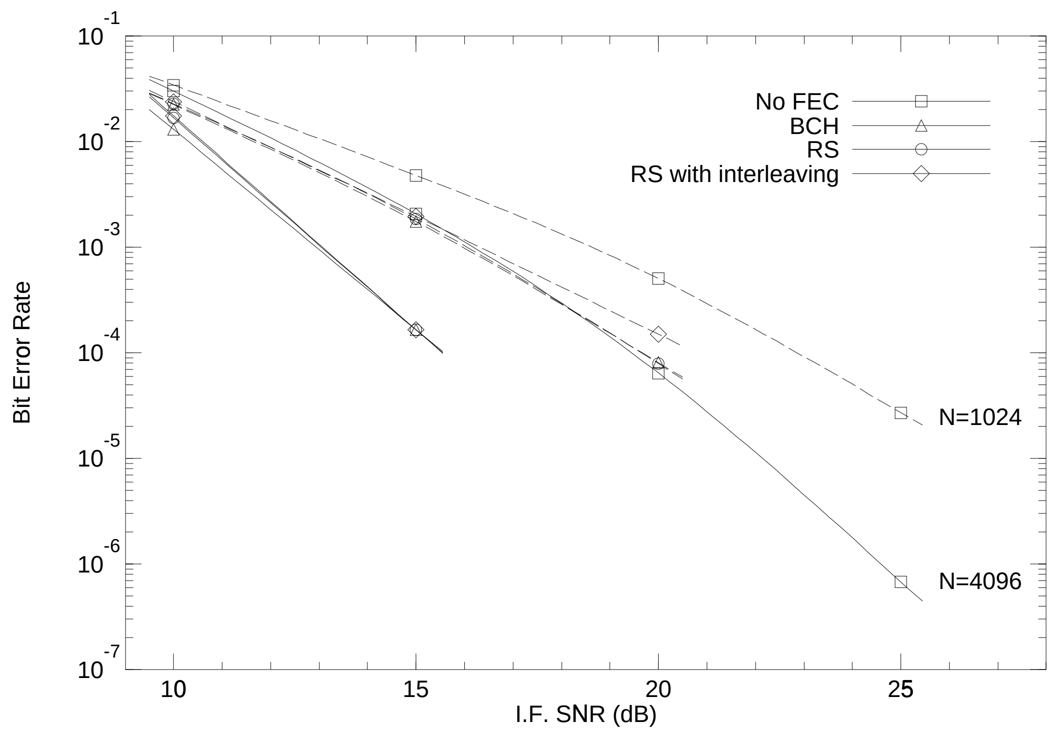

Figure 6.6: BER performance with FEC coding.

6.4 Automatic Gain Control

The purpose of an automatic gain control (AGC) circuit is to increase the receiver's gain during fades in order to keep the receiver output level constant. AGC is implemented as an amplifier whose gain is controlled by the received signal level. AGC is not normally required by FM receivers because, as shown by the SN curves (Figure 5.8, p. 53), the output signal level is constant when the IF SNR is above threshold. However, AGC is required for SSB reception of fading signals to reduce the distortion of the baseband

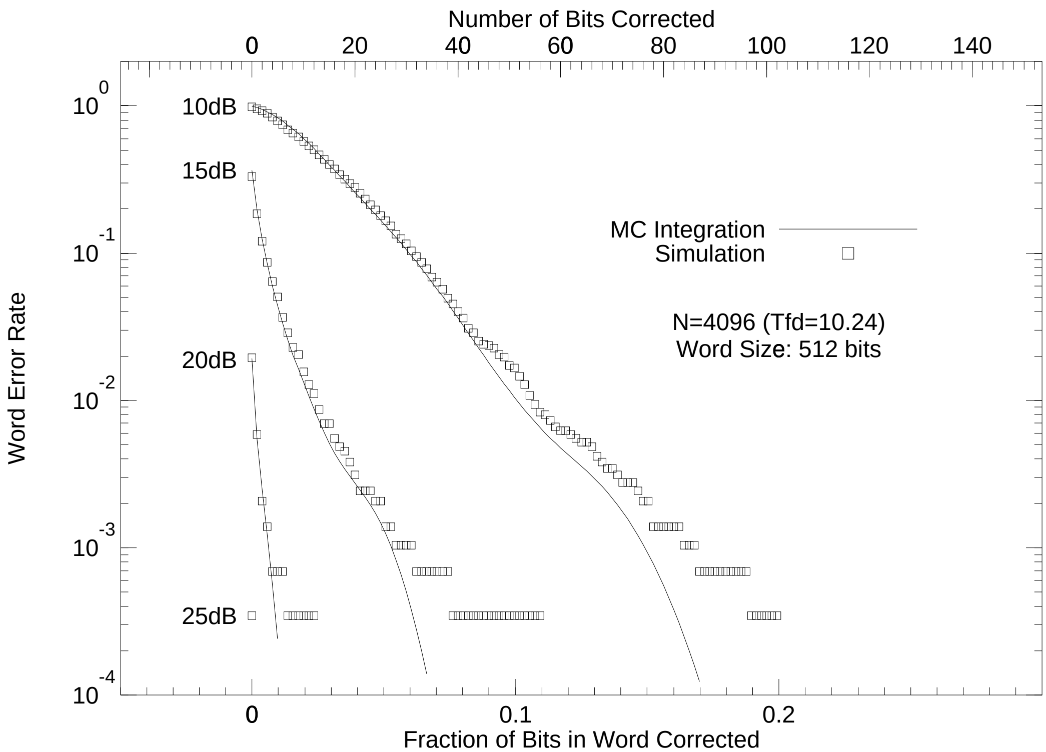

Figure 6.5: WER as function of corrected errors for block size of 4096 samples.

For a block size of 1024 the BCH code has a word size of 511 bits and uses 252 parity bits to correct up to 28 bit errors[2]. The RS code has a word size of 63 6-bit digits (378 bits) and uses 32 parity digits to correct up to 16 digits containing errors. Again, both codes give approximately the same performance. They reduce the required SNR by about 2 dB at a BER of 10^{-2}, 2.5 dB at 10^{-3}, and 3 dB at 10^{-4}. As before, the coding gain is about 3 dB less.

Block interleaving is a method used on burst-error channels to produce a more even distribution of errors within a block. The block is logically divided into a number of sub-blocks. Instead of transmitting the bits in the block sequentially, the first bit of each of sub-block is transmitted, followed by the second bit of each sub-block, and so on. This procedure helps to spread error bursts among the sub-blocks. Figure 6.6 shows

[2] The actual BCH code that corrects 31 errors requires slightly fewer (234) parity bits.

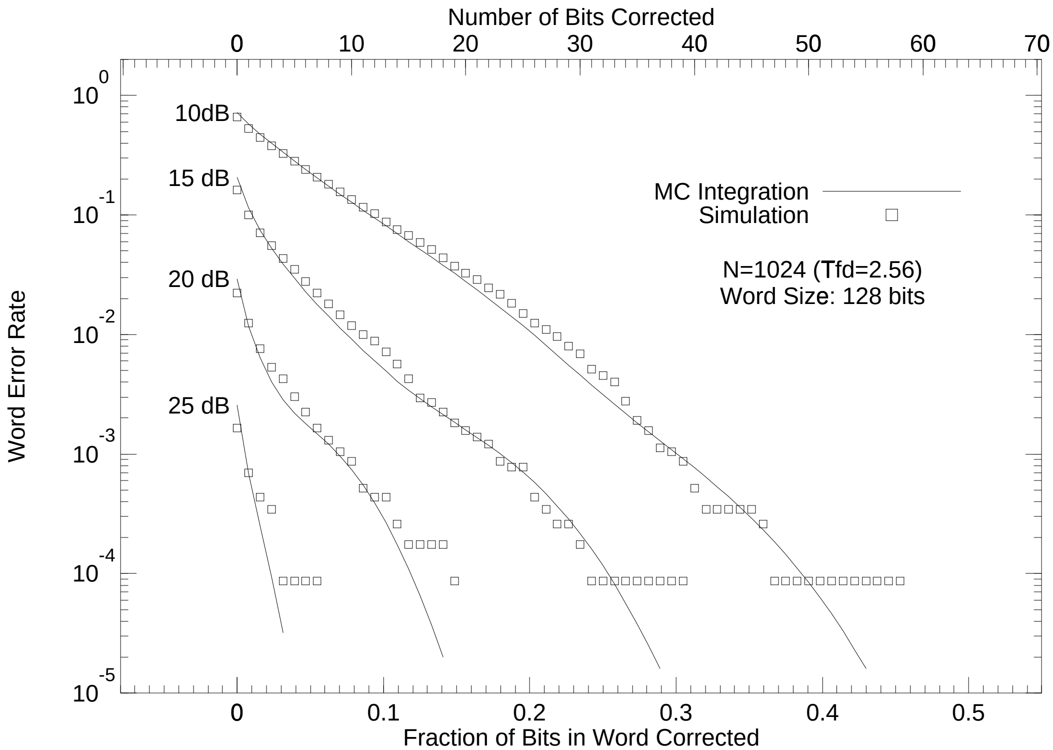

Figure 6.4: WER as function of corrected errors for block size of 1024 samples.

and the decoder outputs the bits as they were received.

