Spread Spectrum Communications Using Chirp Signals

A. Springer[1], W. Gugler[1], M. Huemer[1], L. Reindl[2], C.C.W. Ruppel[3], R. Weigel[1]
[1]University of Linz, Institute for Communications and Information Engineering, Austria
E-mail: springer@mechatronik.uni-linz.ac.at,
[2]University of Clausthal, Institute for Electrical Information Technology, Germany.
[3]Siemens AG, Corporate Technology, ZT KM 1, Munich, Germany.

Abstract— **We report on the use of broadband chirp signals for spread spectrum systems in indoor applications. The presented system concepts make use of chirp transmission and pulse compression. Different modulation schemes for chirp signals resulting in different system performance and complexity are compared in terms of bit error rate for the AWGN channel and for frequency selective indoor radio channels. We present simulations and measurement results from demonstrator systems which use surface acoustic wave (SAW) devices for the generation and matched filtering of the chirp signals. RF and IF frequency and transmission bandwidth of the presented systems are 2.4 GHz, 348.8 MHz, and 80 MHz, respectively. Due to the processing gain of 16 dB - made possible by the use of SAW devices - and the large transmission bandwidth the system is insensitive against frequency selective fading, CW interference and noise.**

I. Introduction

Indoor wireless communications has gained increasing attention over some time and its market share is expecting to grow rapidly in the next years due to advantages over cable networks, such as mobility of users, elimination of cabling and flexibility. Typical applications are cordless phone systems and wireless local area networks (WLAN's) for home and office applications and flexible and mobile data transmission links between sensors, actuators, robots, and controller units in industrial environments. Due to the hostile electromagnetic environment, which includes severe electromagnetic emissions from other devices as well as heavy distortions due to multipath propagation [1] the robustness of the communication link is an extremely important feature of a wireless communication system.

The spread spectrum technology is especially well suited to provide such a robust data transmission even in very noisy radio environments [2]. The critical operations in spread spectrum systems are the spreading and despreading functions in the transmitter and receiver. In common system concepts (direct-sequence and frequency hopping) the synchronization of the despreading code is a difficult task which needs high computational effort. With the well known FM chirp signals and the associated technique of pulse compression with its high processing gain, which is widely used in radar systems [3], another kind of spread spectrum technique can be realized [5],[6],[7]. In this system concept, the spreading is used solely for combating multipath distortions, whereas code-division multiple acces (CDMA) can only be realized if additional coding is introduced [4].

The implementation of the spreading and despreading with chirp signals is easily accomplished by using surface acoustic wave (SAW) chirped delay lines [5]. These devices can be realized at small size and low cost and due to the analog correlation process the complex synchronization circuits can be economized.

After an introduction into the theory of chirp signals we describe different incoherent and coherent modulation schemes for chirp spread spectrum systems. Simulation results and first measurements obtained with a hardware demonstrator are given.

II. Chirp Theory

A chirp waveform [3] can be written as

$$s(t) = a(t)\cos\left[\Theta(t)\right] \qquad (1)$$

where $\Theta(t)$ is the phase, and $a(t)$ is the envelope of the chirp signal which is zero outside a time interval of length T. The instantaneous frequency is defined as

$$f_M(t) = \frac{1}{2\pi}\frac{d\Theta}{dt}. \qquad (2)$$

The chirp rate is defined by

$$\mu(t) = \frac{df_M}{dt} = \frac{1}{2\pi}\frac{d^2\Theta}{dt^2} \qquad (3)$$

and represents the rate of change of the instantanous frequency. We call waveforms with $\mu(t) > 0$ up-chirps an those with $\mu(t) < 0$ down-chirps. For a linear chirp $\mu(t)$ is constant, and hence $f_M(t)$ is a linear function of t, and $\Theta(t)$ is a quadratic function. If we take the waveform to be centered at $t = 0$ it can be written as

$$s(t) = a(t)\cos\left[2\pi f_0 t + \pi\mu t^2 + \varphi_0\right] \qquad (4)$$

where f_c is the center frequency and $a(t) = 0$ for $|t| > T/2$. It is convenient to define the bandwidth B as the range of the instantanous frequency, so that

