and delay spread for the next up-link time slot, and determines modulation parameters that can achieve the highest bit rate while satisfying a certain bit error rate (*BER*th:BER threshold). The MS also estimates C/N_0 and the delay spread for the next down-link, determines the optimum modulation parameters, and informs them to the base stations. In the midamble (Mid) of each burst, a Walsh function that represents the modulation level information and the modified PN sequence [2] to estimate delay profile $h(\tau; t)$ are inserted for the up-link and the down-link. Pilot symbols which are used to compensate for fading distortion [3] are inserted in preamble (Pr), midamble and postamble (Po).

Channel estimation: When a directive antenna is employed, the Doppler spread of the received signal becomes narrower than that received by the omnidirectional antenna, although there exists a certain frequency offset [1]. First, the offset frequency is estimated by monitoring the zero-crossing rate and the phase-rotating direction. Then, the f_{off}-cancelled delay profile $\xi(\tau; t)$ is obtained, where the relation between $\xi(\tau; t)$ and $h(\tau; t)$ is given by

$$\xi(\tau;t) = h(\tau;t)e^{-j2\pi f_{off}t} \qquad (1)$$

The f_{off}-cancelled delay profile for the next up-link and down-link slots are predicted by extrapolating the sequence of $\xi(\tau; t)$, and the actual delay profile for the time slot can be obtained by multiplying $\exp(j2\pi f_{off}t)$ to $\xi(\tau; t)$.

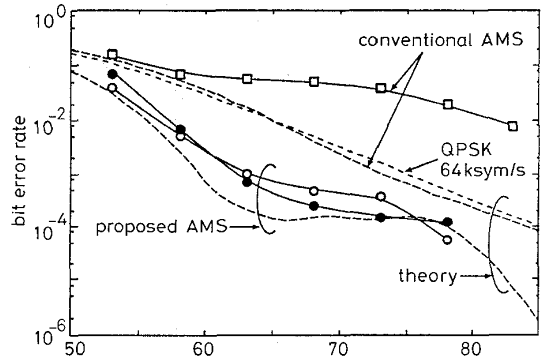

Fig. 3 *BER performance against C/N_0*

$f_dT_F = 0.65$, $B = 4$
● up-link
○ down-link

Simulation: Fig. 3 shows the BER performance of the proposed scheme, where the number of diversity branches (B) is four and (f_dT_F) is 0.65. The modulation scheme is selected from QPSK, 16QAM and 256QAM with their symbol rate of 64 ksymbol/s, and *BER*th = 10^{-3} is selected. The transmitter selection diversity for the up-link and maximal ratio combining diversity for the down-link are executed at the terminal. The employed B-sectored directive antenna has its beamwidth of $2\pi/B$ so as to cover the whole azimuth angle. The directive antenna is assumed to have an ideal pattern $g(\theta)$ given by

$$g(\theta) = \begin{cases} G & -\pi/B \le \theta \le \pi/B \\ 0 & \text{otherwise} \end{cases} \qquad (2)$$

Directional gain G = 6.0dB is assumed. In this paper, C/N_0 is defined as the C/N_0 obtained by using an omnidirectional antenna under the same condition. Fig. 3 also shows the performance of the conventional adaptive modulation scheme and lines show the theoretical values of QPSK, conventional adaptive modulation and the proposed adaptive modulation with selective diversity combining. The BER performance of the conventional adaptive modulation is largely degraded from the theoretical value because the estimation error is so large owing to fast fading variation. Conversely, the BER performance of the proposed AMS is largely improved compared to that for the conventional AMS, and its degradation from the theoretical value is only slight. Fig. 4 shows BER performance against f_dT_F of the conventional and proposed AMS at C/N_0 = 63 and 78dB. In the up-link, the BER performance of the proposed system with $B = 4$ is not degraded at $f_dT_F < 0.8$, while that for the conventional AMS is degraded at $f_dT_F > 0.1$. Although the performance of the down-link is more sensitive to f_dT_F, the degradation is tolerable in comparison with the up-link at $f_dT_F < 0.8$. If a greater number of branches is used, for example $B = 8$, it is possible to keep up with faster fading variation.