Figure 6.6 shows the effect on the BER performance of using rate $\frac{1}{2}$ RS (burst-error) and BCH (random-error) codes with OFDM block sizes of 1024 samples (Tf_d= 2.56, 512 bits) and 4096 samples (Tf_d= 10.24, 2048 bits).

For a block size of 4096 the BCH code has a word size of 2047 bits and uses 1023 parity bits to correct up to 93 bit errors[1]. The RS code has a word size of 255 (8-bit) bytes (2040 bits) and uses 128 parity bytes to correct up to 64 bytes containing errors. Both codes give approximately the same performance. They reduce the required SNR by about 2 dB at a BER of 10^{-2}, 3 dB at 10^{-3}, and 4 dB at 10^{-4}. The improvement in E_b/N_0 (the coding gain) is 3 dB worse than these values because the coding approximately doubles the energy per information bit.

[1]The actual BCH code that corrects up to 93 errors probably requires fewer parity bits.

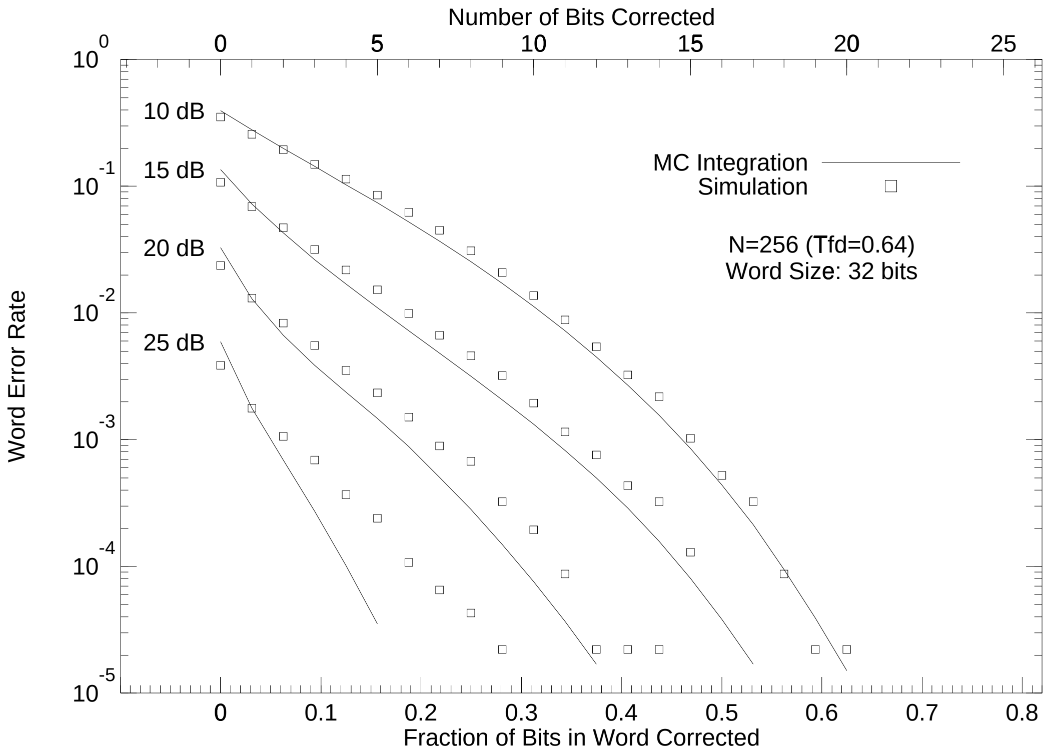

Figure 6.3: WER as function of corrected errors for block size of 256 samples.

6.3.2 BER Performance of Block Codes

Among the most popular FEC block codes are the binary Bose-Chaudhuri-Hocquenghem (BCH) and Reed-Solomon (RS) codes [69]. A binary BCH block code uses codewords of $2^m - 1$ bits. To allow correction of up to t errors, at most mt of these bits are used for parity and the remainder can be used for data. A Reed-Solomon (RS) code uses codewords of $2^m - 1$ m-bit digits. To allow correction of up to t digits, $2t$ of the digits are used for parity. RS codes are often used for burst error channels because they can correct any number of bit errors in each digit.

The effect of using FEC on OFDM was tested using the equivalent baseband channel simulation. For simplicity, the received data bits were corrected by comparing them to the transmitted data instead of implementing the actual coders and decoders. This approach assumes that when an uncorrectable error is detected no error correction is attempted patterns introduced by the channel. FEC has been used for mobile radio data communication because the probability that errors will occur in a long data transmission is quite high [68].

6.3.1 Testing the Independent Error Assumption

To predict the performance of an FEC block code it is helpful to know the distribution of the number of bit errors per code word. In Chapter 5 it was shown that the independent error assumption could be used to predict the WER for OFDM/FM. In this section it is shown that the independent error assumption can be used to predict the distribution of the number of errors per word and thus the performance of an FEC block code.

If bit errors in a code word of k bits occur independently with probability p, the probability of e errors, $P_k(e)$, is binomially distributed:

$$P_k(e) = \binom{k}{e} p^e (1-p)^{k-e}. \tag{6.1}$$

It is necessary to average over the block BERs to predict the average distribution because the BER (p) will be different for each block. This was done by averaging $P_k(e)$ using p's generated with the Monte-Carlo integration program.

A simulation was then run and the actual distribution of the number of errors per word was measured to verify the results obtained with the binomial distribution prediction. The simulation program generated a file of error-free run lengths and a second program then computed the distribution of the number of errors in each word.

Figures 6.3, 6.4 and 6.5 compare the measured and computed distributions. These graphs shows the results for block sizes of 256, 1024, and 4096 samples (normalized durations, Tf_d, of 0.64, 2.56 and 10.24). In each case the bits in the block were split into four words. The simulation length was 1,474,560 bits for each OFDM block size. The distributions obtained from the simulations are very close to those obtained by assuming independent errors within a block.

