DIGITAL SIGNAL PROCESSING FOR RADIO MONITORING*

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Abstract

A radio monitoring system based on a receiver, interfaces and FFT analyzer is described. The controller of the system evaluates spectra and the frequencies and levels of sinusoid signals (carriers) are accurately measured by interpolation of spectral values. The interpolation procedure and a new interpolation algorithm is described. The Cramer-Rao Lower Bound is also calculated for real— and complex-valued input data. Real-life measurement results are also presented.

Introduction

The usual tasks of radio monitoring are the measurements of field strength, frequency and direction of incidence, as well as the observation of spectrum occupancy and identification of sources. Recent developments in technology allow a powerful solution of these tasks by digital signal processing of the intermediate-frequency signal of receivers. Besides of the repeatable and stable high-accuracy measurements, the automatic recognition and classification of radio-frequency signals are also possible [1].

In this paper a monitoring system based on digital signal processing is described, an interpolation algorithm for interpolation of FFT spectra of sinusoid signals is presented and the Cramer-Rao bound to the interpolated parameters together with simulation and measurement results are given.

Radio monitoring with digital signal processing

The high-resolution monitoring system consists of an appropriate antenna and receiver, demodulator and a digital signal processing unit. For calibrated level measurements the antenna and receiver should also be calibrated. Depending on the actual parameters of instruments, a choice of demodulation schemes are available [2]. In case

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of synchronous detection, the receiver should have low phase noise. Recommended performance can be found in [3].

For broadcast monitoring, the digital signal processing unit is a FFT analyzer. The high resolution together with interpolation of spectra may call for an external frequency reference to synchronise all local oscillators to a single source.

Interpolation

The spectra obtained by the FFT analyzer can be either visually analyzed by the operator or processed by a computer. The evaluation of spectra means the identification of sinusoid signals (carriers), measurement of their parameters and measurement of bandwidth. In the following we deal with the measurement of sinusoid signals.

The identification and parameter measurements of sinusoid signals are performed in two steps:

— search for local maxima (course search),
— interpolation of spectral values (fine search).

Different methods can be used for the course search. See e.g., [4], [5].

The interpolation is usually done on the two largest spectral lines of periodogram [6] or to separate several peaks, recursive interpolation of complex spectra is performed [7], [8]. These interpolations, however, do not provide information on the stability of selected peak. A new interpolation algorithm [9] can be summarized as follows:

— absolute values of complex spectra with Hanning weighting are used,
— the negative-frequency terms and periodicity of spectra are neglected,
— the three largest spectral lines around a selected peak (ith line) are used: (L_{i-1}, L_i, L_{i+1})

The estimation of frequency increment ($\Delta\lambda$) of a sinusoid with a frequency: $2\pi(i+\Delta\lambda)/NT$

$$\Delta\lambda 1 = (A-2)/(1+A) \qquad A = L_i/L_{i-1} \tag{1}$$

$$\Delta\lambda 2 = (2-B)/(1+B) \qquad B = L_i/L_{i+1} \tag{2}$$

$$\Delta\lambda 3 = 2(1-C)/3(1+C) \quad C = L_{i-1}/L_{i+1} \quad \text{if} \quad |\Delta\lambda 3| \ll 1 \tag{3}$$

$$\Delta\lambda 3 = D(1-1-2/D^2) \qquad D = 3(1+C)/2(1-C) \quad \text{else.}$$

$$\sigma(\Delta\lambda) = \sqrt{\left((\Delta\lambda 1-\Delta\lambda)^2+(\Delta\lambda 2-\Delta\lambda)^2+(\Delta\lambda 3-\Delta\lambda)^2\right)/2} \tag{4}$$

Lower bound of estimate

A basic question of parameter estimation is the ultimate available accuracy. In spectral interpolation the most widely used bound is the Cramer-Rao Lower Bound (CRLB). Details can be found e.g., in [10] and [11].

The CRLB should be calculated for the commonly used parameter combinations of the digital signal processing. Here we consider a single sinusoid with additive, Gaussian noise, zero mean independent samples with σ^2 variance. We will consider the following four cases: complex or real input signal with or without weighting.