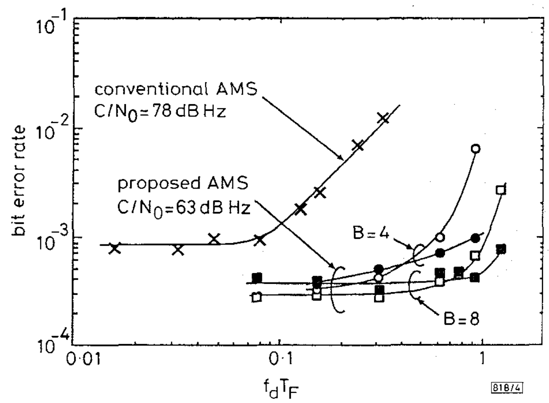

Fig. 4 *BER performance against f_dT_F*

●, ■ up-link
○, □ down-link

Conclusion: This Letter proposes AMS using directive antenna diversity to improve AMS performance in high mobility land mobile communication environments, and evaluates its performance in Rayleigh fading environments by computer simulation. Computer simulation confirms that the proposed AMS can achieve high quality and high bit rate transmission even in fast fading environments.

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Easing radiated EMC problems with spread-spectrum modulation of computer clock signals

D.A. Stone and B. Chambers

Indexing terms: Electromagnetic compatibility, Spread spectrum communication

The effect on radiated interference of the application of frequency-hopping spread-spectrum modulation techniques to the clock frequency of a microcomputer has been investigated. Preliminary results show that this technique is capable of at least an 8dBµV/m reduction in radiated quasipeak interference levels in the near field when using standard EMC measurement test bandwidths.

Introduction: Recent introduction of the European Community (EC) electromagnetic compatibility (EMC) legislation [1], and its imminent enforcement in January 1996, has prompted extensive research into both conducted and radiated electromagnetic interference (EMI). Computer systems, having higher and higher clock frequencies, form a major source of radiated interference and require shielding around some or all of the circuits to ensure that the spectral power density of the clock signal and its harmonics is below the mandatory levels. As virtually all operations in a computer occur in synchronism with each other, all switching events fall at the same instant, thus adding to the measured power density at harmonics of the master clock frequency. To ease the problem of passing the EMC standard tests, this Letter shows that it is possible to reduce the energy measured at harmonics of the clock frequency by spreading it over the available spectrum using digital frequency-hopping spread-spectrum (FHSS) techniques applied to the system master clock. An FHSS technique has been shown to be effective in the reduction of conducted EMI in switched mode power supplies [2, 3], and may be similarly applied to a computer system by changing the clock frequency at regular intervals to ease radiated emissions. An initial investigation into frequency modulation of a computer clock signal with a periodic waveshape has been reported [4], which achieves a form of spectral spreading; the work presented here, however, gives a true implementation of FHSS modulation of a computer clock signal based on a maximal length pseudorandom sequence.

The FHSS system instigates a change in the clock frequency at the end of an integer number of clock cycles, giving a dwell time at any given clock frequency of $1/nf$, where f is the clock frequency for the particular number of cycles, and n is the number of cycles at that frequency. The extent to which energy is spread around the spectrum is given by the dispersion factor $\beta = T\Delta f$, where Δf is the peak frequency deviation of the clock frequency, and T is the period of repetition of the modulation sequence. The modulating sequence used is a maximal length pseudorandom sequence generated from the equation $X_{i+1} = (aX_i + b) \bmod c$, where $c = 2^p$, $b = 2q+1$, $a = 4b+1$, and p, q are any integers [5].

