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Abstract - We show that a free of assumptions approach to the Doppler effect (plane wave and “very small” period assumptions) leads to a Doppler factor which depends on the involved frequencies. The result is that the Doppler effect shifts differently the different frequencies present in the studied electromagnetic radiation.

I. INTRODUCTION

The Doppler factor formula is one of the most frequently derived equation in classical and in relativistic physics as well. In many cases, we make assumptions concerning the magnitude of the involved frequencies, the source-receiver distance and the character of the wave (plane or spherical).

We consider that in a Doppler Effect experiment we compare the time interval between the emissions of two successive wave-crests by the source (emission period), with the time interval between theirs reception, by the observer (reception period). The emission period is measured in the rest frame of the source (proper period of emission). The reception period is measured in the observer’s rest frame (proper period of reception). A Doppler factor formula relates the two periods mentioned above via a Doppler factor. The Doppler factor depends on the followed scenario, which prescribes the relative motion of receiver and source, and on the assumptions made concerning the magnitude of the physical quantities related by it.

We consider that speaking about a physical quantity it is advisable to define the observer who measures it, when and where he performs the measurement and the measuring devices involved in the measurement. We measure time intervals as a difference between the readings of the same clock (proper time interval) or as a difference between the readings of two distant synchronized clocks, of the same inertial reference frame. In a Doppler Effect experiment performed in an electromagnetic wave, we consider the rest frame of the source (Ks(XsOxYs)) and the observer’s rest frame (Kr(XrOrYr)) as well. The axes OxXs and OxXr are overlapped, the corresponding axes are parallel to each other and the relative motion takes place in the positive direction of the overlapped axes. At each point of the plane defined by the axes of the mentioned reference frames, we find instantaneously two clocks belonging to the involved reference frames, each synchronized in its rest frame in accordance with a synchronization procedure proposed by Einstein. The involved clocks are Cs(0,0) and Cr(0,0) located at the origins Os and Or. They read ts=tr=0 when the origins mentioned above are located at the same point in space. An important consequence of the scenario is that at a given time we find at a given point in space a clock Cs(xs,ys) of Ks(xsOxys) reading ts and a clock Cr(xr,yr) of Kr(xr,oryr) reading tr.

We consider separately two cases: The case when source S is at rest and observer R moves relative to it and the case when R is at rest and S moves relative to it. In the case of a scenario which involves a stationary source S and a uniformly moving receiver R, the Doppler formula:

\[ D = \frac{T_S}{T_R} = \frac{f_R}{f_S} = \frac{1 - \frac{v_R}{c}^2}{1 - \frac{v_s}{c}^2} \]  

is in large use. Ts represents a proper time interval measured as a difference between the readings of clock Cs(0,0) attached to the stationary source. Tr represents a proper time interval as well measured as a difference between the readings of a clock Cr attached to observer R. Vs represents the instantaneous velocity of R and D represents a Doppler factor. As not depending on the involved periods (frequencies fS=1/Ts and fR=1/TR) the Doppler factor defined by Eq.(1) is linear. Consequently, if S emits a radiation in which more frequencies are present, the supposed linearity of Doppler shift makes that all the present frequencies are shifted by the same Doppler factor, characteristic for the followed scenario. It is surprising that the Doppler factor (1) does not depend on the source-observer distance.

Authors derive Eq.(1) taking as a starting point the relativistic invariance of the phase of a plane
electromagnetic wave$^3$. The concept of plane wave is a mathematical construction with not much physical support. It is associated with the “very large” source-receiver distance assumption. Using Eq.(1) many Authors fail to mention that fact and use it to support special relativity$^5$.

The purpose of our paper is to derive Doppler formulas free of assumptions concerning how small or how large are the involved physical quantities. Such derivations show that Eq.(1) holds only in the case of “very small” period (“very high” frequency) assumption. The “very small” period assumption is associated with the concept of locality in the period measurement by a moving receiver. Making the locality assumption, we consider that the reception period is small enough in order to consider that R receives two successive wave crests from the same point in space. Moreau$^6$ introduced the concept of non-locality in connection with an accelerating receiver who performs the hyperbolic motion. Measuring the reception period as a time interval between the receptions of two successive wave crests, we infer that during that time interval R has no enough information in order to reckon the reception period.

