ELECTRONICS AND ELECTRICAL ENGINEERING
ISSN 1392 – 1215 2009. No. 8(96)
ELEKTRONIKA IR ELEKTROTECHNIKA

SIGNAL TECHNOLOGY
T121
SIGNALŲ TECHNOLOGIJA

Simulation of Lithuanian Version of DVB–T System

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Introduction

Economical, technical and organizational aspects of terrestrial and cable digital video broadcasting (DVB–T/C) networks rollout and switching off analog video and audio transmission in Lithuania were discussed in details before the beginning or in the early stage of these processes [1, 2]. The first DVB–T network in Lithuania was launched in 2007. After hot discussions between scientists, engineers and officials H.264/MPEG–4 AVC standard for video compression was adopted. The system parameters offered by DVB–T standard [3] were chosen taking into account the country peculiarities and long–term objectives to develop synchronous single frequency network (SFN) and mobile broadcast network in the whole country. These are:

- The outer Reed–Solomon coding RS(204,188, t=8);
- The outer convolutional interleaving with I=12;
- The inner error protection punctured convolutional code with coding rate 3/4;
- The inner block interleaver;
- The number of carriers – 8192 (8K mode);
- Modulation scheme – 64 QAM;
- Guard interval – 1/4.

The Reed–Solomon code allows correcting up to 8 random erroneous bytes in a received word of 204 bytes. The inner coding and inner interleaving are necessary to avoid bursts of erroneous bits being fed to the Viterbi decoder. The inner interleaver consists of a combined bit and symbol interleaver, where “symbol” here refers to the bits being transmitted by one carrier during one symbol time. The requirements on guard interval length determine the number of carriers of an OFDM system. The necessary guard interval length depends on the transmitter distance in a SFN, or on the delay of natural echoes in case of a conventionally planned network. Since SFN is scheduled to be based on existing transmitter sites, it was necessary to use a guard interval of approximately 250 µs. The requirement for a guard interval determines the number of carriers: a guard interval of 250 µs can be achieved with an OFDM system with a symbol time of 1 ms and hence a carrier distance of 1 kHz, resulting in 8000 carriers in an 8 MHz wide channel. An OFDM signal is implemented using an inverse FFT. The FFT size has to be equal to 2^N. So 8000 carriers require N=13. The FFT size will then be 8192. Therefore maximum allowed guard interval length of 1/4 and the "8K mode" are extremely well suited both for single transmitter operation and for small and large SFN networks. The chosen combination of constellation (64 QAM) and code rate (3/4) enables the system to cope with a large variety of channel characteristics.

The first Lithuanian DVB–T receiver TF–401 with Lithuanian tuner module KS–1601 [4] was designed and serial production was launched in 2007. However there are neither models nor simulation results of DVB–T systems like Lithuanian, especially having in mind Lithuanian tuner module KS–1601.

We present below the main information about the known mathematical models of phase noise of different local oscillators (LO) as well as designed SIMULINK models of phase noise and of DVB–T system with the parameters identical to the chosen in Lithuania.

The model of phase noise of local oscillator

Phase noise is the most important parameter of any oscillator and it affects a system perturbing both the amplitude and the phase of an oscillator output. The situation is even more complicated with receivers and results in reciprocal mixing in the mixer. If we mix a clean radio frequency signal with a modulated LO source then a modulated intermediate frequency (IF) will be the result. The effect can be explained by suggesting that the noise components are additional LOs that are offset from the main carrier. Each of them mixes other signals that are offset from the LO by the receivers IF. The noise is the sum of an infinite number of infinitesimal components spread over a range of frequencies, so the signal it mixes into the IF and as a result they are spread into an infinite number of small replicas, all at different frequencies. Therefore the phase noise in an oscillator is known to degrade the bit error ratio (BER) performance of COFDM systems, especially for an input of higher–order modulation like used in DVB–T systems [5].