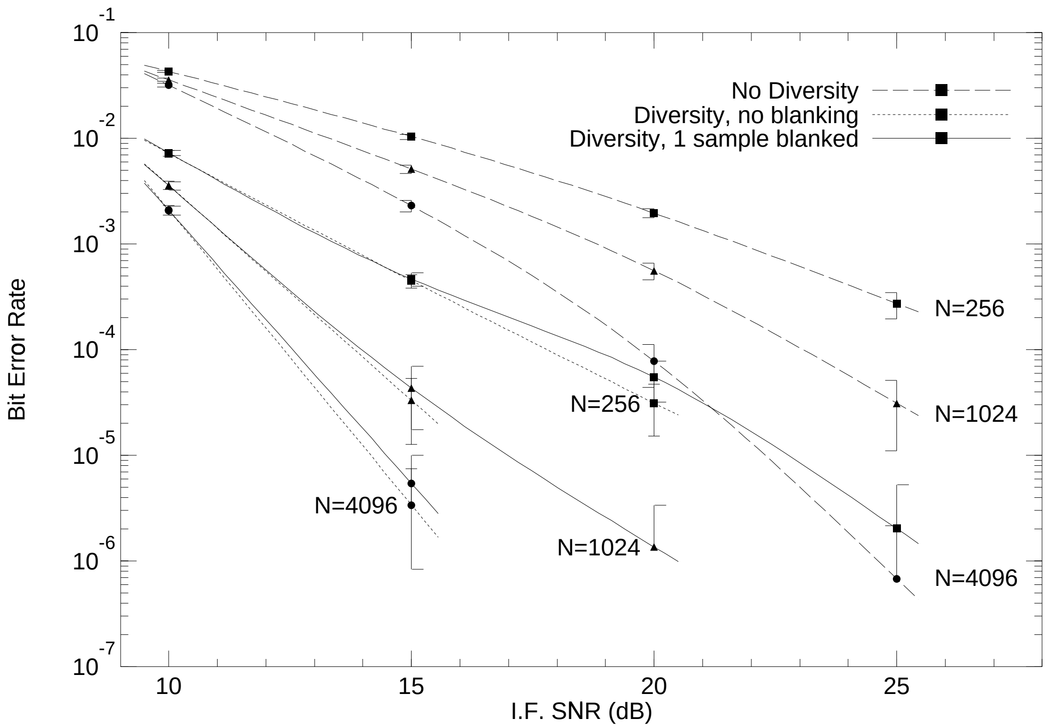

Figure 6.2: BER performance with and without switching diversity. $f_d = 20$ Hz.

Switching diversity also improves the BER performance of FSK systems but the phase transient due to switching "causes a high error probability for a bit received during a switch" [67]. As shown by the results in Figure 6.2, OFDM/FM does not suffer from this problem. For a 20 Hz Doppler rate and a switching threshold of 10 dB below the mean, equation 2.4 gives a level crossing rate of about 14 Hz which corresponds to one switch every 280 bits at the 4000 bps bit rate. A high bit error probability at each switch would have resulted in a much higher BER for the OFDM/FM system.

6.3 Forward Error Correction Coding

Forward error correction (FEC) coding is used in many digital communication systems. FEC coding adds parity bits to the data to allow the receiver to correct certain error discontinuity creates a noise impulse in the recovered baseband signal of an FM receiver. Switching diversity is well suited to a mobile radio system using OFDM since OFDM can average out this impulsive noise. The effect of using switching diversity on the BER performance of OFDM/FM was examined by simulating a switching diversity receiver. Two uncorrelated fading waveforms were generated and the receiver switched between the signals when the signal level on the current antenna dropped below a threshold.

The effect of switching between antennas was simulated by blanking out one sample each time the antenna was switched. This is reasonable because a receiver could be designed to blank (set to zero) the audio output during antenna switching. Therefore zero is the worst-case amplitude of the noise impulse. Since solid-state switches can switch between antennas in a fraction of a microsecond [66], the RF switching time can be made negligible. As a result, the duration of the phase discontinuity (and thus of the noise impulse) after the IF filter will be determined by the IF bandwidth, typically 15 kHz. A single sample (125 μs) is thus likely to be the worst-case duration for the switching discontinuity.

The level of the switching threshold is a compromise between the degradation caused by excessive switching if the threshold is too high and the degradation caused by a low signal level if the threshold is too low. In general, the optimum switching threshold is a function of the IF SNR and the number of diversity branches. For FSK and two-branch diversity a threshold of between 18 dB (13 dB IF SNR), and 20 dB (20 dB IF SNR) below the mean signal power was found to give best performance [67]. However, since OFDM/FM is more tolerant of antenna switching, the optimum threshold might be expected to be somewhat higher. A threshold set at 10 dB below the mean was used. The switch stays on each antenna for a minimum of four samples before switching. This is done to avoid continuous switching when both signals are below threshold.

Figure 6.2 shows the BER performance without diversity and with switching diversity for the case of two antennas.

6.2 Switching Diversity

Space diversity is used on many fading channels to reduce the depth and duration of fades. Space diversity involves using two or more antennas spaced far enough apart (more than about one-half wavelength for mobile receivers) so that the fading at the antennas is not highly correlated. When one antenna is in a fade another one may still be receiving a strong signal. The severity of the fading can be reduced by adding the signals from the different antennas or by switching between them. Time diversity and frequency diversity are also common, but space diversity has the advantages of being insensitive to the fading rate and of not requiring extra bandwidth.

There are several methods of combining the signals from the different antennas [3]. A common and inexpensive type of space diversity is switching diversity. Switching diversity involves switching between antennas whenever the received signal level on the antenna in use drops below a specified threshold. Switching diversity is less expensive to implement than other types of space diversity because only one receiver is required. Figure 6.1 shows a block diagram of a switching diversity receiver.

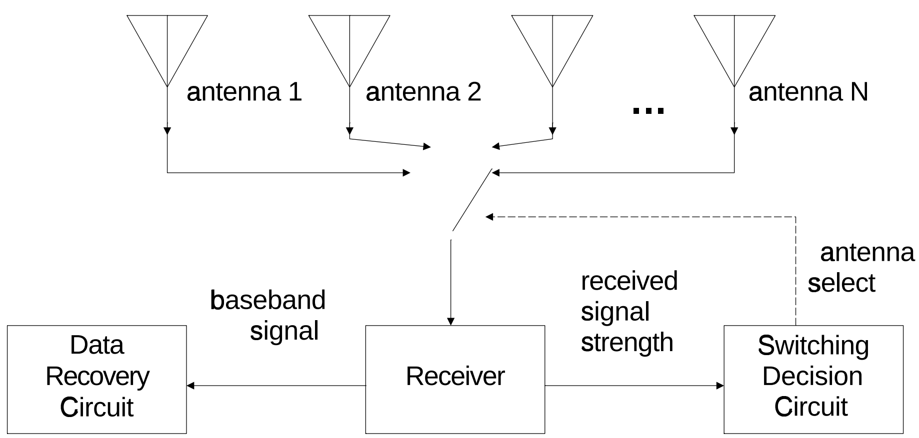

Figure 6.1: Switching diversity receiver.

When the receiver switches between antennas the received signal undergoes a phase shift due to the difference in propagation delay between the two antennas. This phase

Chapter 6

Improving Performance

6.1 Introduction

This chapter examines four methods that might be used to improve the performance of OFDM/FM.

The first method is switching diversity, a type of space diversity. Switching diversity is widely used but it works especially well with OFDM because OFDM is resistant to the transients caused by switching between antennas.

The second method, forward error correction (FEC) coding, is also widely used. It is shown that the distribution of the number of errors in an FEC code word can be predicted by assuming that bit errors occur independently within each OFDM block. The BER performance of two BCH and Reed-Solomon rate $\frac{1}{2}$ codes are also given.

Squelch and AGC were described in Chapter 3. In this chapter the BER bound computations are used to find the optimum squelch and AGC thresholds for large block sizes. It is shown in this chapter that neither technique significantly improves the performance of the OFDM/FM system studied.

The final section describes a novel signal processing technique, decision feedback correction (DFC), that reduces the crosstalk between the OFDM subchannels caused by fading. DFC greatly improves the performance of OFDM/FM when large block sizes are used.

