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AN FM/RDS (RADIO DATA SYSTEM) SOFTWARE RADIO

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ABSTRACT

This paper describes a software radio for the reception of FM (frequency modulation) and RDS (radio data system) signals. Both the FM and RDS receiver use quadrature sampling and simple baseband signal processing. After FM demodulation the signal is further processed by the RDS receiver using matched filtering and feedforward phase estimation.

Index Terms **–** software radio, FM demodulator, RDS receiver

1. INTRODUCTION

This paper presents the design and development of a software radio for the reception of FM/RDS signals. RDS receivers are - besides their use in traditional FM radio receivers - important for navigation systems based on GPS (Global Positioning System). Such systems, if equipped with an RDS receiver or connected to an FM/RDS radio, receive information on road traffic conditions, which is transmitted by broadcast stations via RDS over the Traffic Message Channel (TMC). The implementation of the RDS receiver as a software radio allows easy integration on existing hardware with a general-purpose processor (GPP) or a digital signal processor (DSP).

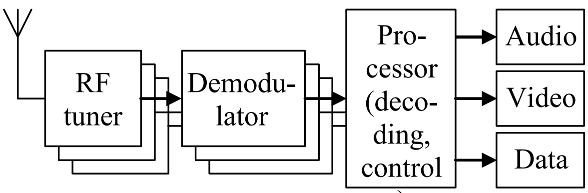

Figure 1 Typical broadcast receiver

Figure 1 shows a high-level block diagram of a typical broadcast receiver. This approach requires a dedicated RF tuner and demodulator for each standard. Even for audio broadcast a variety of analog and digital standards exist, like AM, FM, DAB (digital audio broadcasting), DRM (digital radio mondiale) and HD radio. Moreover, these standards use different frequency ranges and bandwidths in different regions of the word.

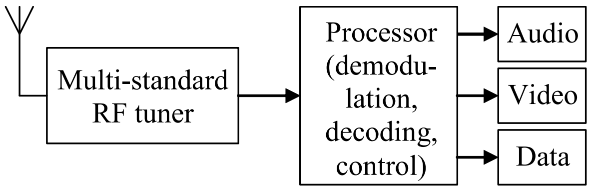

Figure 2 Multi-standard receiver

In a multi-standard receiver, the analog front-end is capable of receiving multiple broadcast standards (Figure 2). Different frequency ranges and bandwidths are typically accommodated by dedicated amplifiers and filters, followed by IQ-downconversion to baseband. The output at zero or low IF (intermediate frequency) is, after analog-to-digital conversion, demodulated and decoded in the processor.

In the following sections, we describe the signal processing of our FM/RDS software radio. Finally, we discuss different options for the implementation of the software radio.

2. FM RECEIVER

FM broadcast signals are transmitted in the frequency range of 87 MHz to 108 MHz. The baseband signal is the audio signal limited to a bandwidth of 15 kHz. Stereo transmission uses a multiplex scheme, where the sum of the left and the right channel, $x_l(t) + x_r(t)$, is transmitted in the lower frequency range and the difference of the stereo channels is modulated using DSB-SC (double-side band suppressed carrier) onto a 38-kHz carrier (Figure 3).

