Title	Design and Implementation of an Indoor Ultrasonic Communication System
Authors(s)	Ballal, Tarig, Saad, Mohamed M., Bleakley, Chris J.
Publication date	2011-06-22
Publication information	Ballal, Tarig, Mohamed M. Saad, and Chris J. Bleakley. "Design and Implementation of an Indoor Ultrasonic Communication System." The Institution of Engineering and Technology, 2011.
Conference details	22nd IET Irish Signals and Systems Conference (ISSC), Dublin, Ireland, June 23-24, 2011
Publisher	The Institution of Engineering and Technology
Item record/more information	http://hdl.handle.net/10197/3858

DESIGN AND IMPLEMENTATION OF AN INDOOR ULTRASONIC COMMUNICATION SYSTEM

Tarig Ballal, Mohamed M. Saad, and C. J. Bleakley

UCD Complex & Adaptive Systems Laboratory
UCD School of Computer Science & Informatics
University College Dublin, Belfield, Dublin 4, Ireland
Tel: + (353) 1 716 2915, Fax: + (353) 1 269 7262, E-mail: {tarig.ballal, chris.bleakley}@ucd.ie
web: http://casl.ucd.ie/, http://www.csi.ucd.ie/

ABSTRACT

In this paper, an indoor communication system using ultrasonic signals is described. The system uses *differential binary phase-shift keying* (DBPSK) for transmitting binary data as a part of an indoor positioning system. A synchronization and decoding approach is proposed that exploits the correlation properties of the DBPSK waveforms. All of the detection and decoding/demodulation processes are performed digitally—no analogue circuits are involved. Experiments that were carried out using off-the-shelf components confirmed the feasibility of the proposed system.

***Index Terms*—** Ultrasonic based communication, DBPSK, correlation, waveform, modulation

1. INTRODUCTION

Ultrasonic based positioning is attractive for indoor location systems. Owing to the low propagation speed of sound, signal parameters such as *time-of-flight* can be estimated more accurately than in *radio frequency* (RF) based systems [1, 2]. Applications of ultrasonic systems include hospitals and ubiquitous computing [3, 4].

Traditionally, ultrasonic positioning systems (e.g., [1, 2, 5]) used RF signals to synchronize system units to allow for measurement of time-of-flight. In these, systems, the RF transceivers can be used for data communications between units.

In [3] and [4], indoor data communication using airborne ultrasound is discussed. The papers considered different factors that affect the design and operation of the systems. However, no details on the actual operation of the communication system were given. The papers suggest using *M-ary frequency-shift keying* (MFSK). The advantage of such modulation scheme is that it requires simple non-coherent detection and does not have the stringent synchronization requirements of coherent detection [6].

This paper proposes a system for ultrasonic data communication utilizing *differential binary phase-shift keying* (DBPSK) over 4 frequency channels. The transmission alternates periodically between channels in a way similar to a *frequency-hopping spread-spectrum* (FHSS) modulation schemes. This approach simplifies the system operation, as will be revealed subsequently.

In section 2, general issues related to the design of the signals are discussed and a design is identified. Section 3 discusses the DBPSK waveform properties that are exploited by the system. Section 4 describes the operation of the system in terms of the detection and the decoding of the signals. In Section 5 the experimental setup is described and test results are reported. Section 6 gives the conclusions of this paper.

2. SIGNAL DESIGN

In a communication system, orthogonal frequency sets should be separated by at least $1/T_b$, where T_b is the symbol duration (bit duration in the case of binary transmission) [7]. The acoustic channel is known for its large Doppler shifts compared to the RF channel [3]. Therefore, in the case of an ultrasonic based communication system, Doppler effect has to be accounted for in the signal design. Here, we propose increasing the frequency separation to

$$\Delta f = \frac{1}{T_b} + 2f_g \tag{1}$$

where f_g is a *guard frequency* meant to accommodate Doppler shifts between $-f_g$ Hz and $+f_g$ Hz. The introduction of such redundancy ensures that the frequency contents of each channel will always remain within a *distinct* frequency band. Fig. 1 depicts an orthogonal frequency set based on Eq. (1).

The number of orthogonal frequencies available for transmission can be calculated from the total available bandwidth,

This work is supported by the Higher Education Authority (HEA) of Ireland under the Programme for Research in Third Level Institutions, cycle 4 (PRTLI4)—the NEMBES project.