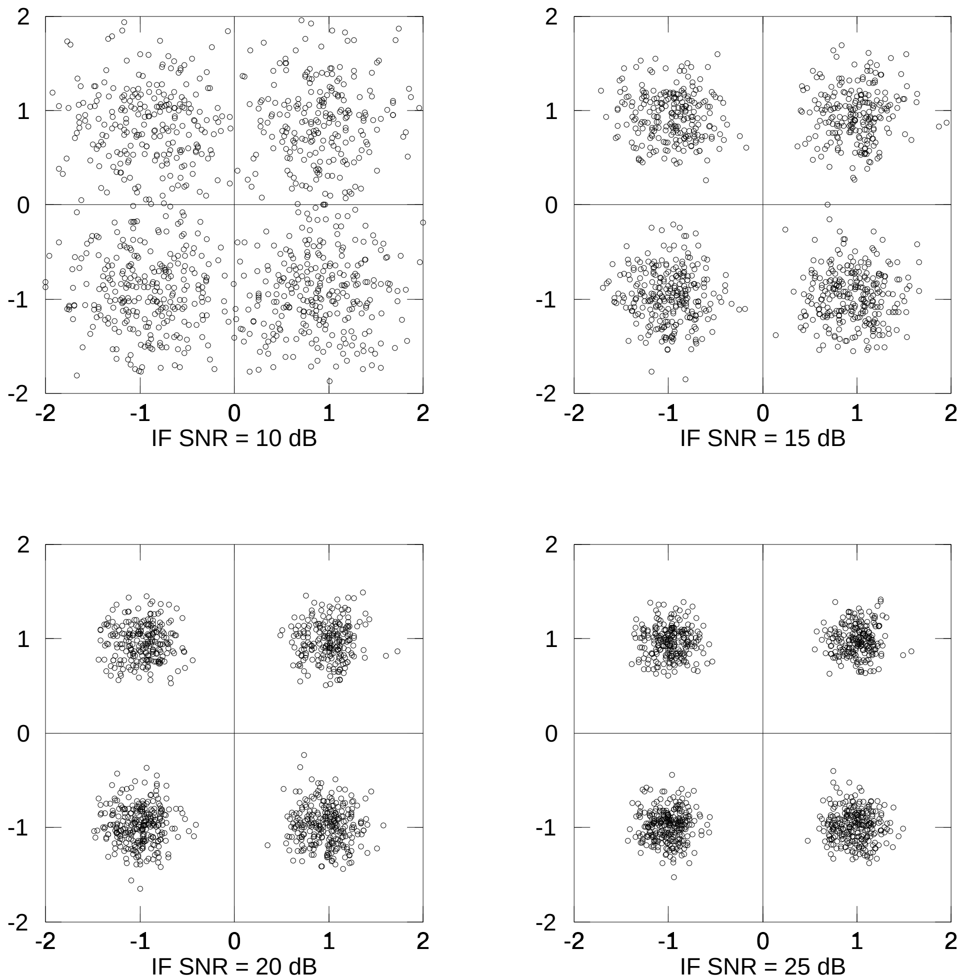

Figure 5.15: Example of received data values.

Equipment and techniques were developed to measure the SN curves, and the BER and WER performance of OFDM/FM over a non-frequency-selective Rayleigh-fading channel. The experimental work demonstrated the need to consider the practical limitations of FM transmitters such as clipping, bandwidth restrictions, preemphasis/deemphasis, and the baseband noise spectrum.

Although bit errors are not independent, it was shown that use of the independent error assumption gives a good approximation to the measured WER. It was also shown that random FM should not be a significant problem.

Performance measurements were made only on one transmitter and receiver combination. Other units or models may have different performance. NBFM equipment that has been modified or designed for digital transmission by modifying the audio processing stages should have better performance. Thus, although the experimental measurements do not show the range of performance that can be achieved with OFDM/FM, the experimental results help to verify the analysis and demonstrate the feasibility of OFDM/FM.

assumption.

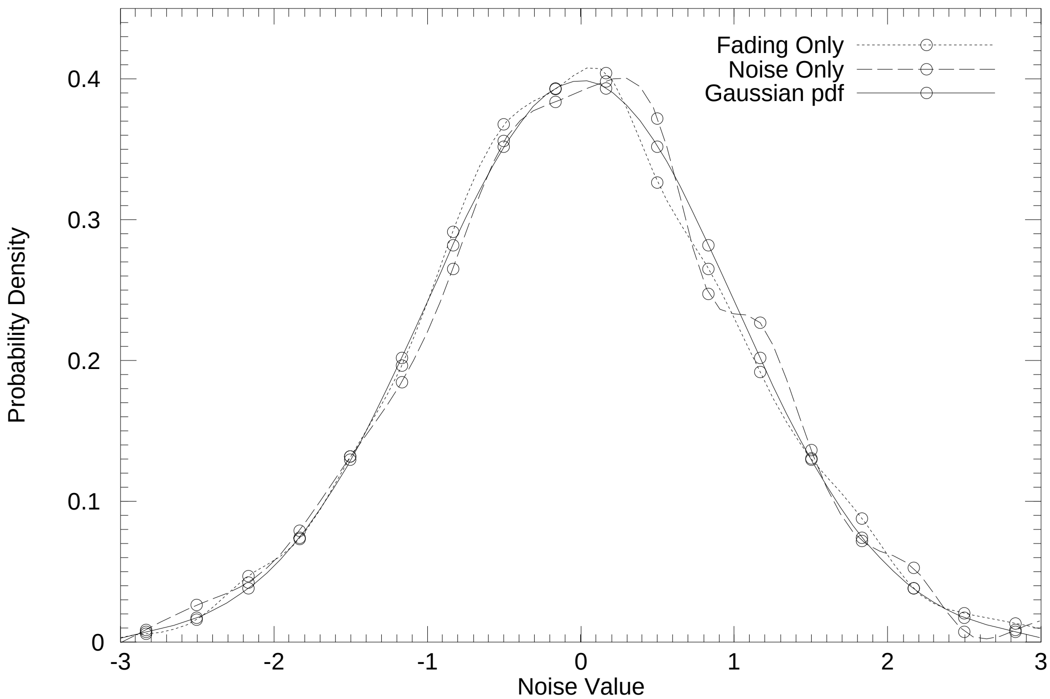

Figure 5.14: Example of probability density of the noise in the frequency domain.

5.8 Example of Received Signal Values

Figure 5.15 shows plots of the received complex data values for four IF SNRs. The plots show the effect of the additive and crosstalk noise. The noise distribution is circularly symmetric as would be expected from two independent and equal-power components.

5.9 Conclusions

The measurements described in this chapter were designed to measure the SN curves of a typical NBFM radio channel, to verify experimentally the BER and WER results obtained through numerical methods, and to check for possible unforeseen implementation problems.

the baseband SNR to about 20.5 dB. Since the limiting baseband SNR due to clipping during the BER measurements was 17.5 dB, random FM noise should not be a significant problem for OFDM/FM with QPSK (4-QAM) subchannel modulation.

5.7 Probability Distribution of the Baseband Noise

An assumption made in Chapter 3 is that the crosstalk interference between subchannels due to fading is normally distributed. The distribution of this crosstalk noise was measured to confirm this assumption. A simulation was performed in which a block of 4096 samples was sent over the fading channel but no noise was added. The probability distribution (a histogram) of the noise due to the crosstalk interference was measured and normalized by dividing the measured noise values by the standard deviation. This normalized distribution is plotted in Figure 5.14. This figure also shows the normal probability distribution (unit variance and zero mean). The measured distribution approximates the normal distribution.

As described in 2, the probability distribution of the baseband noise output of an FM discriminator is not Gaussian. However, the averaging done by the DFT in the OFDM demodulation process produces noise in the frequency domain whose distribution is approximately normal. A second measurement of the probability distribution of the noise after the OFDM demodulation was made to confirm this assumption. A second simulation was done in which there was no fading but impulse noise was added to the OFDM signal. The impulse noise used was composed of noise samples of the same level with an equal probability (0.05) of occurrence at each sample (see Section 4.5.3). This impulse noise waveform is not meant to simulate a realistic source of impulse noise, but only to verify that the averaging of the baseband noise by the OFDM demodulation process produces noise with a distribution that is approximately Gaussian. The measured distribution was normalized as before and is also shown in Figure 5.14. Again, the measured distribution approximates the Gaussian distribution and confirms the Gaussian

5.6 Random FM

Random FM adds a noise component that increases with the Doppler rate and is independent of the IF SNR. At high Doppler rates the random FM noise power might limit the achievable baseband SNR. Available theoretical analyses of the baseband noise produced by random FM [3,48] cannot be used since they do not take into account the effect of deemphasis.