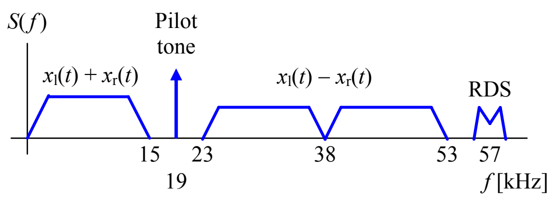

Figure 3 Schematic spectral representation of the FM stereo baseband signal

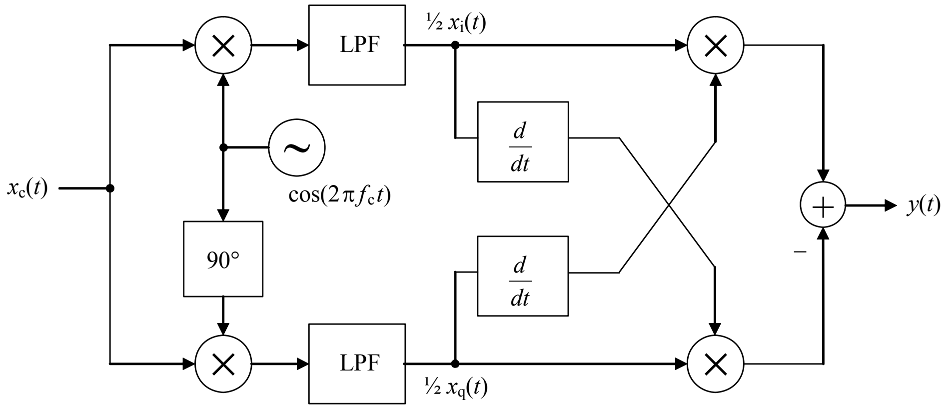

Figure 4 Block diagram of the FM demodulator

This multiplex scheme is compatible with monaural receivers which use the $x_l(t) + x_r(t)$ signal only. A pilot tone with frequency 38 kHz/2 = 19 kHz indicates the stereo signal to the receiver and is used for coherent demodulation of the DSB signal. As shown in Figure 3, the RDS signal is inserted above the stereo multiplex using a subcarrier of 57 kHz.

The baseband signal is FM-modulated onto a carrier using a frequency deviation of $f_\Delta = 75$ kHz. The bandwidth of the modulated signal is

$$B \approx 2(f_\Delta + 2 f_g),$$

where $f_g \approx 60$ kHz is the upper frequency limit of the baseband signal. Thus the FM signal has a bandwidth of 390 kHz.

2.1. Quadrature Demodulation

For FM, a number of demodulation schemes are known [1]. A demodulator particularly suited for digital implementation is shown in Figure 4. It is a quadrature demodulator, where the FM signal is first decomposed into its quadrature components $x_i(t)$ and $x_q(t)$. After further processing as shown in the block diagram the baseband signal is available at the output.

An FM signal is described by

$$\begin{aligned} x_c(t) &= A_c \cos\left(2\pi f_c t + 2\pi f_\Delta \int_0^t x(\tau)\, d\tau \right) \\ &= A_c \cos\left(2\pi f_c t + \varphi(t)\right), \end{aligned}$$

where f_c and A_c are the carrier frequency and amplitude, respectively, f_Δ is the frequency deviation and $x(t)$ is the baseband signal. We express $x_c(t)$ as

$$\begin{aligned} x_c(t) &= A_c \cos\left(2\pi f_c t + \varphi(t)\right) \\ &= A_c \operatorname{Re}\left\{\exp(j\varphi(t)) \exp(j 2\pi f_c t)\right\}. \end{aligned}$$

$\exp(j\,\varphi(t))$ is the complex envelope of $x_c(t)$. Its real and imaginary parts are the I (in-phase) and Q (quadrature-phase) signals $x_i(t)$ and $x_q(t)$,

$$x_i(t) = \cos\varphi(t), \quad x_q(t) = \sin\varphi(t),$$

where $\varphi(t)$ depends on the modulating signal according to

$$\varphi(t) = 2\pi f_\Delta \int_0^t x(\tau)\, d\tau \cdot$$

The quadrature components are obtained by multiplying $x_c(t)$ with $\cos(2\pi f_c t)$ and $-\sin(2\pi f_c t)$. The lowpass filters (LPF) following the multipliers in Figure 4 reject the double-frequency terms located at $2 f_c$, which result from the multiplication [2].

As further shown in Figure 4, $x_i(t)$ and $x_q(t)$ are differentiated and cross-multiplied giving

$$\begin{aligned} y(t) &= x_i(t) \frac{d\, x_q(t)}{dt} - x_q(t) \frac{d\, x_i(t)}{dt} \\ &= \frac{d\,\varphi(t)}{dt} \cos^2 \varphi(t) + \frac{d\,\varphi(t)}{dt} \sin^2 \varphi(t) \\ &= \frac{d\,\varphi(t)}{dt} = 2\pi f_\Delta\, x(t) \end{aligned}$$

where the factor 1/2 was omitted for simplification. As shown, the demodulator output equals the baseband signal $x(t)$ multiplied by an amplitude factor, which depends on the frequency deviation.