B, as

$$K = \frac{B}{\Delta f} = \frac{B}{\frac{1}{T_b} + 2f_g}. \quad (2)$$

Doppler shift at a frequency f_0 under motion with velocity v (normally + or -) is given by $f_d = f_0 v/c$, where is the c is the speed of wave propagation. For ultrasound, typically, the value $c = 343$ m/s is used corresponding to an indoor temperature of 20º Celsius.

Typical off-the-shelf ultrasonic transducers have bandwidths of $B = 4$ kHz around the frequency 40 kHz. Indoor motion velocities are normally 1–2 m/s. Based on these facts, the design parameters are selected as follows. The total bandwidth is divided into 4 channels ($K = 4$), each with a bandwidth (Δf) of 1 kHz. The guard frequency is set to $f_g = 167$ Hz, which can accommodate motions with speeds up to 1.43 m/s. These choices result in a symbol duration, T_b, of about 1.5 ms, which correspond to a symbol rate of approximately 666 symbol/s. If 4-FSK modulation is to be used, this gives a data rate of 1.332 k bit/s.

In the application of interest, short codewords are transmitted by each transmitter device in the environment at random times. Long intervals of silence occur between the codeword bursts from a device to allow for reverberation (multipath) to attenuate and to avoid collision between bursts from different transmitters. The receiver has to listen to the channels continuously to detect the arrival of the data and synchronize to each symbol. With MFSK this can be difficult since the channel used for transmitting a certain datum is determined by that datum. This leads to uncertainty in the channels to be received, and hence makes synchronization harder.

Here we propose transmitting binary data using DBPSK over the four channels in sequence (f_1, f_2, f_3, f_4). The first bit is transmitted over the first channel, the second bit over the second, etc. The pattern can be repeated until all of the data is transmitted. This resembles a FHSS scheme and can provide some resistance against multipath effects [6, 8]. However, the major advantage is that the received hopping pattern is known, which simplifies receiver operation. Also, the (a priori known) orthogonality between consecutive channels and other properties of DBPSK waveforms can be exploited for synchronization, as will be demonstrated subsequently. This is achieved at the cost of a reduction in data rate; due to the differential nature of DBPSK two symbols are required to transmit every single bit, and the data rate drops to 333 bit/s.

3. DBPSK WAVEFORMS

DBPSK represents one binary digit by a phase-shift of 0 radian and the other by a phase-shift of π radian. A startup symbol relative to which data is encoded is always required [6]. This means that to transmit a single binary digit over a certain frequency, two symbols are required. This leads to the aforementioned data rate reduction. Here, we conceive the two

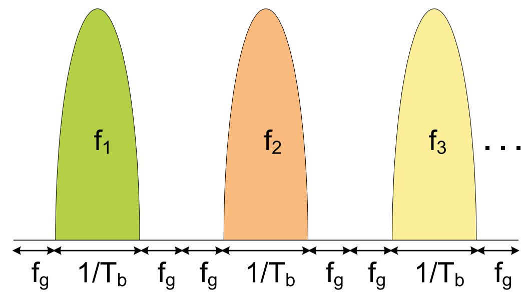


Fig. 1: Orthogonal frequency set.

symbols pertaining to a binary digit transmitted over a certain frequency, as a single waveform. Therefore, the waveforms for binary 0 and binary 1 at frequency $f_k, k = 1, ..., 4$ can be written in the form

$$\begin{aligned} s_{0k}(t) &= A_k \cos(2\pi f_k t)\, g(t) + A_k \cos(2\pi f_k t)\, g(t - T_k) \\ s_{1k}(t) &= A_k \cos(2\pi f_k t)\, g(t) - A_k \cos(2\pi f_k t)\, g(t - T_k) \end{aligned} \quad (3)$$

where T_s is the sampling interval; $A_k = \sqrt{2E_k/T_k}$; E_k is the energy per symbol; T_k is the symbol duration and is assumed to be equal for all k ($T_k = T_b, \forall k$); and $g(t)$ is a function that is defined as $g(t) = 1, 0 \leq t < T_k$, $g(t) = 0$, otherwise.