II. A REALISTIC APPROACH TO THE DOPPLER EFFECT

2.1. Stationary source and moving observer

The scenario we propose involves a source S of electromagnetic waves, located at the origin O$_S$ of its rest frame K$_S$(X$_S$,O$_S$,Y$_S$). An observer R moves with constant velocity $v_R=v$ parallel to the O$_S$X$_S$ axis (to the ground), at an altitude $d$. We consider the problem in the rest frame of the source. In order to find out the result of the measurements performed by R we use the time dilation formula. $R_s(0,d)$ represents an instantaneous position of R when the clocks of the K$_S$(X$_S$,O$_S$,Y$_S$) frame read $t_S=0$. He receives there a wave crest emitted by S at $-d/c$. $R_n(n)\rightarrow$ represents a later instantaneous position of R where he receives the $n$-th emitted wave crest at a time $t_n$, measured by S, at a distance $vt_n$ from its initial position (Figure 1). Pythagoras’ theorem applied to Figure 1 leads to:

$$d^2 + v^2 t_n^2 = c^2 \left( t_n + \frac{d}{c} - nT_S \right)^2 . \quad (2)$$

Solved for $t_n$ Eq.(2) leads to:

$$t_n = T_S \frac{\left( n - \frac{d}{c} f_S \right) + A_n}{\sqrt{1 - \left( \frac{v}{c} \right)^2}^2} , \quad (3)$$

where:

$$A_n = \sqrt{\left( \frac{d}{c} f_S \right)^2 + \frac{v^2}{c^2} \left( n - \frac{2d}{c} f_S \right)} . \quad (4)$$

A clock $C(t_n)$ attached to R and reading $t_R=0$ at location $R_0$, reads $t_{R,n}$ at location R. We have in accordance with the time dilation effect:

$$t_n = \frac{T_R \cdot n}{\sqrt{1 - \left( \frac{v}{c} \right)^2}} \quad (5)$$

and the proper time in K$_R$ when R receives the n-th crest is:

$$t_{R,n} = T_S \frac{n - \frac{d f_S}{c} + A_n}{\sqrt{1 - \left( \frac{v}{c} \right)^2}} . \quad (6)$$

The proper reception period $T_R=t_{R,n}-t_{R,n-1}$, measured in the observer’s rest frame, and the proper emission period $T_S$, measured in the rest frame of the source, are related by:

$$D = \frac{T_S}{T_R} = \frac{f_R}{f_S} = 1 + A_n - A_n - 1 . \quad (7)$$

A first important conclusion is that the Doppler factor $D$ is non-linear, as depending on $f_S$. The result is that if the radiation emitted by S is not monochromatic, the Doppler Effect shifts differently the components of different frequencies. In order to illustrate that fact consider that S represents a radio station which emits frequencies comprised between 16 Hz$<f_S<20000$ Hz or even higher, R being located on a jet plane which can move at supersonic velocities. We have in mind the fact that the acoustic signals are transformed in electromagnetic ones. The electromagnetic wave carries them at a velocity $c$. The Doppler Effect shifts those frequencies in the same way as the electromagnetic ones. We present in Figure 2 the variation of $D$ with $n$ for constant values of $d$ and $v$.

For $f_S \rightarrow \infty$(very high frequency assumption) Eq.(7) becomes:

$$D \propto \sqrt{1 - \left( \frac{v}{c} \right)^2} . \quad (8)$$

For $d \rightarrow \infty$ (plane wave assumption) Eq.(7) becomes:
For $n \to \pm \infty$ (longitudinal Doppler Effect) Eq.(7) becomes:

$$D_{n \to \pm \infty} = \frac{1}{\pm v c - \pm 1}.$$  

As we see, the plane wave assumption and the very high frequency assumption lead to the same result. Because the first wave crest was emitted in the positive direction of the OY axis ($\theta_R = 90^\circ$), we can consider that in the plane wave assumption R moves in a plane wave in which the rays are parallel to the OY axis, Eqs.(8) and (9) describing the transversal Doppler shift.