There is no lack of papers dealing with miscellaneous oscillators phase noise theories and mathematical models [6–11]. All authors come to one common conclusion: the phase of the noise is uniformly random and the instantaneous magnitude of the noise obeys a probability distribution very similar to normal distribution. However power spectral density (PSD) of the noise depends on the oscillator type.

Considering LO as an amplifier with positive feedback Leeson derived the expression for the noise PSD in so called Leeson's equation form [7]. Additional estimation of the flicker noise of an active device puts Leeson's equation into an expanded shape [8]

$$\Phi(f_m)=\left[1+\frac{1}{f_m^2}\left(\frac{f_0}{2Q_L}\right)^2\right]\frac{NkT\Delta F}{P_{avs}}\left(1+\frac{f_c}{f_m}\right),\quad \Delta F=1\,, \tag{1}$$

where $\Phi(f_m)$ – phase noise PSD, f_m – carrier offset frequency, f_0 – carrier center frequency, f_c – flicker corner frequency of the active device, Q_L – Q factor of loaded tank circuit (loaded Q), P_{avs} – average power through the resonator, N – noise factor of the active device, k – Boltzman constant, T – absolute temperature.

Phase noise is usually specified in dBc/Hz at a given offset, where dBc is the level in dB relative to the carrier.

Representation of an oscillator phase noise PSD according to the equation (1) in logarithmic scale is given in Fig. 1.

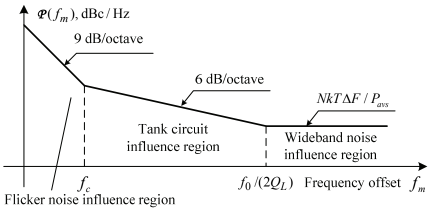

Fig. 1. Representation of phase noise PSD of an oscillator

The close to carrier noise with a slope of 9 dB/octave is due to the flicker noise of the active device and has a cut–off at the flicker corner frequency f_c. Typically, it is less than 1 kHz. The 6 dB/octave section is due to phase noise according to Leeson's equation and is a function of loaded Q, noise factor, power and temperature. It is spread from f_c to $f_0/(2Q_L)$, which is in the region of a few MHz. Above carrier offsets of $f_0/(2Q_L)$ noise is broad–band noise as defined by $NkT\Delta F/P_{avs}$.

The values of parameters f_c, Q_L, f_0, P_{avs} from equation (1) depend on LO circuitry. In the main MOSFETs have a higher f_c than JFETs or bipolar transistors which is usually below 2 kHz for the latter. Loaded Q value fluctuates approximately between 5 and 50.

However the main type of oscillators used in DVB–T set–top boxes is a variable frequency oscillator (VCO) used as a component of a phase locked loop (PLL). Such VCOs require a method of converting the PLL control voltage to frequency and this is normally done with a varactor diode (Vari–capacitance diode). Unfortunately, any noise on the PLL control voltage and any internally generated noise will modulate the carrier, increasing the overall phase noise performance. The increase of the phase noise in dBc/Hz is given by the formula [8]:

$$\Phi(f_m)=20\lg\frac{K_{VCO}U_n}{\sqrt{2}f_m}\,, \tag{2}$$

where K_{VCO} – the VCO gain constant, U_n – the equivalent noise voltage modulating the varactor.

Therefore, the total phase noise of the open loop VCO will be the power sum of the oscillator phase noise given by the Leeson's equation added to the varactor phase noise just given. Varactor modulation noise is the most significant in broadband high frequency VCOs, because the VCO gain constant is large. Simulation and measurement results [8–11] show in the extreme varactor noise and flicker noise can dominate the main cause of noise in an oscillator – that generated by the resonator.

Full PLL system (closed loop VCO) mostly consists of active components such as VCOs, mixers, frequency multipliers, frequency dividers, active or passive filters, reference signal sources which are vulnerable to the surroundings. Therefore, it is likely to have an unexpected result such as a peak or a non–uniform phase noise. The different spectrum calculation and measurement techniques used by different authors show similar results [8–11]. The PLL (closed loop VCO) output spectrum follows that reference input signal spectrum for low offset frequencies and open loop VCO spectrum for large offset frequencies. In between, the output spectrum is almost constant. The offset frequency beyond which the PLL output spectrum follows the open loop VCO spectrum is approximately equal to the bandwidth of the PLL. At high offset frequencies the PLL output PSD is slightly higher than the open loop VCO spectral density. This is because there is no loop filter present in the circuit to remove the high frequency noise component of the phase noise of the reference signal. If a loop filter is included, the PLL output coincides with the open loop VCO output for high offset frequencies.