Although we did assume a perfectly synchronised local oscillator signal, the demodulator also works when the oscillator has a phase or frequency offset relative to the carrier of $x_c(t)$. A phase offset $\Delta\phi$, i. e. the local oscillator signal is $\cos(2\pi f_c t - \Delta\phi)$, cancels out in $y(t)$. A frequency offset $\cos(2\pi (f_c - \Delta f)\, t)$ results in a d. c. component proportional to Δf in $y(t)$ and can be removed easily.

2.2. Implementation Aspects

The demodulator of Figure 4 is implemented completely in the digital domain. The analog front-end is described in Section 4. A very simple digital version of the quadrature demodulation results, if the sampling rate is $f_s = 4 f_c$. This quadrature sampling scheme [3] is illustrated in Figure 5. The input signal $x_c(t)$ is sampled at $t = n\, T_s$, $T_s = 1/4 f_c$. The discrete-time signal $x_c(n)$ is multiplied with the discrete-time versions of $\cos(2\pi f_c t)$ and $-\sin(2\pi f_c t)$. As shown in Figure 5, even samples $n = 2k$ multiplied by $(-1)^k$ represent the in-phase component, and odd samples $n = 2k + 1$ multiplied by $(-1)^k$ represent the quadrature component:

$$x_i(2k) = (-1)^k x_c(2k) \,,$$
$$x_q(2k+1) = -(-1)^k x_c(2k+1) \,.$$

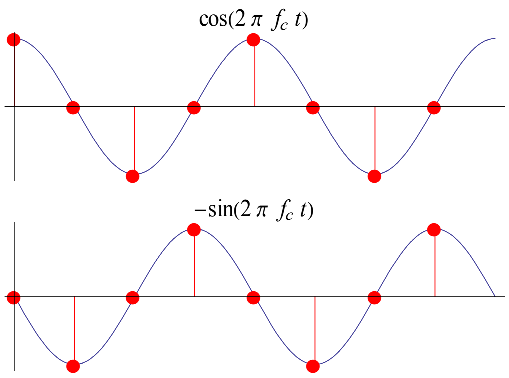

Figure 5 Sampling instants of the locally generated quadrature carriers for $f_s = 4 f_c$

Quadrature sampling results in quadrature components sampled at a rate $f_s/2 = 2 f_c$. The sampling instants are T_s apart. In order to obtain quadrature components sampled at the same instants, the signals can be interpolated.

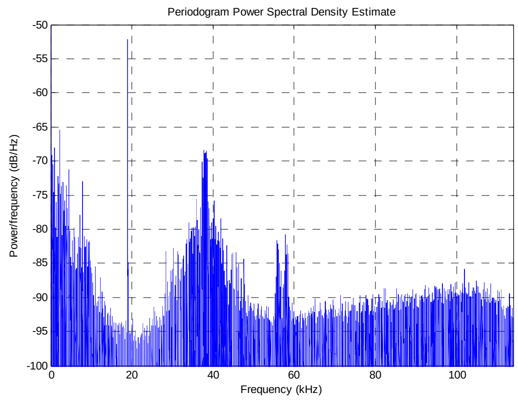

Figure 6 Measured spectrum of the FM stereo baseband signal

The differentiators can be implemented as FIR filters. A first-order approximation based on the difference of subsequent samples allows a very efficient implementation. Using this simplification and quadrature sampling, the demodulator can be implemented with a few lines of code.

Figure 6 shows the spectrum of an FM signal recorded with a sampling oscilloscope und post-processed with Matlab. It clearly shows the audio signal in the range up to 15 kHz, the pilot tone at 19 kHz, the stereo multiplex signal centered at 38 kHz and the RDS signal at 57 kHz.

3. RDS RECEIVER

RDS allows the transmission of digital data in parallel to the audio signal. RDS enables services like program identification, automatic tuning, radio text, traffic message channel (TMC) and many others.