The joint propability density function of the sample vector [4]:

$$f_c(Z,\alpha) = (\sigma^2 2\pi)^{-N} \exp\left[-1/2\sigma^2 \sum_{n=0}^{N-1} (X_n-\mu_n)^2+(Y_n-\nu_n)^2\right] \tag{5}$$

$$f_r(Z,\alpha) = (\sigma\sqrt{2\pi})^{-N} \exp\left[-1/2\sigma^2 \sum_{n=0}^{N-1} (X_a-\mu_n)^2\right] \tag{6}$$

for complex and real input signals respectively, where

$$X_n = h_n(s(t_n)+n(t_n)) \quad 0 \leqq n \leqq N-1 \tag{7}$$

$$Y_n = h_n(\hat{s}(t_n)+\hat{n}(t_n)) \quad 0 \leqq n \leqq N-1 \tag{8}$$

$Z=X+jY$ and h_n is a sample of the weighting function (9)

$$s(t) = A_0\cos(\omega_0 t+\theta_0), \quad \hat{s}(t) = A_0\sin(\omega_0 t+\theta_0) \tag{10}$$

$$\mu_n = h_n(A\cos(\omega t_n+\theta)), \quad \nu_n = h_n(A\sin(\omega t_n+\theta)) \tag{11}$$

$$\alpha = [\omega, A, \theta]^T, \quad t_n = t_0+nT = (n_0+n)T, \quad T = 1/fs \tag{12}$$

The Fischer information matrix (J) is defined by:

$$\mathbf{J}_{ij} = E\left\{\frac{\partial}{\partial\alpha_i}(\ln f(Z;\alpha))\frac{\partial}{\partial\alpha_j}(\ln f(Z;\alpha))\right\} \tag{13}$$

The following bound holds for the variance of estimation (CRLB):

$$\text{var}\{\hat{\alpha}_i\} = [\mathbf{J}^{-1}]_{ii} \tag{14}$$

After elementary calculations the Fischer information matrix can be written as follows:

$$\mathbf{J} = (1+c)/2\cdot\mathbf{J}^c+(1-c)/2\cdot\mathbf{J}^r, \tag{15}$$

where $c=0$ for real input signal
$c=1$ for complex input signal, and

$$\mathbf{J}^c = \begin{bmatrix} A^2T^2\sum(n_0+n)^2h_n^2 & 0 & A^2T\sum(n_0+n)h_n^2 \\ 0 & \sum h_n^2 & 0 \\ A^2T\sum(n_0+n)h_n^2 & 0 & A^2\sum h_n^2 \end{bmatrix} \tag{16}$$

$$\mathbf{J}^r = \begin{bmatrix} J_{11}^r & J_{12}^r & J_{13}^r \\ J_{21}^r & J_{22}^r & J_{22}^r \\ J_{31}^r & J_{32}^r & J_{33}^r \end{bmatrix} \tag{17}$$

$$J_{11}^r = -A^2T^2 \sum (n_0+n)^2 h_n^2 \cos 2\varphi_n$$

$$J_{12}^r = J_{21}^r = AT \sum (n_0+n) h_n^2 \sin 2\varphi_n$$

$$J_{13}^r = J_{31}^r = -A^2 T \sum (n_0+n) h_n^2 \cos 2\varphi_n$$

$$J_{22}^r = \sum h_n^2 \cos 2\varphi_n$$

$$J_{23}^r = J_{32}^r = -A \sum h_n^2 \sin 2\varphi_n$$

$$J_{33}^r = -A^2 \sum h_n^2 \cos 2\varphi_n$$

and $\varphi_n = \omega t_n + \theta$, moreover Σ denotes $\sum_{n=0}^{N-1} (\,.\,)$.

It is easy to show that every element of J^r tends to zero if the argument φ changes rapidly enough, that is:

$$\mathbf{J}^r \to \mathbf{0} \quad \text{if} \quad \omega_0 \gg 2\pi/NT \tag{18}$$

(18) and (16) mean that the CRLB of α is independent of the frequency of sinusoid for complex signals and the dependence can be neglected if (18) holds for real signals. (15), (16) and (17) can be evaluated numerically if the actual parameters are known. Simplifications can be made and simple closed forms for $\hat{\alpha}_i$ can be given with the following assumptions:

— no weighting, $h_n = 1$,
— $n_0 = 0$,
— $N \gg 1$.