$$B = |\mu| T. \qquad (5)$$

The impulse response of a matched filter for a linear chirp signal is again a linear chirp signal but with a chirp rate of opposite sign. If a chirp waveform is fed into its matched filter the output signal typically has a narrow IF peak at the chirp center frequency. If we consider chirp waveforms with flat time domain envelopes and take the matched filter to be centered at $t = 0$ an analytical expression for the output waveform $g(t)$ of the matched filter can be given. We have

$$g(t) = h(t) * s(t) = \varphi_{ss}(t) \qquad (6)$$

0-7803-6323-X/00/$10.00 (c) 2000 IEEE

where $\varphi_{ss}(t)$ is the autocorrelation function of $s(t)$. It can be shown [9] that $\varphi_{ss}(t)$ is given by

$$\varphi_{ss}(t) = \sqrt{BT}\frac{\sin\left\{\pi Bt\left(1-\frac{|t|}{T}\right)\right\}}{\pi Bt}\cos(2\pi f_0 t) \qquad (7)$$

for $-T < t < T$. The envelope has its maximum at $t = 0$, and its first zeros at $t \approx \pm 1/B$. It is convenient to specify the pulse width as $1/B$. The ratio of the input and output pulse widths is therefore given by the time-bandwidth product TB which is known as compression ratio or processing gain. Another important parameter is the sidelobe rejection, which is about 13 dB for chirp signals with rectangular time domain envelope $a(t)$. A common method of reducing the sidelobes is to apply amplitude weigthing to the chirp signals.

III. Binary Orthogonal Keying Using Up- and Down-Chirp Signals

As up- and down-chirp signals are almost orthogonal they can be used for a binary orthogonal keying (BOK) modulation scheme as is shown in Fig. 1 a) in the time frequency plane. If a '1'-bit is sent, an IF pulse at the chirp center frequency stimulates an up-chirp filter, if a '0'-bit is sent, a down-chirp filter is stimulated. To increase the data rate beyond the limit imposed by the chirp duration the chirp signals have to overlap in time. In the receiver the signal is fed into the SAW compressor filters. The up-chirp filter is matched to the '0'-bit signal, the down-chirp filter to the '1'-bit signal. Due to the overlapping and the fact that up- and down-chirps are not exactly orthogonal disturbing crosscorrelation functions in both filter outputs occure. Together with the time spreading of the correlation peaks caused by the multipath channel this limits the achievable data rate. For the same reasons the theoretical BER of a system using two orthogonal signals

$$P_e = \frac{1}{2}\mathrm{erfc}\left(\sqrt{\frac{E}{4N_0}}\right). \qquad (8)$$

can not be reached exactly (Fig. 2).

IV. Binary Orthogonal Keying Using Pulse Position Modulation

When applying pulse position modulation (PPM) to chirp spread spectrum transmission only one chirp waveform is needed for transmission and therefore disturbances due to crosscorrelations do not occur. Furthermore we economize one SAW-filer in the transmitter and one in the receiver. The schematic of this system is depicted in Fig. 1 b). We do not send the data in equidistant time steps. If we send a '1'-bit the chirp signal is sent at a fixed time Δt before and if we send a '0'-bit the signal is sent at the same fixed time after the reference clock. An increase in the data-rate by overlapping the chirp signals is possible without any disturbances. Therefor the limiting factors for the systems data rate are the fixed time shift Δt and multipath fading. Since there are no crosscorrelations the exact BER given in eq. (7) is reached in the AWGN case.

V. $\pi/4$ DQPSK Modulation of Chirp Signals

Traditionally, the phase shift keying (PSK) modulation modifies the phase of the carrier signal. For the coherent demodulation of any M-ary PSK signal the absolute phase of the signal has to be known. To overcome the problems associated with the necessary phase synchronisation the technique of phase comparison detection (DQPSK) is used. This also solves the problems of carrier recovery. In the presented system the well known $\pi/4$ DQPSK modulation scheme is used. Different from traditionally $\pi/4$ DQPSK systems we do not modulate the carrier. Instead, the IF-pulse which is stimulating the SAW-chirp filter is phase modulated.

The schematic of the chirp $\pi/4$ DQPSK system is depicted in Fig. 3. The input data are split in an inphase- and quadraturephase component (I and Q component) and fed into the modulator. Another input of the modulator is a short IF pulse at the chirp center frequency. The output of the modulator, which is the $\pi/4$ DQPSK modulated IF pulse, stimulates the transmit chirp (expander) filter resulting in the desired phase modulated chirp signal. The received signal which is disturbed by the frequency selective radio channel and additive white noise is fed into the SAW compressor filter followed by the $\pi/4$ DQPSK demodulator which is shown in Fig. 4.