EMI measurements: In practice, the perceived effect of spectral spreading is determined by the resolution bandwidth (RBW) of the system making the measurements. Under EC EMC test conditions, the RBW specified for radiated EMC measurements above 30MHz is 120kHz; therefore, any change to the clock frequency which causes it to remain within the RBW of the measurement system will not register as a change in the measured signal strength. The EC EMC emission limits are specified for measurements taken with a quasipeak detector whose charge and discharge time constants are 1 and 550ms, respectively, at this RBW. The effect of both the RBW and the time constants have therefore to be taken into consideration when choosing a peak frequency deviation (Δf) and length (n) for the modulating sequence.

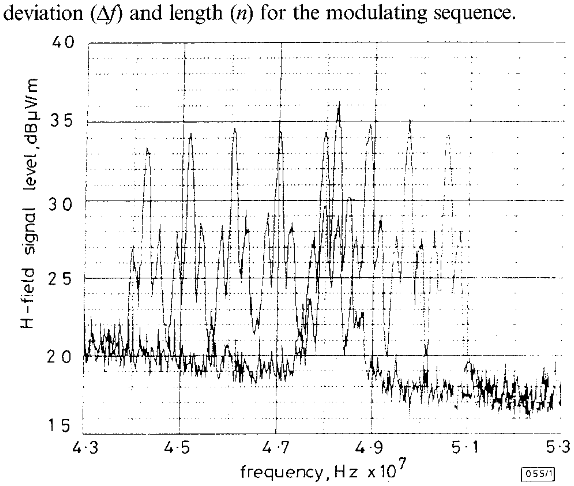

Fig. 1 *3rd harmonic of 16MHz clock*

Constant frequency operation, and spread with a maximal length pseudorandom sequence
RBW = 120kHz, n = 8

Practical measurements: Practical results presented in this Letter were taken with a Farnell 5-1000A 150kHz to 1GHz spectrum analyser and an H-field probe from the Farnell 7405 near-field probe set. An FHSS clock module was built which allows the clock to be modulated with various modulation sequence types having sequences lengths (n) of up to 256. The peak frequency deviation (Δf) from the nominal 16 MHz clock frequency is variable up to ±4MHz. The clock module was used with a commercially available process-controller module based around the 87C196KC16 16MHz microprocessor and to demonstrate the effectiveness of the technique, near-field emissions were measured in the vicinity of the microprocessor chip.

Fig. 1 shows some preliminary results of the effect of using a Δf of 1MHz (frequency range from 15MHz to 17MHz) and a sequence length n = 8, on the third harmonic of a nominally 16MHz clock. The spreading of the energy to eight distinct frequencies can clearly be seen, as can the reduction in peak values at the individual frequencies. The quasipeak value for the constant frequency clock was 35dBμV/m with a peak reading of 36dBμV/m; however, the spread-spectrum clock showed a quasipeak reading of 27dBμV/m against a peak reading of 34dBμV/m. This shows a reduction in the quasipeak levels of 8dBμV/m at the third harmonic of the clock frequency. Fig. 2 shows the spreading effect with a sequence length n = 256, the longer sequence giving a more continuous spread across the frequency range. The quasipeak levels in this case were 26dBμV/m with a peak level of 34dBμV/m; showing very little difference between the two sequence lengths at this harmonic of the clock. Fig. 3 shows the effect of spectral spreading on the 11th harmonic of the clock, under the same conditions as Fig. 2. Here the measured quasipeak levels were 35dBμV/m with a 36dBμV/m peak level for the constant frequency operation, and 23dBμV/m with a peak of 34dBμV/m for the FHSS operation. This shows a 12dBμV/m reduction in the quasipeak level at this clock harmonic.