2.2. Stationary observer and moving source

We consider the experiment in the observer’s rest frame. The source S moves with constant velocity $v_S = v$ parallel to the OX axis (to the ground), at an altitude $d$. In a free of assumptions approach (Figure 3), $S_0$ represents an instantaneous position of S at $t_S = 0$, when it emits a wave crest received by R at $d/c$. $S_n$ represents a later position of S at time $nT_S'$, measured in $K_E$, when it emits the $n$-th wave crest at a distance $nT_S'v$ at from $S_0$.

The $n$-th wave crest is received by R at $t_n$ given by:

$$t_n = nT_S + \frac{r_n}{c} = nT_S' + \frac{d^2 + n^2T_S'^2v^2}{c^2}.$$  

A clock comoving with S measures between the emissions of two successive wave crests a proper time interval $T_S$ related to $T_S'$ by:

$$T_S' = \frac{T_S}{\sqrt{1 - \frac{v^2}{c^2}}}.$$  

so one can write:

$$t_n = T_S \frac{n + B_n}{\sqrt{1 - \frac{v^2}{c^2}}},$$

where:

$$B_n = c^{-1} \sqrt{(nv)^2 + (d/T_S')^2(1 - v^2c^{-2})}.$$  

The time interval between the receptions of two successive wave crests is:

$$T_R = t_n - t_{n-1} = T_S \frac{1 + B_n - B_{n-1}}{\sqrt{1 - \frac{v^2}{c^2}}}.$$  

We are now able to define a Doppler factor characteristic for the followed scenario:

$$D = \frac{T_S}{T_R} = \frac{f_R}{f_S} = \sqrt{1 - \frac{v^2}{c^2}}.$$  

As expected, we have in the case of the very high frequency and plane wave assumptions:

$$D_{n \to \pm \infty} = \frac{1}{\pm v c - \pm 1},$$  

whereas in the case of the longitudinal Doppler Effect we have:

$$D_{n \to \pm \infty} = \frac{1}{\pm v c - \pm 1}.$$  

In order to illustrate the results obtained above we present in Figure 4.a the variation of $D$ with $n$ in the case when S moves with supersonic velocities at low altitudes, whereas in Figure 4.b we consider the case when S moves with relativistic velocities but at astronomic altitudes, a situation of interest for astronomers.

III. A JUMP INTO THE FUTURE

We consider now that there are objects that can fly with relativistic velocities at low altitudes. Such scenarios lead to high Doppler shifts and illustrate in a convincing manner the way in which the Doppler Effect shifts differently the different frequencies present in the electromagnetic radiation. We present in Figure 5 the variation of $D$ with the order number $n$ of the emitted wave crest in the case of a stationary source and a uniformly moving observer. His motion takes place at an altitude $d = 10^8$ m, at a velocity $0.6 c$. In order to illustrate the fact that during the reception of two successive wave crests R has no enough information in order to reckon the period we mark the points which lead to a calculated value of $D$. We present in Figure 6 the variation of $D$ with $n$ in the case of the scenario which involves a stationary observer and a source moving with velocity $v = 0.6 c$ at an altitude $d = 10^8$ m. The calculated values of $D$ are marked in order to illustrate the fact that during the reception of two successive wave-crests R has no
enough information in order to reckon the Doppler factor $D$.

**IV. CONCLUSION**

The Doppler shift formulas derived without making the usual assumption of “plane wave” (“very high” frequency) clearly show the non-linear character of the Doppler factor. As a result, the Doppler Effect differently shifts the different frequencies present in the electromagnetic radiation. That fact could be of importance in Astronomy. In the case of the longitudinal Doppler Effect ($\theta = 0^\circ$ or $\theta = 180^\circ$) and in the case of the transversal Doppler Effect the different frequencies are equally shifted. In the case of oblique incidence, the Doppler Effect shifts more the lower frequencies.

**REFERENCES**


**Figure captions**

Figure1. Doppler effect experiment. It involves a stationary source of light located at the origin of its rest frame. An observer R moves with constant velocity $v$ parallel to the OX axis (to the ground) at an altitude $d$.

Figure2. The variation of the Doppler factor $D$ defined by Eq.(7) following the scenario presented in Figure 1, with the order number $n$ of the emitted wave crest and for different values of the frequency $f_S$ emitted by the source: 1 - 20 Hz; 2 - 100 Hz; 3 - 2000 Hz; 4 - 40000 Hz. a) : R moves with supersonic velocity $v=6x340$ m/s at an altitude $d=10^2$ m; b) : R moves with relativistic velocity $v=0.6c$ at an altitude $d=1.5x10^8$ m.