If we assume a signal

$$x_1(t) = A\cos(\omega t - \varphi_1 + \varphi_{\mathrm{ch}})$$

at the input of the demodulator, the signal after the delay T_{Sym} and a phase shift of $\Phi = \omega T_{\mathrm{Sym}}$ becomes

$$x_2(t) = A\cos(\omega(t + T_{\mathrm{Sym}}) - \varphi_2 + \Phi + \varphi_{\mathrm{ch}}),$$

where φ_1, φ_2, φ_{ch} and T_{Sym} are the phase of the present symbol, the phase of the delayed symbol, the phase shift due to the channel, and the symbol duration, respectively. We also assume that the phase of the channel φ_{ch} is constant during a symbol period since the indoor radio channel does not change significantly during this time intervall (200 nsec at a data rate of 10 Mbps). At the output of the demodulator the I- and Q-data after low pass filtering and sampling are

$$\begin{aligned} y_I &= \frac{1}{2}A^2\cos(\varphi_1 - \varphi_2) \\ y_Q &= \frac{1}{2}A^2\sin(\varphi_1 - \varphi_2) \end{aligned}$$

which represents the known constellation in the I-Q diagram of any $\pi/4$ DQPSK system. Due to the differential detection the phase shift due to the channel is canceled. In case of the $\pi/4$ DQPSK chirp system the data rate is only limited by the time duration of the autocorrelation pulse and the time spreading due to multipath fading.

VI. System Simulation

System simulations have been carried out to evaluate the performance of the proposed modulation concepts. ADS from HP-EEsof has been used for the simulations and all computations have been carried out in complex baseband. Measured data for the transfer function of the SAW chirp filter have been used in the simulations. For the hardware demonstrator of the PPM system up- and down-chirp filters have been designed and fabricated from $LiTaO_3$-X112rotY substrate using standard optical lithography technique. An optimized weighting of the magnitude of the filter transfer-function was employed to obtain an improved sidelobe rejection while keeping the compressed pulse narrow. Fig. 5 depicts the transfer function and the group delay of an up-chirp filter. The center frequency, bandwidth, and time duration of the chirp signals used in the system simulations as well as in the hardware demonstrator are 348.8 MHz, 80 MHz, and 500 nsec, respectively. The chirp rate of the filters is about ± 40 MHz/μs which results in a dispersion time of about $0.5\mu s$. This corresponds to a time-bandwidth product of 16 dB.

As is shown in [8] the exact average bit error rate (BER) of a $\pi/4$ DQPSK system under AWGN conditions is given by

$$P_e = \frac{1}{2}(P_{eI} + P_{eQ}) = \frac{1}{2}\Bigg[1 - Q\left(\sqrt{\tfrac{E_b}{N_0}(2+\sqrt{2})}, \sqrt{\tfrac{E_b}{N_0}(2-\sqrt{2})}\right) + Q\left(\sqrt{\tfrac{E_b}{N_0}(2-\sqrt{2})}, \sqrt{\tfrac{E_b}{N_0}(2+\sqrt{2})}\right)\Bigg]$$

where Q is the Marcum Q-function. If we assume a data rate of 10 Mbps in our system we get an autocorrelation pulse every 200 ns. The duration of the autocorrelation pulse is about 20 ns. In Fig. 6 the theoretical BER of the traditional $\pi/4$ DQPSK system and the BER of our system with a data rate of 10 Mbps and ideal data clock recovery is plotted. By increasing the data rate up to 67 Mbps an autocorrelation pulse occurs every 30 ns. As can be seen the simulated BER coincides well with the theoretical BER of a $\pi/4$ DQPSK system. If data clock recovery by means of a PLL is applied in the simulation, the results from Fig. 6 show a loss of about 1 dB.

In a further step we considered an indoor radio channel with non line-of-sight (NLOS) condition and intersymbol interference (ISI) according to the model used in the IEEE 802.11 wireless LAN standardization group. For improved detection both outputs of the demodulator are integrated over the duration of one symbol and sampled according to the data clock recovery. Thus we were able to sum up the energy of all radio channel paths during one symbol periode. The BER including the indoor radio channel, the data clock recovery and the integration can be seen in Fig. 6.