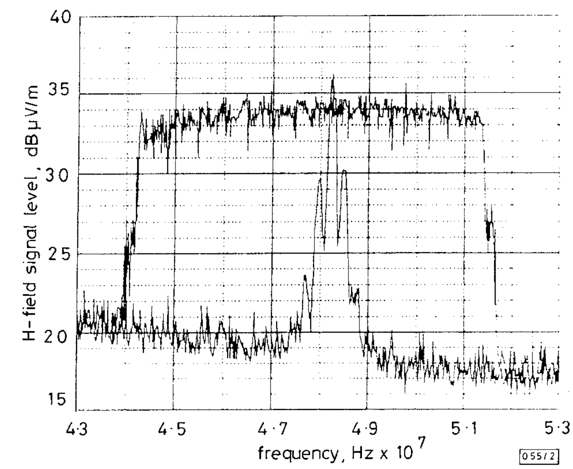

Fig. 2 *3rd harmonic of 16MHz clock*

Constant frequency operation, and spread with a maximal length pseudorandom sequence
RBW = 120kHz, n = 256

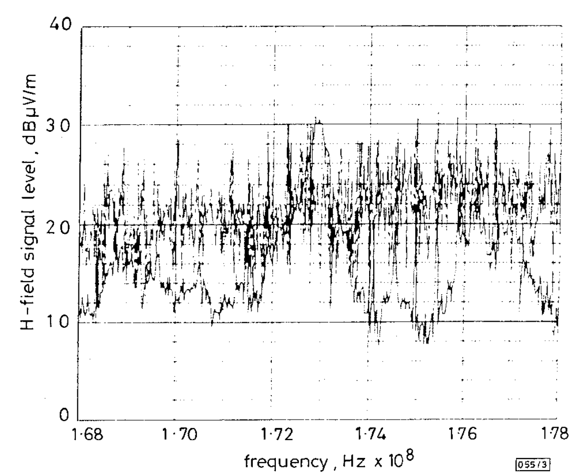

Fig. 3 *11th harmonic of 16MHz clock*

Constant frequency operation, and spread with a maximal length pseudorandom sequence
RBW = 120kHz, n = 256

Conclusion: We have shown the effectiveness of using an FHSS clock generator in a microcomputer system to reduce the quasipeak levels of electromagnetic emissions, and thus ease the problems of EMC compliance of digital microcomputer systems. This technique is capable of at least 8dBμV/m reduction in the quasipeak levels at low-order harmonics of the clock. Higher-order harmonics show an increased reduction in the quasipeak levels, consistent with the increase in the spreading effect with the harmonic number, whilst the RBW of the detector remains the same. Work is now under way to optimise the Δf, n and sequence type for maximum effect. The prototype microcomputer system functioned normally at all times when fed with the FHSS clock, being required to execute a self-test program whenever the clock modulation strategy was changed.

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Interference suppression for CDMA overlay of a narrowband waveform

J. Wang and L.B. Milstein

Indexing terms: Code division multiple access, Interference suppression

The performance of CDMA networks which overlay a narrowband signal to increase spectral efficiency is presented. Such systems employ interference suppression filters in the CDMA receivers to minimise the interference caused by the narrowband waveform.

Introduction: In this Letter, the use of interference rejection techniques to code division multiple access (CDMA) systems which are designed to share a common spectrum with a narrowband waveform, is presented. It is assumed that the narrowband signal is being overlaid by a CDMA network to increase the overall spectral efficiency of the band, and that, therefore, does not represent intentional jamming. Rather, the interference that they impose on the CDMA waveforms is the unavoidable result of attempting this type of spectral sharing, and it is the purpose of the interference suppression technique to minimise the resulting degradation to system performance.