Figure3. Doppler effect experiment. It involves a stationary observer located at the origin of its rest frame. A source S moves with constant velocity parallel to the OX axis (to the ground) at an altitude $d$. Figure 4. The variation of the Doppler factor $D$ defined by Eq.(16) with the order number of the emitted wave crest for different values of the emitted frequency $f_S$: 1 - 20 Hz; 2 - 100 Hz; 3 - 2000 Hz; 4 - 40000 Hz; a) Source S moves with supersonic velocity $v=6x340$ m/s at an altitude $d=10^2$ m; b) Source S moves with relativistic velocity $v=0.6c$ at an altitude $d=1.5x10^8$ m.

Figure 5. The variation of the Doppler factor $D$ with the order number $n$ of the emitted wave crest for different values of the emitted frequency $f_S$: 1 - 20 Hz; 2 - 2000 Hz; 3 - 20000 Hz; 4 - 40000 Hz. In order to illustrate the fact that between the reception of two successive wave crests the observer has no enough information to reckon the frequency we have marked the time when a value of the frequency is calculated. a): the case of a stationary source and an observer moving with relativistic velocity $v=0.6c$ at low altitude $d=10^4$m; b) the case of a stationary observer and source moving at low altitude $d=10^4$ m and with relativistic velocity $v=0.6c$. 

![Figure 1](image1.png)

![Figure 2](image2.png)

![Figure 3](image3.png)

![Figure 4](image4.png)

![Figure 5](image5.png)
Abstract – In an attempt to reduce costs, many organizations are using collaboration software tools (such as NetMeeting, Agillion, CentraNow, Done.com, HotOffice, eRoom, MagicalDesks, TeamWave Workplace, and Vicinities.com) for delivering synchronous online training. Such applications, offer a variety of interactive features that can be used for training purposes but are not fully equipped to provide an instructionally sound learning experience. This paper presents the drawbacks of inexpensive collaboration tools and provides solutions for how to overcome technical and instructional limitations of such software applications when used in training.

Keywords: collaboration software, synchronous training

I. INTRODUCTION

Recent Internet-based technology has revolved around creating interactive meeting (or collaboration) software, which allows geographically dispersed individuals to work together on the Web. Anyone with a decent Pentium processor or PowerMac, an Internet/Intranet connection, and a browser can use such software that provides an array of collaboration features, ranging from simple chat rooms to complex audio, streaming video and multimedia interaction (Webb, 2004).

The most frequently used and inexpensive collaboration tools are NetMeeting, Agillion, CentraNow, Done.com, HotOffice, eRoom, MagicalDesks, TeamWave Workplace, and Vicinities.com. These software tools allow users from remote locations to share and work on the same applications in real time, exchange ideas during threaded discussions and white boards, and even answer polls on critical issues (Webb, 2004).

Due to the large range of interaction capabilities and information sharing that these Web-based interactive meeting software offer, and particularly because of their low costs, training organizations have started to use them for providing online synchronous instruction to students. Most of this software is either entirely free (e.g., NetMeeting), free for a minimum of users (e.g., Done.com, eRoom, and Vicinities.com), or offered at a very low price (Agillion, CentraNow, MagicalDesk, TeamWave Workplace, HotOffice).

Synchronous learning, which implies the simultaneous presence in time of students and teachers at a training event, is considered a popular instructional method due to the ability to provide student-student interaction (peer-learning), student-instructor interaction (mentored learning), while offering a more solid framework for calibration and expectations to keep students on track. Synchronous distance education draws from the solid foundation of traditional instruction, while reaching a geographically dispersed student population (Christensen & Cowley-Durst, 1998).

While synchronous distance education appeals to educators and trainers, its implementation requires expensive equipment, complex infrastructure, and technical support personnel with specialized skills. Such requirements lead to increased costs of operation and ownership.

Corporate training departments and academic training organizations prefer to maintain the benefits of synchronous instruction yet implement it at reduced costs due to constantly decreasing budgets allocated to training endeavors. Consequently, many trainers prefer to use low-priced live meeting software for training purposes (Sherry, 1996).