The baseband SNR at a high (nominally 75 dB) IF SNR was measured while the carrier was fading to see if random FM had a significant effect. The level of the modulating signal was reduced so that the noise due to random FM could be measured above the noise due to clipping. The modulating signal level was reduced by 3.5 dB so that the limiting baseband SNR increased to 24.2 dB. Figure 5.13 shows the measured random FM noise power as a function of Doppler rate.

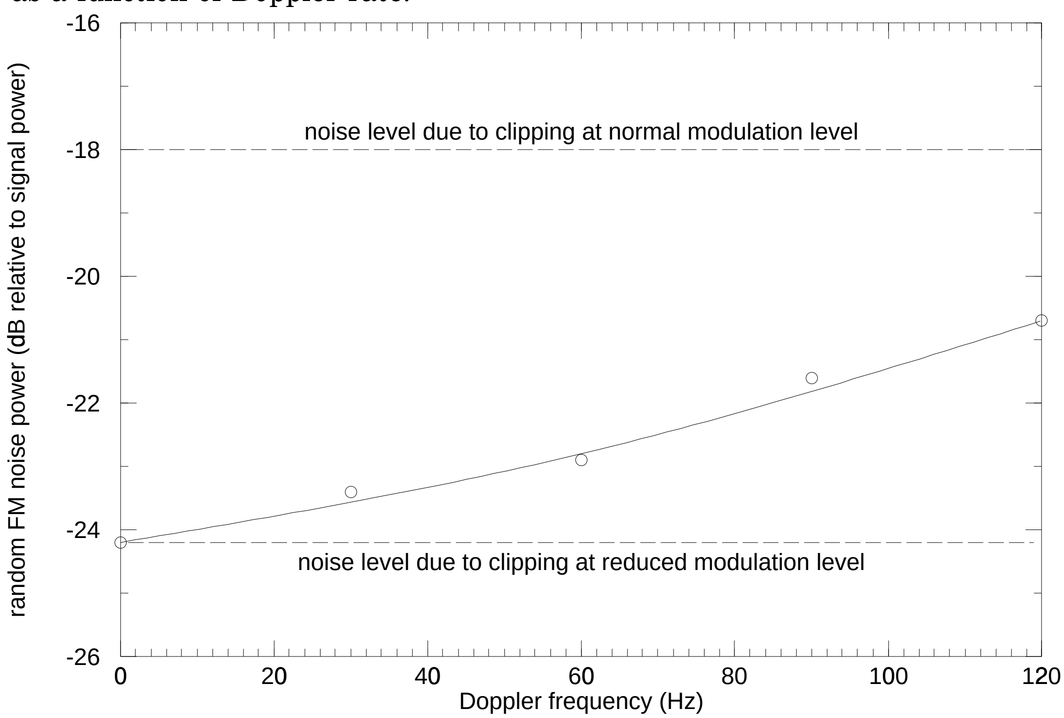

Figure 5.13: Random FM noise power versus Doppler rate.

At a Doppler rate of 120 Hz (over 140 km/h at 900 MHz) the random FM noise reduced

Test for Independence

The assumption that errors occur independently of each other is often used when evaluating the effectiveness of various error correction methods. A Wald-Wolfowitz statistical test for the independence of errors was performed on the bit error pattern to test the independence of errors [62,63,64,65]. This test is based on the expected number of runs (continuous sequences of bits with or without errors) for a given bit error rate. For a large number of runs the distribution of the number of runs (R) tends to a Gaussian with mean given by [63]:

$$\mathrm{E}\left[R\right] = 1 + \frac{2n_1 n_2}{n} \tag{5.1}$$

and a variance given by

$$\sigma_R^2 = \frac{2n_1 n_2 (2n_1 n_2 - n)}{n^2 (n-1)} \tag{5.2}$$

where n_1 is the number of bits with errors, n_2 is the number of bits without errors and n is the total number of bits.

The test for independence on the OFDM bit error patterns was done by normalizing the number of runs in each OFDM block by using the above expression for the mean and variance and the measured values of n_1 and n_2 for each block. This normalized count of the number of runs was averaged over several hundred blocks and was repeated for several block sizes and SNRs. Only blocks with 10 or more runs were considered so that the number of runs would be approximately normally distributed. This average normalized number of runs was significantly (at the 0.95 confidence level) negative (less than zero) except for the two tests for SNRs of 25 dB for block sizes of 1024 and 4096 samples (the Doppler rate was 20 Hz). These results indicate that there were fewer runs than would have been expected if the errors had occurred independently.

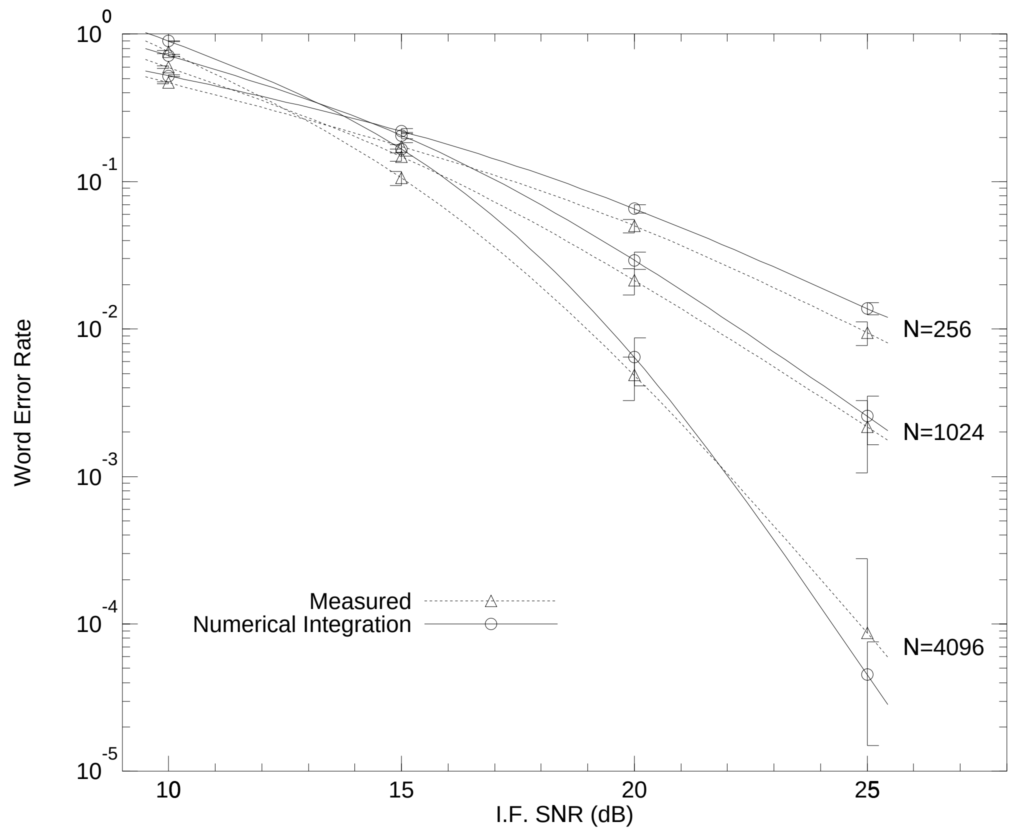

Figure 5.12: Measured and computed WER.

bursty errors will have a lower WER than the random error channel because the errors will tend to be concentrated into fewer words. By making blocks longer and thus putting more words into each block there is less variation in the $\mathrm{BER}_{\mathrm{block}}$ from word to word and the channel errors appear to be more random. Increasing the block length thus increases the WER because the bit errors are more random. However, increasing the block length also reduces the BER and this will reduce the WER.

At low SNRs the different block sizes have approximately the same BER as shown in Section 5.5.1. Thus at low SNRs using shorter blocks will produce more bursty errors and a lower WER than longer blocks. At high SNRs the longer blocks have much lower BERs than the shorter blocks and this leads to lower WERs for longer blocks. This explains why the WER curves intersect in Figure 5.12.

rate is changed by the same amount.