With the above assumptions the CRLB expressions ($\Delta\omega = 2\pi/NT$):

$$\frac{\text{var}\,\{\hat{\omega}\}}{\Delta\omega^2} \geqq \frac{3\sigma^2}{\pi^2 A^2 N} \quad \text{if } A \text{ known or not } \omega \text{ and } \theta \text{ unknown} \tag{19}$$

$$\frac{\text{var}\,\{\hat{\omega}\}}{\Delta\omega^2} \geqq \frac{3\sigma^2}{4\pi^2 A^2 N} \quad \text{if } \theta \text{ known and } A \text{ known or not} \tag{20}$$

$$\text{var}\,\{\hat{A}\} \geqq \frac{\sigma^2}{N} \quad \text{for all combinations} \tag{21}$$

$$\text{var}\,\{\hat{\theta}\} \geqq \frac{4\sigma^2}{A^2 N} \quad \text{if } A \text{ known or not } \omega \text{ and } \theta \text{ unknown} \tag{22}$$

$$\text{var}\,\{\hat{\theta}\} \geqq \frac{\sigma^2}{A^2 N} \quad \text{if } \omega \text{ known and } A \text{ known or not} \tag{23}$$

Simulation and measurement results

From Expn (19) follows that with digital signal processing of 12 bit mantissa and $N=1024$, the quantization and rounding noise of the FFT process enable an ultimate interpolation factor of at least 10000. This high interpolation factor is advantageous in radio-frequency measurement to reduce measuring time when an accuracy of frequency measurement in the order of $10^{-10}...10^{-12}$ might be necessary.

The following figures give some detailes about the simulation and real-life measurement result conducted at the Technical University of Budapest, Microwave Department and the Hungarian Measuring Station at Tárnok.

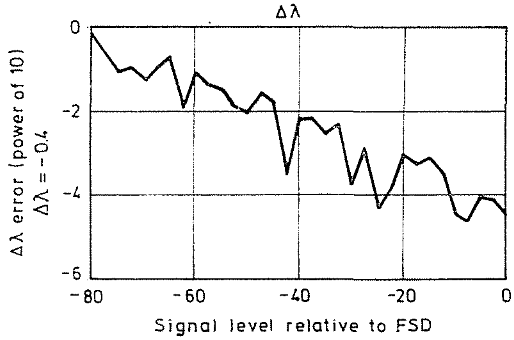

Fig. 1. Simulation result of interpolation. The time buffer of a Brüel Kjaer Type 2033 Signal Analyzer was loaded by $N=1024$ samples of a real sinusoid with $i=256$, $\Delta\lambda$ nominal$=-0{,}4$. The A amplitude of sinusoid was changed in 2,5dB increments between 0 dB and -80 dB relative to full scale. The bias and standard deviation of $\Delta\lambda$ for 100 independent runs are shown

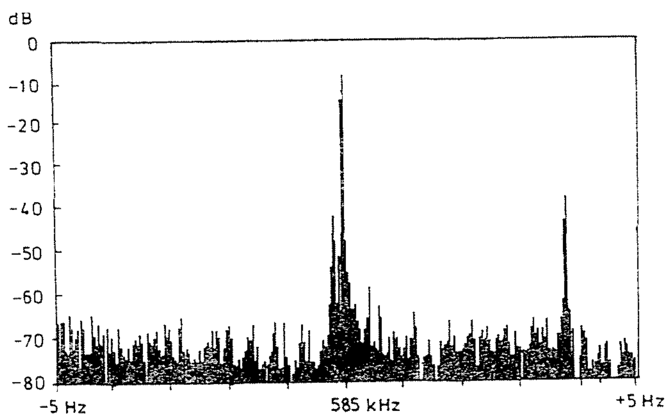

Fig. 2. Spectrum obtained by the FFT analysis of downconverted radio-frequency signal. Channel: 585 kHz. Horizontal span: +5 Hz. Vertical scale: relative level in dB µV/m. Date and time: 13th of March, 1987, 18:40—19:40 local time