A pair of signals, $s_1(t)$ and $s_2(t)$, is said to be orthogonal if it satisfies $\int_{-\infty}^{+\infty} s_1(t)s_2(t) = 0$ [7]. Clearly, each pair, $[s_{0k}(t), s_{1k}(t)], \forall k$, represents an orthogonal pair. Since the frequency set of $f_k, k = 1, .., 4$ is designed to be sufficiently orthogonal, then the waveforms; $s_{bk}(t), \forall k, b = 0, 1$; are also sufficiently pairwise orthogonal.

Now, consider a discrete version of Eq. (3), which is easily obtained by replacing the continuous-time variable t by nT_s, where $n = 0, 1, 2, ...$ and T_s is the sampling interval. This results in the discrete waveforms, $s_{0k}[nT_s]$ and $s_{1k}[nT_s]$, or simply $s_{0k}[n]$ and $s_{1k}[n]$. For the orthogonality property to be maintained for these sampled signals, the sampling theorem has to be respected. Namely, the sampling rate, $F_s = 1/T_s$, has to satisfy $F_s \geq 2W$, where W is the highest frequency in the frequency band of interest. The maintenance of the orthogonality property can be directly implied from the relationship between the Fourier transforms of the continuous-time and the corresponding discrete-time signals.

Fig. 3 plots the auto- and cross-correlations of an example pair, $[s_{01}[n], s_{11}[n]]$ at a frequency of 40 kHz and another orthogonal pair, $[s_{02}[n], s_{12}[n]]$ at 41 kHz, in the noise-free case. Fig. 3 (a) shows the autocorrelation and the cross-correlations of $s_{01}[n]$ with the other three orthogonal waveforms, whereas, Fig. 3 (b) shows similar correlations pertaining to $s_{11}[n]$. Each waveform consists of 512 samples.

The autocorrelations looks much like those of narrowband signals; the magnitude grows towards the zero lag point and the autocorrelation has both negative and positive peaks. The positive peak occurs exactly at zero lag. Due to the phase-shift in the middle of the waveform, the autocorrelation of $s_{11}[n]$ has two smaller peaks in addition to the main peak. The cross-correlation of $s_{01}[n]$ and $s_{11}[n]$ (orthogonal waveforms at the same frequency) has four (two negative and two positive) peaks and two (one negative and one positive) *valleys*. This cross-correlation has its minimum absolute value at zero lag. The cross-correlation of each of $s_{01}[n]$ and $s_{11}[n]$ with the other orthogonal pair (lower plots) are generally smaller in magnitude compared to the upper plots, and they do not show a significant variation in magnitude towards zero lag.

These correlation properties can be exploited to implement a simple and robust detection scheme without precise symbol synchronization, as will be shown in the next section.

4. SYSTEM OPERATION

The operation of the system can be summarized by the following two steps:

I. Detection of the arrival of the frequency sequence.

II. Decoding of the binary bits.

The decoding requires synchronization. Next, it will be shown that that synchronization and decoding can be performed simultaneously.

4.1. Detection

To detect the arrival of the first cycle (4 frequencies) of the transmission, matched filters or equivalently bandpass filters can be used [8]. Here, we benefit from the *discrete Fourier transform* (DFT) [9] to detect the frequencies of interest. DFT can normally be employed as a filter bank and is computationally attractive [10].

The detection system is depicted in Fig. 2. This scheme is inspired by FHSS world [8] and operates as follows. The receiver takes a window of N samples of the received signal, $x[n]$, with N equivalent to the number of samples in the period of a frequency duration ($2T_b$). The DFT of the data is computed and the energy in each of the four bands is estimated, as indicted by the summation over frequency band B_k. The estimated energy from each of the four stages is compared to a suitably selected threshold. If the threshold is exceeded, a binary output, b_k, that indicates this fact, is set to take the value 1, otherwise, it remains 0. The binary outputs pertaining to different bands are delayed such that the system output will accumulate a value of 4 only when the four frequencies are received in the correct order. This is supposed to protect against false detection due to in-band interferences. Detection is declared when the output reaches the value 4. Note that the binary signals, b_k, are just indicators of the arrival of the corresponding frequencies and have nothing to do with the actual data encoded therein.