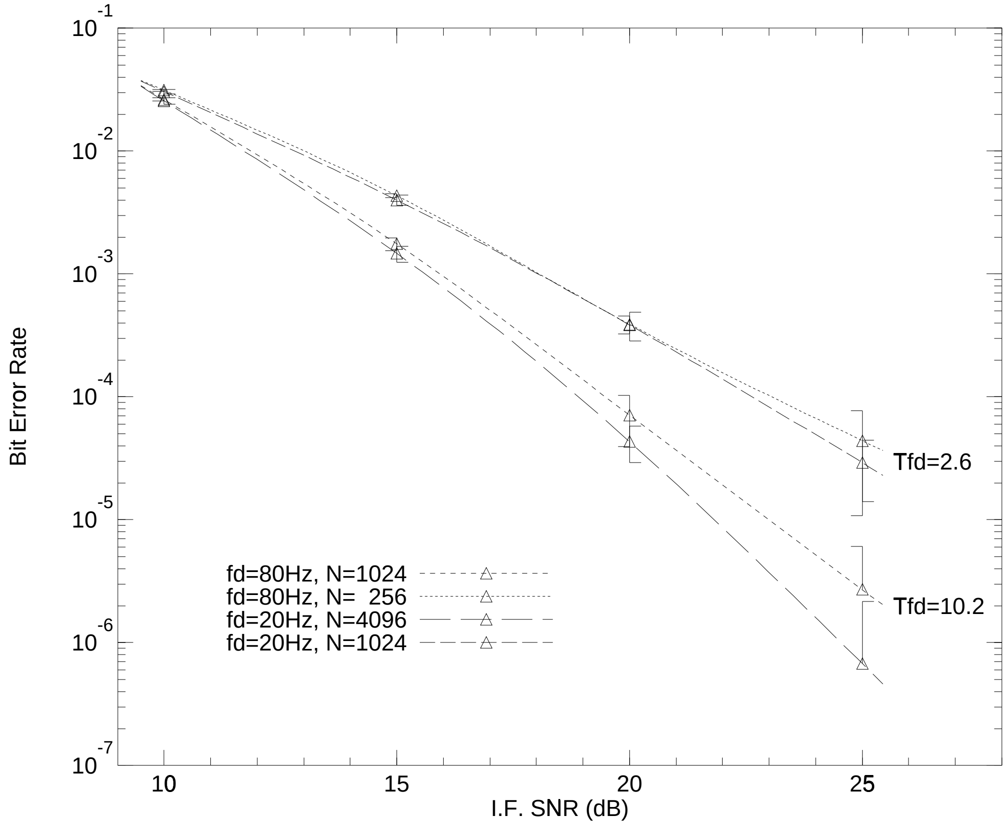

Figure 5.11: Measured BER for two values of Tf_d. Lower limits on error bars are not shown if lower limit is off the graph.

5.5.4 WER Results

Figure 5.12 compares the experimental WER measurements and the WER predicted with the Monte-Carlo integration program and the assumption that the errors occur independently (see Chapter 3). The two curves agree to within about 1 dB. This indicates that the independent error assumption can be useful in predicting the WER. The WER is always quoted for a word size of 128 bits.

Consider two channels that have the same BER but the errors on one channel tend to occur in bursts while the errors on the other channel occur at random. The channel with

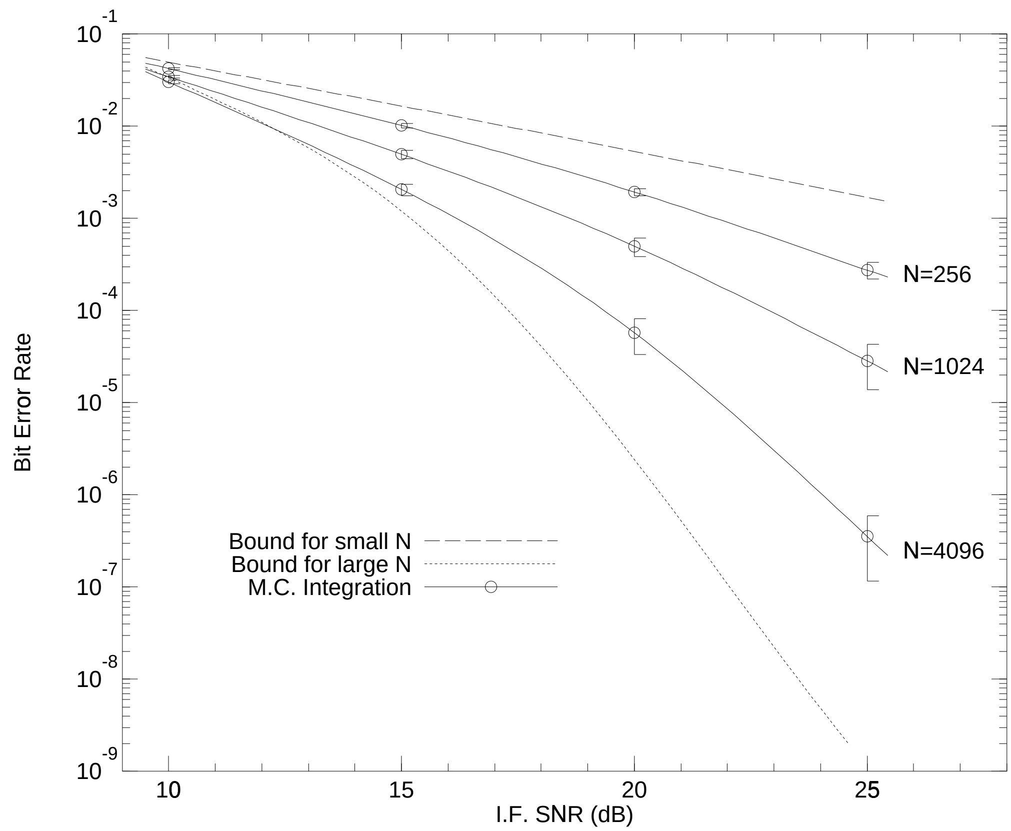

Figure 5.10: Bounds on the BER.

5.5.3 Effect of Block Duration and Doppler Rate (Tf_d)

The BER performance of OFDM is determined by the time averages of three variables during a block. If the duration of the block is made longer and time scale of the fading is increased by the same amount (the Doppler rate decreased), the time statistics of the averages will not change. It would therefore be expected that the OFDM BER performance is determined by the product, Tf_d, of the block duration T and the Doppler rate, f_d. This relationship should hold until the Doppler rate is high enough to produce a significant amount of random FM noise (section 5.6).

Figure 5.11 shows the measured and computed BER performance for Doppler rates of 20 and 80 Hz (100 km/h at 850 MHz) and block sizes of 256, 1024 and 4096 samples. As expected, the BER performance remains the same if either the block size or the Doppler