VII. Measurements

Demonstrators for the BOK and the PPM chirp modulation schemes were built operating in the ISM band at 2.45 GHz. Regarding measurement results of the BOK demonstrator we refer to [5]. Fig. 7 shows measurements of the PPM chirp system at a data rate of 1 Mbps. The plot above shows the output of the chirp compressor filter without any multipath fading. The distance between transmitter and receiver in the laboratory was 2 m and the line-of-sight (LOS) component predominated. The different time intervalls between the autocorrelation peaks according to the pulse position modulation can clearly be seen. In the lower plot the output of the chirp compressor filter at a distance of about 15 meters with two reinforced concrete walls in between transmitter and receiver is depicted. At least two propagation paths can be distinguished. The limiting factor for this kind of modulation is the time shift according to the PPM and time spreading due to multipath fading.

VIII. Conclusion

We presented chirp spread spectrum systems for robust wireless communications in indoor environments. The systems take advance of the low cost, small size and power efficient SAW technology. SAW devices are used for chirp generation in the transmitter and for pulse compression in the receiver. Due to the analog correlation process the usually complex synchronization task can be greatly simplified. Different modulation schemes were discussed and simulation results as well as measurements of the data throughput were presented. Simulations and measurements indicate that the proposed systems overcome the disturbances due to frequency selective fading and CW interferers by the transmission of ultra broadband signals. Therefore the presented chirp spread spectrum technique is a well suited solution for very robust wireless communication systems in highly distorted environments.

References

[1] H. Hashemi, ***The Indoor Radio Propagation Channel,*** Proceedings of the IEEE, Vol.81, No.7, (1993).

[2] R. P. Dixon, ***Spread Spectrum Systems with Commercial Applications,*** New York: Wiley, (1994).

[3] D. P. Morgan, ***Surface Wave Devices for Signal Processing,*** Elsevier: Amsterdam, (1985).

[4] M. Kowatsch, J. T. Lafferl, ***A Spread Spectrum Concept Combining Chirp Modulation and Pseudonoise Coding,*** IEEE Transactions on Communications, Vol.29, No.6, (1981).

[5] A. Springer, M. Huemer, L. Reindl, C.C.W. Ruppel, A. Pohl, F. Seifert, W. Gugler, R. Weigel, ***A Robust Ultra Broadband Wireless Communication System Using SAW Chirped Delay Lines,*** IEEE Transactions on Microwave Theorie and Techniques, Vol.46, No.12, (1998).

[6] M. Huemer, A. Pohl, W. Gugler, A. Springer, R. Weigel, F. Seifert, ***Design and Performance of a SAW Based Chirp Spread Spectrum System,*** Proc. 1998 IEEE MTT-S International Microwave Symposium, (1998).

[7] W. Gugler, A. Springer, R. Weigel, H.P. Kpfer, ***Simulation of a SAW-Based WLAN Using Chirp-Pi/4 DQPSK Modulation,*** Proceedings 1998 IEEE Ultrasonics Symposium, (1999).

[8] L. Miller, J. Lee, ***BER Expressions for Differentially Detected π/4 DQPSK Modulation,*** IEEE Transactions on Communications, Vol.46, (1998).

[9] Ch. E. Cook, M. Bernfeld, ***Radar Signals - an introduction to theory and application,*** Academic Press, New York, (1967)

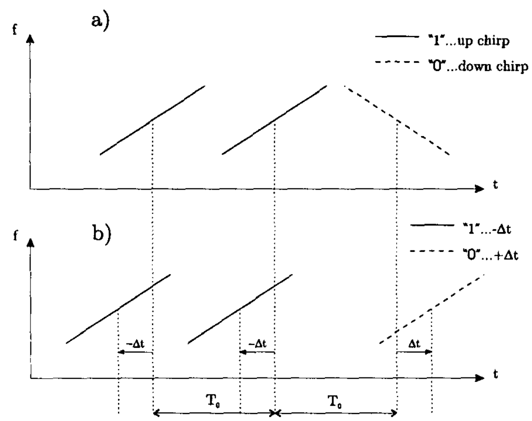

Figure 1. Binary orthogonal keying schemes.