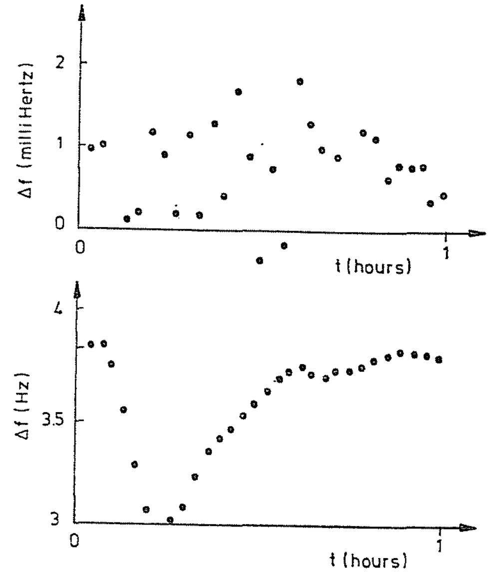

Fig. 3. Frequency offset relative to the nominal 585 000 Hz as a function of time, measured in the two largest peaks in the spectrum

a. Frequency offset of the largest peak

b. Frequency offset of the second largest peak. The random variation of Fig. 3.a. is within +2−1 millihertz and is due to the sky-wave propagation. The periodic and smooth variation of Fig. 3.b. is about 1 Hz peak-to-peak. This indicates the poor stability of transmitter

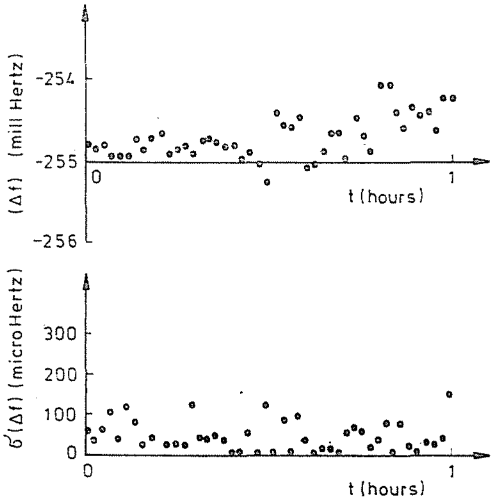

Fig. 4. Results of stability measurement on the Solt (Hungary) transmitter. Date and time: 13th of March, 1987, 14:00—15:00 local time

a. shows the offset of transmitter frequency relative to the nominal 540000 Hz. In

b. the deviation of interpolation is plotted. The two curves clearly show the stable carrier frequency with a constant frequency offset (−254,5 mHz average)

References

1. JONDRAL, F. Automatic Classification of High-Frequency Signals, Signal Processing, October 1985, pp 177—190.
2. CCIR Report 272—4: Frequency Measurements at Monitoring Stations, CCIR Report 273—5: Field Strength Measurements at Monitoring Stations, ITU, CCIR, Geneva, 1986.
3. NOVÁK, I.: Computer-Controlled High-Resolution Radio Monitoring System, Proceedings of the 1986 Wroclaw EMC Symposium, Part III, pp 1035—1044.
4. RIFE, D. C., BOORSTYN, R. R.: Single-Tone Parameter Estimation from Discrete-Time Observations, IEEE Tr. on Information Theory, No *5*, Sept. 1974, pp. 591—598.
5. NARDUZZI, C., OFFELLI, C.: Real-Time High Accuracy Measurement of Multifrequency Waveforms, Proceedings of ICASSP 1987.
6. RIFE, D. C., VINCENT G. A.: Use of the Discrete Fourier Transform in the Measurement of Frequencies and Levels of Tones, The Bell System Technical Journal, February 1970 pp. 197—228.
7. JAIN, V. K.: High-Accuracy Analog Measurements via Interpolated FFT, IEEE Tr. on Instrumentation and Measurement, No *2*, June 1979, pp. 113—121.
8. RENDERS H., SCHOUKENS J., VILAIN G.: High Accuracy Spectrum Analysis of Sampled Discrete Frequency Signals by Analytical Leakage Compensation, IEEE Tr. on Instrumentation and Measurement, No *4*, December 1984, pp 287—192.
9. NOVÁK, I.: Radio-Frequency Measurements with Interpolation in FFT Spectra, IMEKO 6th TC7 Symposium, Budapest, Hungary June 10—12, 1987.
10. RIFE, D. C. and BOORSTYN, R. R.: Multiple Tone Parameter Estimation From Discrete-Time Observations, The Bell System Technical Journal, November 1976, pp. 1389—1410.
11. FRIEDLANDER, B., PORAT, B., A General Lower Bound for Parametric Spectrum Estimation, IEEE Tr. on Acoustics, Speech and Signal Processing, No *4*, August 1984, pp. 728—733.

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