A precise phase noise model could be designed having measurement results of the specific local oscillator. In our case, we should have been satisfied with a few values of the measured local oscillator's phase noise power spectrum density (PSD), which have been presented in [4].

Having in mind the above presented analysis, we decided to employ in Fig. 1 depicted phase noise PSD graph as a basis to design the SIMULINK model of phase noise generator. To get the model noise as close as possible to the noise of a real oscillator, we corrected the absolute values of the graph in Fig. 1 according to the measurement data [4] and then synthesized an appropriate digital filter using SIMULINK FDA Tool block. As a noise source for excitation of the digital filter the Gaussian noise generator was used. The noise at the output of the filter is represented in the magnitude–angle form. For the baseband simulation the complex form is needed. Therefore the model was supplemented with the block converting noise representation to complex form. The complete phase noise generator model is presented in Fig. 2.

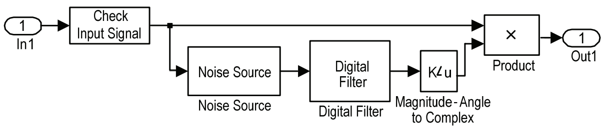

Fig. 2. The model of the phase noise generator

The model of the DVB–T system

The SIMULINK demo version model (file name – commdvbt.mdl) was chosen as a prototype of our model. Unfortunately, this model is designed for 2K mode system. To customize it to our requirements we needed slightly to change the structure of some blocks and to revolutionize the file comdbvttablegen.m describing the system parameters. The changes affected the data stream structure of all blocks. The final version of the model is depicted in Fig. 3.

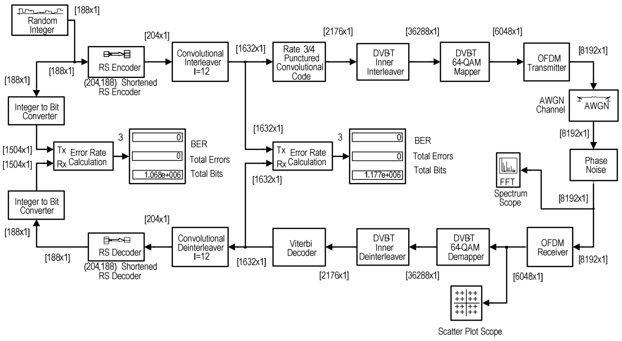

Fig. 3. The model of the DVB–T system

The model contains the following blocks:

- A source of random integers between 0 and 187;
- An outer encoder (Reed–Solomon encoder);
- An outer interleaver (convolutional interleaver);
- An inner encoder (punctured convolutional encoder);
- An inner interleaver (in–depth);
- A baseband 64–QAM modulator (mapper);
- An OFDM transmitter (calculator of Inverse Fast Fourier Transform);
- An additive white Gaussian noise (AWGN) channel;
- A source of phase noise
- An OFDM receiver (i.e. calculator of Fast Fourier Transform);
- A baseband 64–QAM demodulator (demapper);
- An inner deinterleaver;
- An inner decoder (i.e. Viterbi decoder);
- An outer deinterleaver;
- An outer decoder;
- Error statistic calculators;
- Display icons that show the error statistics (bit error rate – BER, the total number of errors, and the total number of bits) while the simulation runs;
- A scatter plot scope that shows the received signal constellation;
- A spectrum scope that shows the received signal spectrum;
- The numbers near the connection lines indicate the dimensionality of data streams.

Simulation results

Some results illustrating displays gained during the simulation process are presented in Fig.4. Fig 4a shows the signal spectrum plot in the case of noiseless channel and removed phase noise.