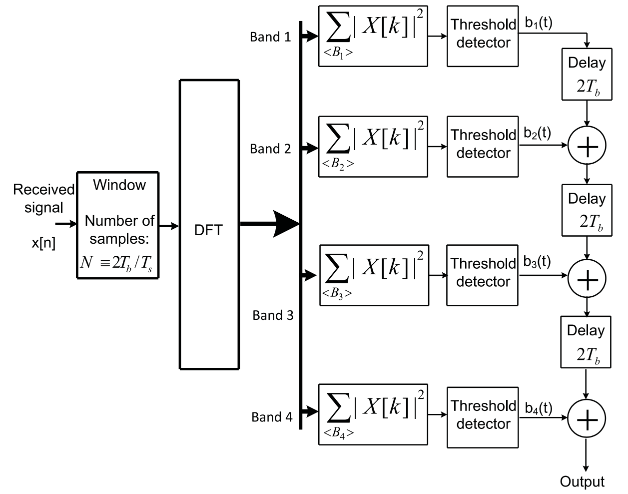


Fig. 2: Detection System.

Due to misalignment, windowing can result in a portion of a received frequency being in one window and another portion in the next window. If the proportion is close to 0.5, this may lead to a duplication of detection. Naturally, the receiver will select the first detected sequence. This kind of alignment results in an uncertainty of a half window length in the location of the start of the pattern. This should be dealt with in the decoding stage.

4.2. Decoding

Let us redefine the received sequence, $x[n]$, such that $x[0]$ is the received sample that coincides with the start of the frequency sequence according to the detection stage. Also, let $s_{bk}[0]$ coincide with the same starting point. As preceded, there is an uncertainty of $N/2$ samples in estimating this discrete time. With this in mind, the portions of the cross-correlations that are relevant to the decoding of the data carried in frequency f_k ($k = 0, .., 4; b = 0, 1$) are [9]

$$E_{bk}[l] = \sum_{n=l}^{l+N} x[l] s_{bk}[n-l],$$
$$l = -\frac{N}{2} - (k-1)N, ..., -\frac{N}{2} - kN \quad (4)$$

Now, define the function

$$\Delta E_k[l] = |E_{0k}[l]| - |E_{1k}[l]|. \quad (5)$$

This function combines the contributions of the correlations of the two different waveforms at each frequency. This definition is aimed at decreasing any peaks other than the main peak (see Fig. 4), and therefore, reducing the probability of picking a wrong peak when sources of error (e.g., noise) are present.

In the ideal case, it is expected that $|E_k[l]|$ will have its peak value at $l = p_k$, exactly when the test data s_{bk} has the best alignment (or synchronization) with the corresponding frequency in the received signal. To determine which binary is being received, $E_k[p_k]$ is tested. If $E_k[p_k] > 0$, then binary 0 is received; and if $E_k[p_k] < 0$, then 1 is received.

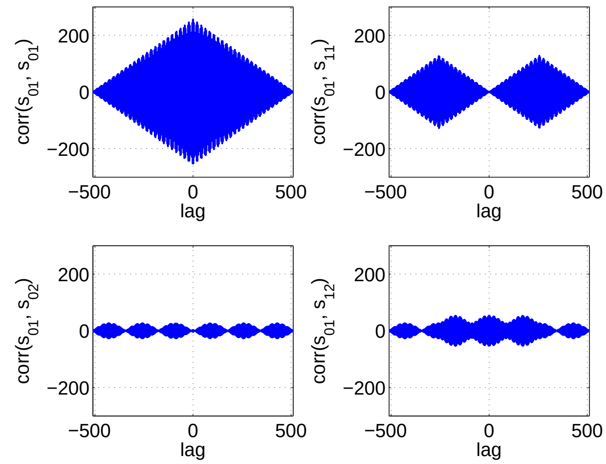
Fig. 4 depicts examples of the functions involved in the decoding process for one frequency, where binary 1 is received. Fig. 5 plots bit-error rate versus E_b/η_0, where η_0 is the noise spectral density of an *added white Gaussian noise* (AWGN). The figure represents the detection and decoding performance obtained from simulation for DBPSK data transmitted over a frequency of 39 kHz as a part of a sequence of frequencies of 38, 39, 40 and 41 kHz and $T_b = 1.5$ ms. Performance is clearly seen to improve as E_b/η_0 increases. The bit-error rate decreases to very small values for high E_b/η_0 ratios.