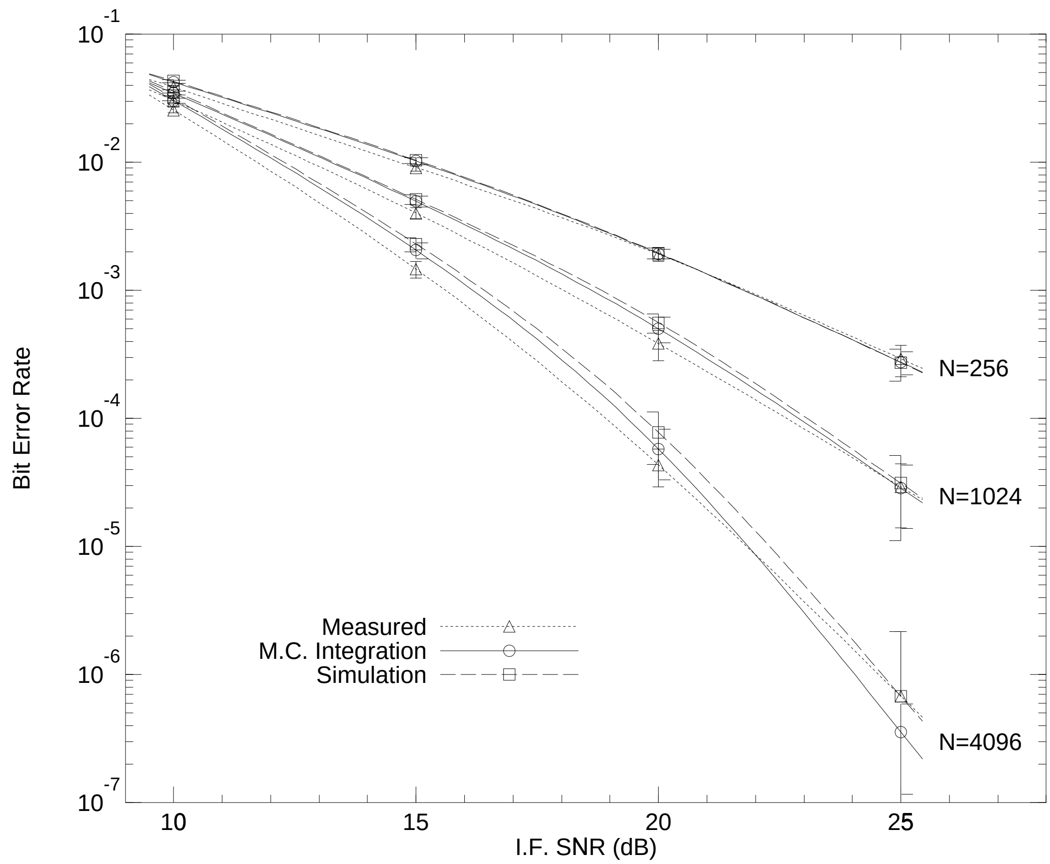

Figure 5.9: BER Results for $f_d = 20$Hz. The error bars show the 95 percent confidence intervals. The error bars offset to the right are for the Monte Carlo (M.C.) integration results, those to the left are for the simulation results, and those that are centered are for the measured results.

methods give results that are within about 1 dB.

5.5.2 Bound Results

Figure 5.10 shows the large- and small-block bounds (section 4.2) on the BER obtained using the parameter values given above. As expected, the results obtained in the previous section lie within the bounds.

Doppler rate	20 Hz
corresponding vehicle speed	25 km/h at 850 MHz
sampling frequency	8 kHz
baseband frequencies used	1 kHz to 3 kHz
bit rate (during block) (R)	4000 bps
samples per block	256, 1024, 4096 samples/block
block duration	32, 128, 512 ms
bits per block	128, 512, 2048 bits/block
IF SNR	10, 15, 20, 25 dB
IF noise bandwidth (W)	14.9 kHz
E_b/N_0	16, 21, 26, 31 dB

Table 5.1: Summary of parameter values used in experimental measurements.

5.5 BER and WER Measurements

The purpose of these measurements was to measure the BER and WER of an OFDM/FM system over the non-frequency-selective Rayleigh fading channel. The results are then compared against results obtained using the numerical techniques with the measured SN curves.

5.5.1 BER Results

Table 5.1 summarizes the values of some of the experimental parameters.

Figure 5.9 compares the BER results obtained using three methods: Monte-Carlo integration (section 4.3), baseband signal processing simulation (section 4.4), and measurements using the experimental channel. The Monte-Carlo integration and the baseband signal processing simulations used the measured SN curves shown in Figure 5.8.

The measurements were organized as 12 trials of 60 blocks per trial with 4096 samples per block. For the simulations and the experimental BER measurements, this represents about 1.5 million bits. For the Monte-Carlo integration this represents a fading waveform duration of about 6 minutes. As shown by the 95 percent confidence interval error bars [55,62], there are large uncertainties at low BERs. The three different BER evaluation power (S) was computed as the square of the mean of the received data values. The noise power was computed as the variance (second central moment) of the received data values. The signal and noise powers were measured over the subchannels that were used (1 to 3 kHz). The program could also display signal and noise power in certain frequency bands so that the distribution in frequency of the signal and noise powers could be measured.

5.4.8 SN Curve Results

The measured SN curves are given in Figure 5.8 along with those of an ideal discriminator.

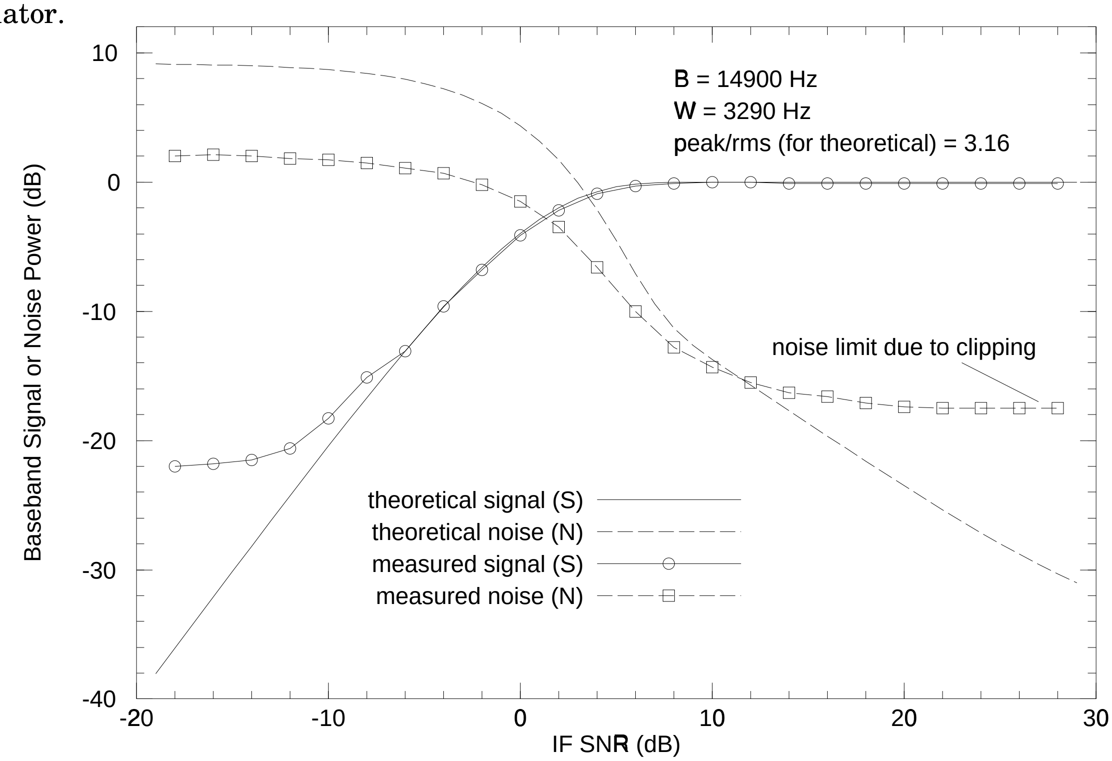

Figure 5.8: Theoretical and measured SN curves.

Each point is an average of 64 1024-sample measurements. The measurements are normalized to give a maximum signal (S) level of 0 dB.