The problem of using inexpensive collaboration tools for providing synchronous training is that such applications are not properly equipped to produce and provide a sound instructional experience for students. This paper outlines the features and limitations of inexpensive collaboration tools and how technical and instructional drawbacks can be avoided. The paper also presents ideas for using inexpensive collaboration tools not only during training delivery, but also for training planning and development.
II. FEATURES OF INEXPENSIVE COLLABORATION SOFTWARE

Below are listed the most commonly available and used features of low-cost collaboration tools (e.g., NetMeeting, Agillion, CentraNow, Done.com, HotOffice, eRoom, MagicalDesks, TeamWave Workplace, and Vicinities.com).

Program Sharing. This feature allows sharing of multiple programs between a virtually unlimited number of participants. During a training event, the instructor can enable students to view shared programs in a frame, which makes it easy to distinguish between shared and local applications on students’ desktop. Instructors can also switch between multiple shared programs, approve students’ requests to work in a program, and allow or prevent others from working in an application. From an instructional perspective, this feature promotes learning because it allows students ample opportunities for hands-on practice.

White boards. The Whiteboard feature allows real-time collaboration with others via a graphic interface, which is typically similar to Microsoft's Paint program. When the whiteboard feature is invoked, it will typically appear in a window that can be seen by all users, and all users can collaboratively work on the document/object. Using the Whiteboard feature, students have the ability to:

- Review, create, and update graphic information (e.g., artwork, GIFs).
- Manipulate contents by clicking, dragging, and dropping information on the whiteboard with the mouse.
- Cut, copy, and paste information from any application into the whiteboard.
- Use different-colored pointers to easily differentiate between students’ comments.
- Save the whiteboard contents for future reference or for distribution amongst the students in the class.

Real-Time Chat. This feature supports real-time typed (text-based) conversations among an unlimited number of people. The chat feature allows students to type text messages to communicate with others during a class session, or to chat with one person or a group of people across multiple computers. The instructional value of this feature is that students can send a private message to instructors, therefore avoiding the potential pressure of revealing their question to the entire class. Students may also exchange private opinions and/or questions amongst themselves. In short, the real-time chat feature mimics the traditional classroom environment and it enhances it by providing better student privacy.

Audio/Video Conferencing. This feature allows the sharing of training content and applications using video and audio. Even though most inexpensive collaboration software tools do not provide optimal video/audio capabilities, at the bare minimum they do offer the ability to:

- Send and receive real-time video images at small resolutions.
- Send video and audio to a user who doesn't have video hardware.
- Use a video camera to instantly view objects, such as hardware devices, that are displayed in front of the lens.
- Ensure that people hear each other by adjusting the automatic microphone sensitivity level setting.

From an instructional perspective, the use of video may be effective when presented in the beginning of the training, to enable students and instructors to connect and give each other a visual reference. Video may also help when used to demonstrate psychomotor skills (e.g., repairing a piece of equipment), or when used to motivate and change someone’s attitude (e.g., presenting the story of an expert performer who is monetarily recognized for top behavior on the job).

File Transfer. This feature allows the instructor to send one or more files to everyone attending the class, or to one or more selected participants. A practical example of the instructional value of this feature is that a teacher can send a file to a student who can work on it and send it back during the class session.

File Storage. Using this feature, students and teachers have the ability to store and access information and create a secure, shared space that holds documents. Some collaboration tools will even allow version control features and keyword search. This feature is instructionally practical because students may work on documents simultaneously, save their work, and retrieve it when back on their jobs.

Security. Most inexpensive collaboration applications provide user authentication, password protection, and data encryption. Consequently, students and teachers are able to store and access data in a secure manner.

Some of the aforementioned collaboration software tools provide unique features, such as the ability to poll participants (CentraNow), schedule sessions automatically (Done.com), route documents through a pre-defined cycle (eRooms), password protect documents for certain users (HotOffice), offer multiple language capabilities (MagicalDesk), and the ability to customize the look and color of your workspace (Vicinities.com). All these features have the ability to boost the instructional experience during a synchronous online class.
III. LIMITATIONS OF COLLABORATION SOFTWARE WHEN USED FOR SYNCHRONOUS TRAINING

The limitations of inexpensive collaboration software, when used for synchronous training, are noticeable in two areas: technical and instructional.