5. EXPERIMENTAL TESTS

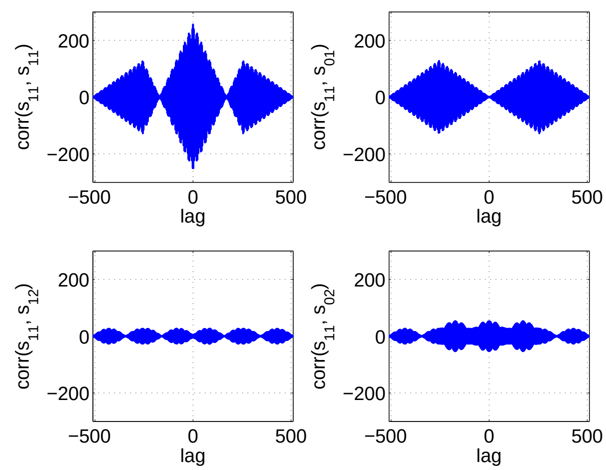
The proposed communication system was tested experimentally in a real office environment. The experimental setup consisted of an ultrasonic transmitter (400ST160 from Prowave), an ultrasonic sensor (SPM0204UD5 from Knowles), a sensor interface, a DSP board (SMT361A from Sundance) and a PC. The sensor interfacing circuit was designed to amplify the sensor outputs from a few millivolts to a maximum of two volts that could be admitted by the DSP board. Also the interface performed high pass filtering that suppressed all sound contents in the audible range.

Throughout the experiments, signals were sampled at a rate of 167.857 samples/s, which results in approximately 512 samples per frequency. The frequency channels 38, 39, 40 and 41 kHz were used. In the tests, codewords of 8 bits (two cycles of the frequency set) were transmitted at random intervals. The test space is depicted in Fig. 6, with the locations of the transmitter and receiver marked and the test codewords given in the attached table. In all of the tests, the transmitter was situated at the location marked Tx_1, whereas the receiver was moved between two different locations; Rx_1 and Rx_2. The difference between the two receiver locations is that Rx_1 was *sufficiently far* from any rigid reflector, while Rx_2 was about 49 cm from a rigid wall. According to the system parameters, the latter distance is sufficiently small for *early* multipath pertaining to a certain frequency to catch up with the direct path reception of the same frequency.

For each receiver location, codeword-1 was transmitted 15 times at random intervals. The same is repeated for codeword-2 and codeword-3. The results are summarized in Table 1, where the errors are classified according to the number of bits in error in each of the 8-bit codewords. It can be seen that for Rx_1, the receiver managed to detect and decode each of the three codewords correctly without a single bit error in all case. On the other hand, for Rx_2, multipath (due to



(a)



(b)

Fig. 3: Example of Correlations functions for $s_{01}[n]$ and $s_{11}[n]$.

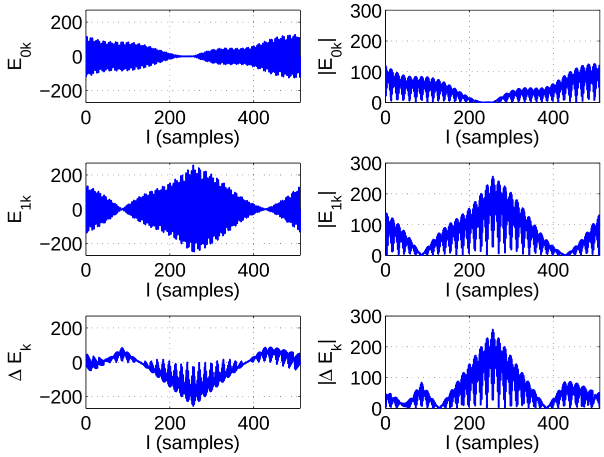


Fig. 4: Example of the functions involved in the decoding process.

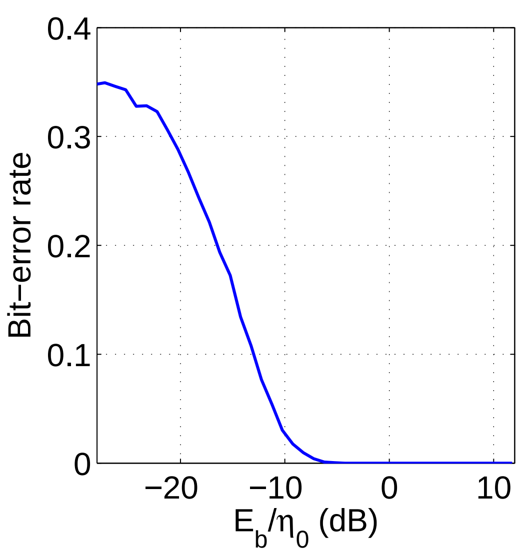


Fig. 5: Bit-error rate versus E_b/η_0.