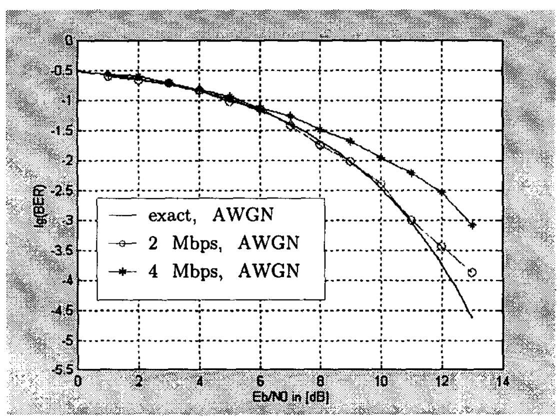

Figure 2. BER of a BOK chirp system using tw o orthogonal siganls.

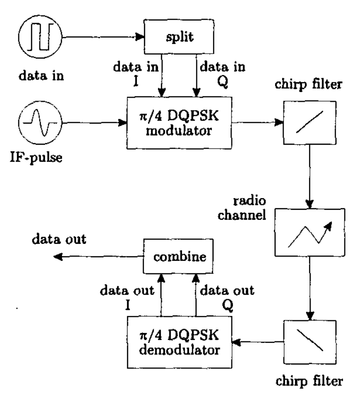

Figure 3. Schematic of the chirp $\pi/4$ DQPSK system.

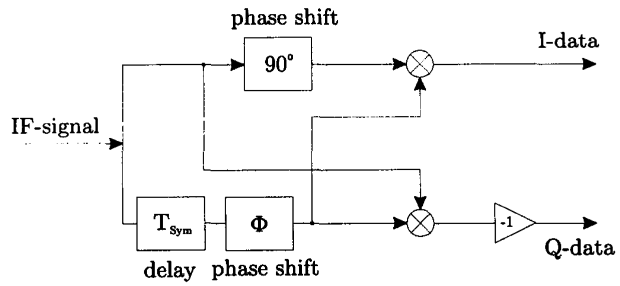

Figure 4. Chirp $\pi/4$ DQPSK demodulator.

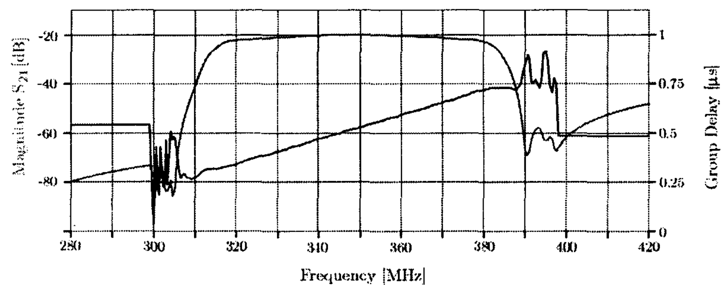

Figure 5. Measured transfer function and group delay of an optimized weighted chirp filter.

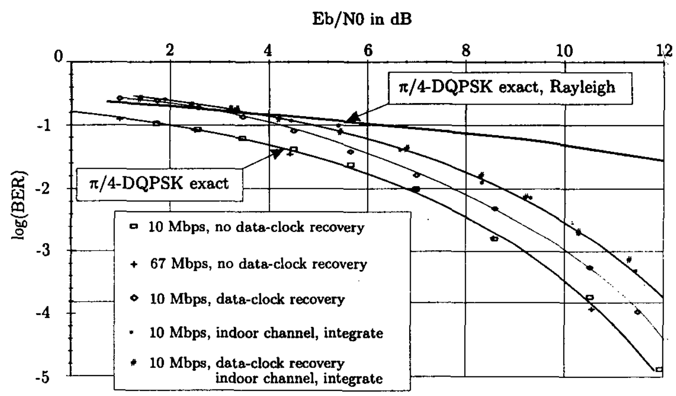

Figure 6. BER of the chirp $\pi/4$ - DQPSK system.

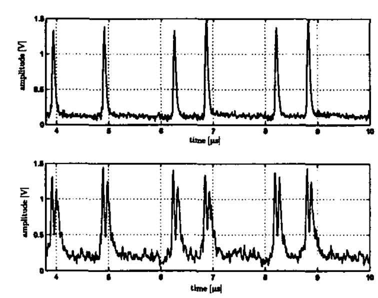

Figure 7. Measurements of a chirp PPM system.