Note, that the Gaussian and phase noise corrupt the received signal so that each constellation point transforms into a cluster of points. But the clusters are different. The phase of the Gaussian noise phasor is distributed uniformly. So the cluster of constellation points representing the sum of phasors of transmitted signal and Gaussian noise take the circle shape [11], see Fig. 4b. For the case where phase noise has been added to the signal, two effects are distinguished: common phase error (a rotation of the signal constellation) and inner carrier interference (similar to additive Gaussian noise) [11]. Common phase error part is proportional to the received carrier. However its phasor is perpendicular to the carrier's phasor. It corresponds to rotation of the signal constellation by an appropriate angle. The implication is that the clusters are close to arcs, see Fig. 4c. There is also "thermal–noise–like" addition, i.e. blurring rather than rotation of the constellation. In our case the common phase error effect prevails over the inner carrier interference.

Fig. 4d shows signal constellation in the case of presence of Gaussian and phase noises.

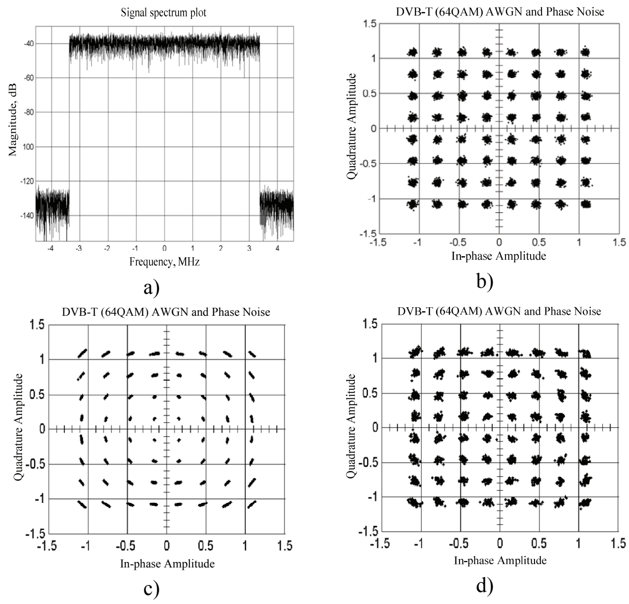

Fig. 4. Spectrum plots (a); Constellations: b) – SNR=30 dB, phase noise negligible), c) – SNR=100 dB, phase noise –68 dBc at 1 kHz, d) – SNR=30 dB, phase noise –68 dBc at 1 kHz

However the main objective of our simulation was to evaluate the performance of the system corrupted by Gaussian and phase noise. Therefore we carried out some simulation experiments with different signal–to–noise ratios (SNR) in AWGN channel and with different phase noise levels at different frequency offsets. Bit stream of 10^9 bits was used during each experiment.

Some graphs illustrating simulation results are represented in Fig. 5.

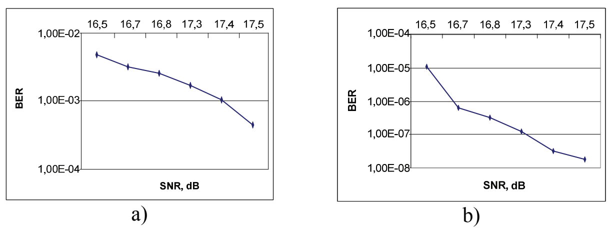

Fig. 5. BER versus SNR in the channel; a – after the Viterbi decoder; b – after RS decoder

Simulation results enabled us to evaluate the SNR and phase noise levels at which the critical values of BER, inadmissible for qualitative reception of signal, could be obtained. Critical BER value after Viterbi decoder (bellow denoted as CV1) is equal to 10^{-3} and critical BER value after RS decoder (bellow denoted as CV2) is equal to 10^{-7}. Our results show:

- Critical value CV1 is reached when SNR=17,1 dB, and CV2 – when SNR=16,9 dB. The estimated confidence interval is the same.
- Critical value CV1 is reached when phase noise level is –59,5 dBc/Hz at 1 kHz offset, and CV2 – when phase noise level is –61 dBc/Hz at 1 kHz offset. The difference between these two numbers is statistically unreliable, as they both fall into the same confidence interval. The maximum phase noise level of the tuner module KS–1601 at 1 kHz offset is –78 dBc/Hz [4]. The implication is that having noiseless channel the tuner has approximately 17 dBc/Hz reserve in phase noise.
- Having 10 kHz frequency offset critical value CV1 is reached with phase noise level –69 dBc/Hz, and CV2 – with –70,5 dBc/Hz. The confidence interval is the same for both numbers. However the maximum phase noise level of the tuner at 10 kHz offset is –82 dBc/Hz [6]. So the reserve is 12 dBc/Hz, i.e. 5 dBc/Hz less than at 1 kHz offset.
- Having maximum values of phase noise specified in [4] (–78 dBc/Hz at 1 kHz offset and –85 dBc/Hz at 10 kHz offset), CV1 and CV2 are reached with the same SNR level ≅17,6 dB. The implication is that maximum level of the tuner KS–1601 phase noise reduces the performance of the system as far as a reduction of SNR in the channel approximately by 0,5 dB.

Acknowledgements

This work was supported by the Lithuanian State Science and Studies Foundation under contract No B23

Conclusions

Simulation results of DVB–T system with parameters settled in Lithuania demonstrate creditable resistance of the system to the Gaussian channel noise and tuner phase noise. Critical BER values after Viterbi and RS decoders are reached with SNR≅17 dB or with phase noise level ≅–60 dBc/Hz at 1 kHz offset and ≅–70 dBc/Hz at 10 kHz offset. The maximum level of the tuner KS–1601 phase noise reduces the performance of the system as far as a reduction of SNR in the channel approximately by 0,5 dB.

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Received 2009 09 02

V. Deksnys, A. Čitavičius. Simulation of Lithuanian Version of DVB–T System // Electronics and Electrical Engineering. – Kaunas: Technologija, 2009. – No. 8(96). – P. 99–102.

This paper deals with the SIMULINK model of DVB–T system with parameters settled in Lithuania. Special attention has been given to the design of the tuner phase noise model. The simulation results are presented. The critical values of signal–to–noise ratio in the channel and phase noise level of the tuner determining the critical value of bit error rate were estimated. Ill. 5, bibl. 11 (in English; abstracts in English, Russian and Lithuanian).

В. Декснис, А. Читавичюс. Моделирование литовской версии системы цифрового наземного телевизионного вещания // Электроника и электротехника. – Каунас: Технология, 2009. – № 8(96). – С. 99–102.

Рассматриваются вопросы построения СИМУЛИНК модели цифровой наземной ТВ системы с параметрами принятыми в Литве. Особое внимание уделяется построению модели фазового шума местного генератора приемника. Приводятся результаты моделирования. Оценены допустимые уровни отношения сигнал – шум в канале и фазового шума местного генератора приемника, вызывающие критический уровень ошибочных битов. Ил. 5, библ. 11 (на английском языке; рефераты на английском, русском и литовском яз.).

V. Deksnys, A. Čitavičius. Lietuviškos skaitmeninės antžeminės TV sistemos versijos modeliavimas // Elektronika ir elektrotechnika. – Kaunas: Technologija, 2009. – Nr. 8(96). – P. 99–102.

Nagrinėjami skaitmeninės antžeminės TV sistemos su Lietuvoje priimtais parametrais SIMULINK modelio sudarymo ir modeliavimo klausimai. Ypatingas dėmesys atkreiptas į TV priedėlio vietinio generatoriaus fazinio triukšmo modelio sudarymą. Pateikiami modeliavimo rezultatai. Įvertintas leistinas signalo ir triukšmo santykis kanale ir fazinio triukšmo lygis imtuve, lemiantys leistiną ribinį klaidingų bitų dažnį. Il. 5, bibl. 11 (anglų kalba; santraukos anglų, rusų ir lietuvių k.).