Mathematical model: The transmitted signal of the kth user is given by

$$S_k(t) = \sqrt{2P} b_k(t) a_k(t) \cos(2\pi f_0 t + \theta_k) \qquad (1)$$

where P and f_0 denote the transmitted power and the carrier frequency, respectively; θ_k is a random phase; $b_k(t)$ is a random binary sequence representing the data; $a_k(t)$ is the spreading sequence, also modelled as a random binary sequence. Each data bit has a duration of T_b seconds, each chip of the spreading sequence has duration T_c seconds, and the processing gain is defined as $N = T_b/T_c$. The channel between the kth user and the receiver of interest is modelled as a flat Rayleigh fading channel, and is characterised by three random variables, β_k, τ_k and θ_k, which are, respectively, defined as the gain, delay and phase of the kth signal. The interference in the channel is assumed to be a non-fading narrowband BPSK signal and is given by

$$J(t) = \sqrt{2J} d(t) \cos[2\pi(f_0 + \Delta)t + \eta] \qquad (2)$$

where Δ stands for the offset of the interference carrier frequency from the carrier frequency of the CDMA signals, and J and η denote the received interference power and phase, respectively. The binary information sequence $d(t)$ has bit rate $1/T_j$. Therefore, the interference bandwidth is $B_j \simeq 2T_j^{-1}$, and we assume that B_j is much less than B_s ($B_s \simeq 2T_c^{-1}$), the spread bandwidth of the CDMA waveform. The ratio of the interference bandwidth to the spread spectrum bandwidth p and the ratio of the offset of the interference carrier frequency to half of the spread spectrum bandwidth q are defined as

$$p = B_j/B_s = T_c/T_j \qquad (3)$$

and

$$q = \Delta/(B_s/2) = \Delta T_c \qquad (4)$$

respectively. Neglecting the channel noise, the received signal $r(t)$ can be represented as

$$r(t) = \sqrt{2P} \sum_{k=1}^{K} \beta_k b_k(t-\tau_k) a_k(t-\tau_k) \cos(2\pi f_0 t + \phi_k) + J(t) \qquad (5)$$

where K denotes the number of active users and $\phi_k = \theta_k - \mu_k - 2\pi f_0 \tau_k$.

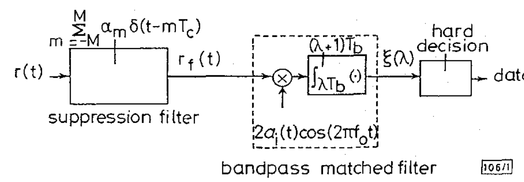

Fig. 1 *Receiver model of CDMA overlay system*

The receiver is shown in Fig. 1; it contains a narrowband interference suppression filter with two sides, a bandpass matched filter and a hard decision device. The resulting suppression filter output is given by

$$r_f(t) = \sum_{m=-M}^{M} \alpha_m r(t - mT_c) \qquad (6)$$

Assuming that user i is the reference user ($\phi_i = 0$ and $\tau_i = 0$) and that $f_0 T_c$ is an integer, the decision variable $\xi(\lambda)$, where λ denotes the λth data bit of the reference user, can be written as

$$\xi(\lambda) = \int_{\lambda T_b}^{(\lambda+1)T_b} r_f(t) 2a_i(t) \cos(2\pi f_0 t) dt$$

$$= D(\lambda) + J(\lambda) + \sum_{k=1,(k\neq i)}^{K} I_{k,i} \qquad (7)$$

where $D(\lambda)$ is the desired signal term of the reference user at the zero-th tap of the suppression filter. Given β_i, the conditional useful signal power is equal to $2P\beta_i^2 T_b^2$. $J(\lambda)$ is caused by the interference, and its variance is equal to

$$T_b^2 \frac{J}{N} \sum_{m_1=-M}^{M} \sum_{m_2=-M}^{M} \alpha_{m_1} \alpha_{m_2} \sigma_j^2(m_1, m_2)$$

where

$$\sigma_j^2(m_1, m_2) = \int_{-1}^{1} (1-|x|) \left(1 - \frac{|x|}{N}\right) \mathrm{sign}(1 - |x - m_1 + m_2|)$$
$$\times \cos[2\pi q(x - m_1 + m_2)] dx \qquad (8)$$