3.1. RDS Modulation and Coding

The RDS baseband signal has a bit rate of 1187.5 bit/s. The information is transmitted in groups with a length of 104 bits. A group comprises 4 blocks of 26 bits. A block is composed of a 16-bit information word and a 10-bit CRC for error detection (Figure 7). To enable group synchronization, offset words A, B, C and D are added modulo 2 to the CRC of blocks 1, 2, 3, and 4, respectively [4].

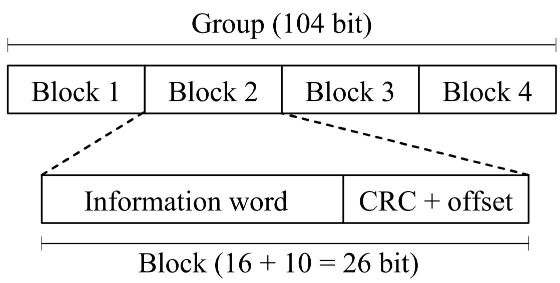

Figure 7 RDS frame structure

The baseband signal is differentially encoded and Manchester encoded, resulting in a symbol rate of 2375 baud. A square-root raised-cosine filter with a roll-off factor $\alpha = 1$ serves as a pulse shaping filter. Finally, the signal is modulated onto the carrier using binary phase-shift keying [2]. The modulated signal has a bandwidth of approximately 4.8 kHz. The carrier frequency of 57 kHz is an integer multiple of the frequency of the 19-kHz stereo pilot tone. The symbol rate is the carrier frequency divided by 24.

3.2. Principle of Operation

After FM demodulation, the RDS signal is extracted by a bandpass filter (BPF). A typical receiver structure for demodulation and decoding is shown in Figure 8. After coherent demodulation, the signal is applied to a filter matched to the transmit pulse shape, i. e. a square-root raised-cosine filter with $\alpha = 1$ [2]. The detected symbols are Manchester decoded. After differential decoding, the bit stream is searched bit-by-bit for valid check words. A valid check word is found if the syndrome calculated by the CRC decoder corresponds to one of the offset words. When block and group synchronization is achieved, the RDS messages can be further processed.

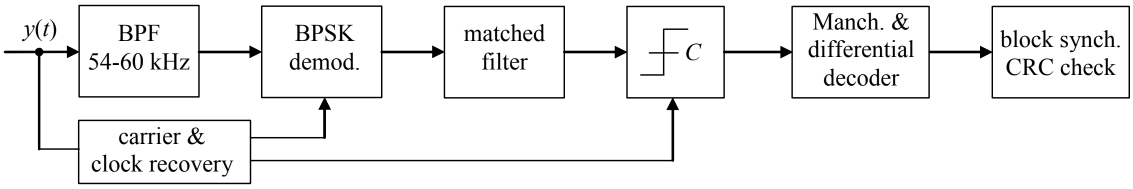

Figure 8 Block diagram of a typical RDS receiver

Although the pilot tone can be extracted and used for coherent demodulation, this approach is not ideal for digital implementation because of the filtering required. Therefore, our receiver uses quadrature sampling and a feedforward algorithm for phase recovery [5].

Quadrature sampling was described in Section 2.2. Here, the sampling rate is $f_{s2} = 4{\cdot}57$ kHz = 228 kHz. The baseband signals are applied to the matched filters followed by downsampling to one sample per symbol.

The BPSK signal is given by

$$y(t) = \mathrm{Re}\left\{\sum_k p(t-kT)\,\exp\big(\mathrm{j}(2\pi f_c t + \theta_c + \varphi_k)\big)\right\}$$

$$= \mathrm{Re}\left\{\sum_k p(t-kT)\,\exp\big(\mathrm{j}(\theta_c + \varphi_k)\big)\exp\big(\mathrm{j}(2\pi f_c t)\big)\right\}$$

$$= \mathrm{Re}\left\{\, y_{lp}(t)\,\exp\big(\mathrm{j}(2\pi f_c t)\big)\,\right\},$$

where $p(t)$ is the transmit pulse shape, θ_c is the unknown carrier phase and $\varphi_k \in \{0, \pi\}$ depends on the transmitted symbols. After quadrature sampling, matched filtering and downsampling, assuming $p(nT-kT) = 1$, we have

$$y_{lp}(k) = y_i(k) + \mathrm{j}\,y_q(k) = \exp\big(\mathrm{j}(\theta_c + \varphi_k)\big).$$