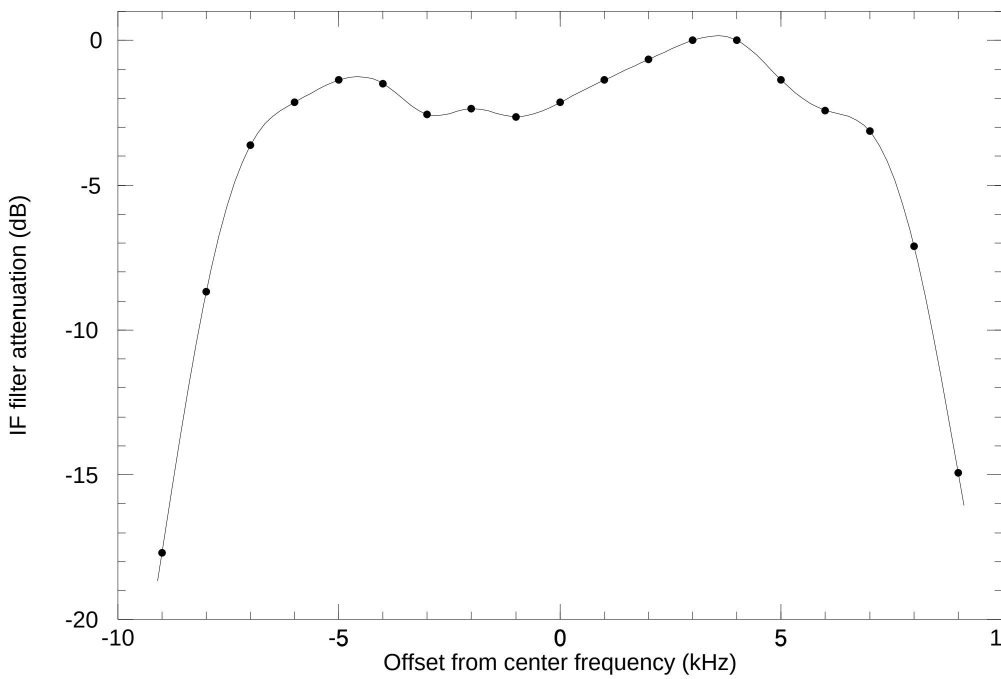

Figure 5.7: Measured IF filter response.

filter will depend on the baseband spectrum of the modulating signal.

Third, the effect of the clipping at the transmitter will depend on the probability distribution of the modulating signal. The signal used to measure the SN curves must have the same probability distribution as the OFDM signal in order to properly measure the amount of distortion due to clipping.

For these reasons it was necessary to use a signal with the same power spectrum and probability distribution as the OFDM signal. Since the OFDM signal itself has the appropriate spectrum and probability distribution, an OFDM signal was used as the test signal.

If there is no fading, there will be no crosstalk between subchannels due to fading and the noise received on each subchannel will be due solely to additive noise and distortion effects such as clipping.

The simulation program was modified to make SN curve measurements. The signal

5.4.6 Converting from IF SNR to E_b/N_0

There are two common ways to specify the quality of the IF signal. The first method is the ratio of the average signal power to the average noise power – the signal-to-noise ratio (SNR). This is the traditional method used for analog communication systems. The second method is the ratio of bit energy to noise power spectral density (E_b/N_0). E_b/N_0 is often used to compare the energy efficiency of digital communication systems. Both measurements are used in this thesis.

The FM baseband channel is defined by the baseband SN curves as a function of IF SNR. However, results are given as a function of E_b/N_0 when comparing OFDM with other digital modulation schemes.

To convert from SNR to E_b/N_0, the ratio of bit rate (R) to IF bandwidth (B) must be known. The bit energy, E_b (Joules, or Watts/Hz), can be computed as $E_b = S/R$, where S is the received signal power (Watts) and R is the data rate (bits/second). The noise power spectral density, N_0 (Watts/Hz), can be computed as $N_0 = N/B$, where N is the power output (Watts) of a filter of noise bandwidth B (Hz). Thus $E_b/N_0 = \mathrm{SNR} \times B/R$. SNR and E_b/N_0 are both unitless quantities.

5.4.7 SN Curve Measurement Method

The SN curves were measured by transmitting a broad-bandwidth baseband signal over the channel and measuring the received baseband signal and noise powers. A broadband signal was required for several reasons.

First, as shown in Figure 5.7 the IF filter does not have constant gain across its passband[4]. The power at the input of the discriminator will thus depend on the IF spectrum of the FM signal. Therefore, to make SN curve measurements that will apply to OFDM modulation, a signal with same IF spectrum as an OFDM signal is required.

Second, the effect of the transmitter's preemphasis filter and the receiver's deemphasis

[4]Measurement details are given in Appendix D.

subchannels can be scaled. This *software preemphasis* has the effect of increasing the received signal power in the low-frequency subchannels.

The amount of software preemphasis was chosen experimentally. A value of -10 dB per decade was chosen because it tended to produce bit errors that were equally distributed among the subchannels. This software preemphasis was applied prior to the transmitter's $+20$ dB/decade. This same value was used when measuring the SN curves.

5.4.4 Bit Rate

All of the results in this thesis use QPSK (4-QAM) encoding on each subchannel. The phase of each subchannel can be represented as a complex quantity ($\pm 1 \pm j$). An inverse DFT modulates a block of N bits ($N/2$ complex values) into an OFDM signal, producing N real samples. These N samples (or bits) are transmitted over the channel in a block of duration N/f_s where f_s is the sampling rate. The nominal bit rate is thus f_s bps and the nominal bandwidth is $f_s/2$.

However, the channel is bandlimited and not all of the subchannels can be used. This reduces the data rate. The sampling rate (f_s) was 8 kHz and, as explained in Section 5.4.2, only subchannels between 1 and 3 kHz were used. The overall data rate was therefore 4 kbps.

5.4.5 Measuring the IF SNR

The RF (and thus IF) SNR was measured with an HP 8558B spectrum analyzer. The noise power measurement involves measuring the noise bandwidths of the spectrum analyzer and receiver IF filters as well as making corrections for the effect of the noise distribution on the spectrum analyzer's envelope detector and logarithmic display, and the effect of the spectrum analyzer noise floor. The details of the RF SNR measurement are presented in Appendix D.

The allocation of power must also take into account the frequency responses of the transmitter's post-clipping low-pass filter and the receiver's audio circuits. These filters are not flat and therefore some subchannels are attenuated more than others. Those subchannels with higher attenuation have lower SNRs and thus higher BERs. Since the total transmitted baseband power is fixed by clipping and *then* passes through the transmitter's low-pass filter, the use of subchannels with higher attenuation also reduces the total received power.

Brief measurements of E_b/N_0 performance using various frequency ranges showed that a good compromise was to use frequencies between 1 and 3 kHz and to transmit the same power on each subchannel.

Another approach that was not tested but is widely used [37,41,42,61] is to encode more bits on subchannels with higher SNRs.

5.4.3 Software Preemphasis

As described in Chapter 2, the spectrum of the discriminator output noise depends on the IF SNR. Above the FM threshold, the noise has a parabolic spectrum while below the FM threshold the noise spectrum is flat. The relative amounts of noise falling in the different subchannels will therefore depend on the IF SNR.

The use of standard +20 dB per decade preemphasis and −20 dB per decade deemphasis produces a flat noise spectrum when the IF SNR is above threshold. This is appropriate for a non-fading channel that normally operates above threshold. However, the discriminator output noise will have more power at lower frequencies when the IF SNR fades below threshold. The effect of −20 dB per decade deemphasis in this case is to produce a noise spectrum with relatively more noise power at low frequencies. Fading therefore causes low-frequency subchannels to have more noise and these subchannels will have a higher BER.

To equalize the SNRs of the received subchannels, the powers transmitted on the two sub-sections describe how the baseband signal and noise powers were measured and present the SN curves of the experimental channel.

5.4.1 The Modulating Signal Level

As described in Chapter 2, the FM transmitter clips the peaks of the modulating signal to limit the peak deviation and avoid interference with adjacent channels. The effect of clipping on the OFDM signal was included in the SN curves by considering clipping to be equivalent to an additive noise source.

It is difficult to measure the amount of clipping of the modulating waveform because:

- preemphasis occurs before clipping and this makes the amount of clipping depend on the frequency content of the signal, and
- the low-pass filter following the clipper smooths out the signal and distorts the effects of clipping.

However, the additive noise effect due to clipping can be measured (see Section 5.4.7).

Clipping, in the absence of other impairments, sets an upper limit on the baseband SNR, the *limiting baseband SNR*. The modulating signal level should be set so that the limiting baseband SNR does not prevent the system from achieving its BER performance objectives. The modulating signal level was set to give a limiting baseband SNR of 17.5 dB, a level for which the BER was very small[3] ($<10^{-6}$).

5.4.2 The Baseband Frequency Range

The transmitter's clipping circuits limit the total transmitted baseband power and it is necessary to allocate this power among the subchannels. This allocation involves a compromise between concentrating the power in fewer subchannels to reduce the BER or using more subchannels (to increase the bit rate) and increasing the BER.

[3] The effect of clipping varies from block to block since the modulating signal is different for each block.