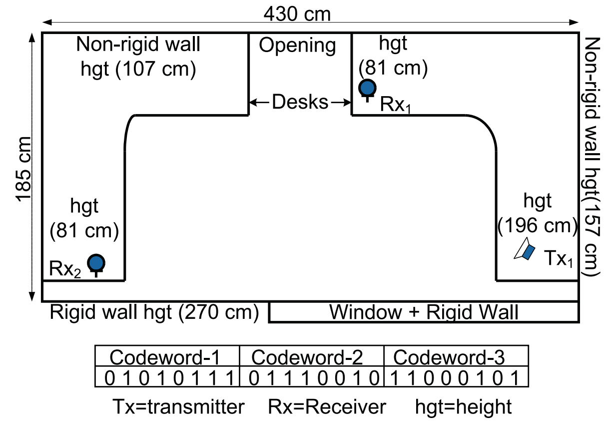


Fig. 6: Experimental Setup.

reflections from walls) seems to have affected performance. It can be seen that 1-bit errors are more dominant, while larger bit errors are quite rare. The latter results highlight the difficulties that surrounds the operation of the systems in a reverberant environment such as indoors. Namely, acoustic systems can be expected to face performance loss near rigid reflectors. To overcome such difficulties, error correction codes can be utilized, which requires increasing the transmitted data size. This is acceptable in applications with low system throughput requirements.

Table 1: The number of occurrences of different number of bit errors in 15 transmissions per codeword for receiver locations Rx_1 and Rx_2.

	$\mathbf{Rx_1}$			$\mathbf{Rx_2}$		
Error	CW-1	CW-2	CW-3	CW-1	CW-2	CW-3
1-bit	0	0	0	2	5	3
2-bit	0	0	0	0	1	0
3-bit	0	0	0	0	0	1
>3-bit	0	0	0	0	0	0

6. CONCLUSIONS

This paper presented an ultrasonic based binary communication system for indoor environments. The system was designed to transmit transmitter identification and synchronization information to aid the operation of an ultrasonic based indoor positioning system. Experimental results demonstrated the robustness of the proposed communication system. However, performance was found to degrade close to rigid reflectors such as walls, a point that is proposed for future consideration.

7. REFERENCES

[1] A. Ward, A. Jones and A. Hopper, "A New location technique for the active office," *IEEE Personal Communications*, vol. 4, no. 5, pp. 42–47, Oct 1997.

[2] N. B. Priyantha, A. Chakraborty and H. Balakrishnan, "The cricket location-support system," *Proc. of MobiCom 2000*, Boston, MA, USA, Aug 6–11 2000, pp. 32–43.

[3] S. Holm, O. B. Hovind, S. Rostad and R. Holm, "Indoors data communications using airborne ultrasound," *IEEE Int. Conf. on Acoust., Speech, Sign. Proc. (ICASSP2005)*, Mar 2005, vol. 3, pp. 957–960.

[4] S. Holm, "Airborne ultrasound data communications: The core of an indoor positioning system," *IEEE Int. Ultrasonics Symp.*, Sep 2005, pp. 1801–1804.

[5] Y. Fukuju, M. Minami, H. Morikawa and T. Aoyama, "DOLPHIN: An autonomous indoor positioning system in ubiquitous computing environment," *IEEE Workshop on Software Technologies for Future Embedded Systems*, May 2003, pp. 53–56.

[6] S. Haykin, "Communication Systems," *Wiley*, 4th edition, 2001.

[7] J. G. Proakis, "Digital Communications," *McGraw-Hill*, 4th edition 2000.

[8] D. Torrieri, "Principles of Spread-Spectrum Communication Systems," *Springer*, 2005.

[9] J. G. Proakis and D. G. Manolakis, "Digital Signal Processing," *Prentice-Hall, Inc.*, 3rd edition, 1996.

[10] C. S. Burrus, C. S. and T. W. Parks, "DFT/FFT and Convolution Algorithmsg," *John Wiley & Sons*, 985.