The random phase φ_k is removed by taking the square of the complex envelope, because $2\varphi_k \in \{0, 2\pi\}$:

$$y_{lp}^2(k) = \exp\big(\mathrm{j}(2\theta_c + 2\varphi_k)\big) = \exp\big(\mathrm{j}2\theta_c\big).$$

Thus the carrier phase is obtained from the phase angle of the squared samples of $y_{lp}(k)$. Because we have to deal with noisy samples, we average over multiple symbols,

$$\hat{\theta}_c(k) = \frac{1}{2}\arg\left(\sum_{n=-N}^{N} y_{lp}^2(k+n)\right),$$

and the phase correction is applied to the complex envelope:

$$y_{lp,corr}(k) = y_{lp}(k)\,\exp\big(\mathrm{j}\hat{\theta}_k\big) \approx \exp\big(\mathrm{j}\varphi_k\big).$$

Figure 9 shows scatter plots obtained by post-processing measured signals with Matlab. The scatter plots show the samples in the complex plane, with the real part given by the in-phase component and the imaginary part given by the quadrature phase component of the complex envelope. The upper figure shows the complex envelope before the phase correction is applied. The lower figure shows the typical BPSK signal constellation, which was obtained after applying the phase correction algorithm.

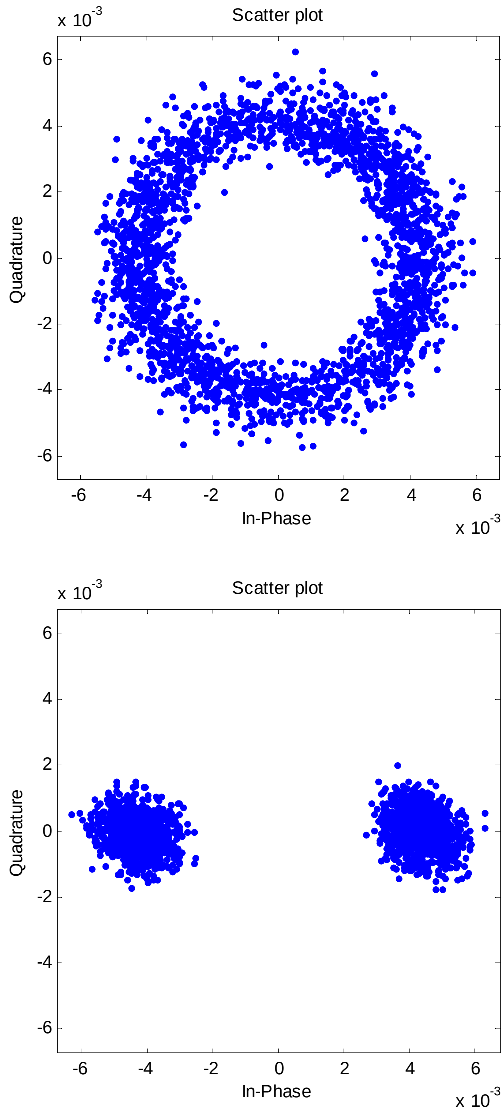

Figure 9 Signal constellation before and after phase recovery

4. IMPLEMENTATION OF THE SOFTWARE RADIO

For software radios, a number of implementation options ranging from custom hardware and software development to the use of off-the shelf components and free software are available (see e. g. [6], [7], [8]). We decided to use a C6713 DSP Starter Kit (C6713 DSK) from Spectrum Digital with a Texas Instruments TMS320C6713 225 MHz floating point DSP. As RF front-end, we use a MAXIM direct-conversion to low IF tuner IC MAX2170. At the output, the analog down-converted FM signal centered at an IF of 2.048 MHz is available. The IF signal is sampled with a high-speed ADC THS1206 (Texas Instruments). For this device an evaluation module (EVM) is available, which is compatible with the C6713 DSK.

Figure 10 shows a block diagram of the software radio. In the first quadrature sampling stage, where the IF signal from the analog front-end at 2.048 MHz is processed, bandpass sampling is applied in order to reduce the sampling rate. With bandpass sampling, the acceptable sampling rates are given by

$$\frac{2f_H}{n} \le f_s \le \frac{2f_L}{n-1}, \quad n \le \left\lfloor \frac{f_H}{B} \right\rfloor,$$

where f_H and f_L are the upper und lower frequency limit and $B = f_H - f_L$ is the bandwidth of the signal [3]. An acceptable sampling rate means that there is no aliasing in the spectrum of the undersampled signal. In our application, with a center frequency f_{IF} = 2.048 kHz, B = 400 kHz and n = 3, the sampling rate must be in the range 1.499 MHz $\le f_s \le$ 1.848 MHz.

Bandpass sampling comes along with a frequency shift $f_d = f_{IF} - f_s$ of the signal spectrum. In order to combine bandpass sampling with quadrature sampling, the condition $f_s = 4 f_d$ results in f_s = 1.6384 MHz. Since the FM demodulator tolerates a small frequency offset, the sampling rate is set to f_s = 1.632 MHz, which simplifies further signal processing.

While bandpass sampling reduces the required sampling rate, the signal-to-noise ratio becomes worse because (n – 1) noise bands of bandwidth B between dc and f_L are aliased into the signal band. In this respect bandpass sampling differs from shifting the signal downward in frequency with an analog mixer, which preserves the signal-to-noise ratio. With bandpass sampling, it is important to reduce the noise outside the signal bandwidth.

At the output of the FM demodulator, the signal has a sampling rate of $f_{s1} = f_s/2$ = 816 kHz. As explained in Section 3.2, quadrature sampling at the RDS receiver input requires a sampling rate f_{s2} = 228 kHz. This ratio is given by the continued fraction expansion

$$\frac{816}{228} = 4 + \cfrac{1}{-2 + \cfrac{1}{-3 + 1/3}} = \frac{68}{19}.$$

After resampling with f_{s2} and bandpass filtering, the in-phase and quadrature components of the RDS signal are derived from the samples. A sampling rate of $f_{s2}/2$ corresponds to 48 samples per symbol. The signal is further processed and decoded as described in Section 3.

Currently the project is in the stage of hardware testing and code development. For code development, two approaches will be tested: writing code in C, and the use of Matlab model-based design tools from (Simulink, Real-time Workshop and Target for TI C6000).

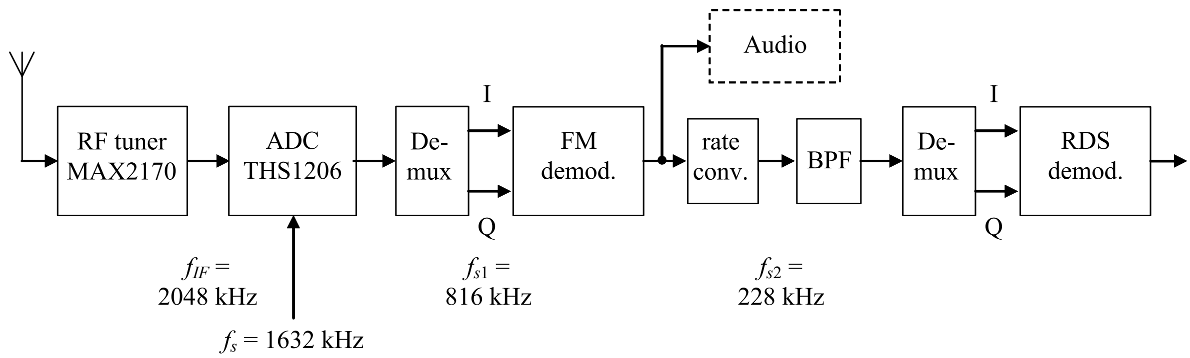

Figure 10: Block diagram of the demodulator

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