A. Technical Limitation of Inexpensive Collaboration Software

Low-cost collaboration software tools do not provide the ability to quiz students and check their comprehension levels. In addition, they do not allow instructors to use any pre-test options, which would indicate the current level of students’ knowledge/skills. Pre-tests features would enable instructors to adjust the pace and flow of the class such that it matches students’ current experience and expertise (Auerbach, 2004).

When using inexpensive collaboration software, students do not have the ability to “raise their hands” (in more expensive virtual classroom software, students can do this by pressing certain icon options provided in the application). This feature would enable students to ask the teacher to modify the pace or flow of the instruction, which have a direct impact on training effectiveness.

The aforementioned live meeting software tools do not enable the “breakout groups” capability, which allows students to be divided into teams and interact around a specific issue. Being able to divide students in small teams and assign to them varied group tasks rests at the foundation of collaborative learning (Vygotsky, 1978).

Scheduling, tracking, and/or recording mechanisms are also missing in inexpensive live meeting software. Such tools do not have the capability to link to a learning management system (LMS) and do not allow the storing of students’ training history (e.g., course completions, scores/grades, training path, need for re-enrollment, etc.). In addition, these tools do not enable the recording of a class session so that students who are absent can re-play it or so that the instructor can include pre-recorded sessions in new classes when taking a break.

Furthermore, when using inexpensive live meeting tools, instructors do not have the ability to “see” who is absent (which student has either left the learning space or is not paying attention to the class). This defeats the purpose of an instructor-led environment, which is supposed to offer better class control and the assurance that everyone leaves the classroom with improved knowledge/skills.

Audio and video capabilities are under-developed in most inexpensive collaboration software. In order to avoid the stilled nature of online training and mimic the traditional classroom atmosphere, a lot of instructors would like to take advantage of video conferencing capabilities with collaboration software. Unfortunately, the poor visual quality of video does not currently attract or hold interest among the student population (Aldrich, 2004). Research shows that video conferencing capabilities are not that popular yet due to bandwidth and resolution limitations. Educators use them more than corporate training departments; both are awaiting the technology to be optimized before complete adoption (Mael, 2003).

In addition, most inexpensive collaboration software tools do not allow students to engage in asynchronous activities (e.g., starting threaded discussions prior to the live meeting and continuing them after the training is complete). A balanced combination of asynchronous and synchronous training options would benefit students who are not always able to align their schedules so they can be present with others in a training event at the same time.

B. Instructional Limitations of Inexpensive Collaboration Software

Even though collaboration tools may offer enough interaction features that accommodate teaching certain instructional objectives (e.g., how to manipulate data in an Excel spreadsheet, how to repair a network element, etc.), they are still inappropriately used because the instructors are either not trained on how to use such tools effectively or instructional designers who develop classes that are to be delivered via such media do not possess enough instructional design experience to recommend and develop successful instruction.

There are currently several providers of virtual classroom solutions that do offer students the optimal classroom experience from a distance, overcoming most of the limitations listed in this section: impeccable video and audio quality, taking control of the classroom, accessing administrative software on a Web-server, etc. Examples of virtual classroom providers are Centra, Interwise, Lotus Learning Space, Avalon Information Technologies, Pathlore, Horizon Long Distance Learning, and others (Wells, 2004). However, these sophisticated options for synchronous distance education come with very high price tags, mainly due to the high costs for servers and access license fees.

As previously mentioned, training organizations undergoing austere financial times are currently striving to avoid increased costs related to adopting sophisticated distance learning technology. The following section outlines ideas on how to overcome limitations of inexpensive collaboration software when used for training purposes.
IV. SOLUTIONS FOR THE EFFECTIVE USE OF INEXPENSIVE COLLABORATION SOFTWARE IN TRAINING

Even though economical collaboration technology may be instructionally imperfect, it may still be engineered to provide active student participation, engage deeper levels of thinking, and, in short, positively transform educational practices at low costs.

A. Overcoming technical limitations

Whenever possible, if using inexpensive collaboration software in training, the classroom even should be delivered via high-speed connections to ensure seamless voice and data transmission (preferably a corporate intranet or LAN) or, at a minimum, offered via DSL and high-speed cable. Fortunately, current technology is advancing and soon training providers will have access to increasingly sophisticated wireless connection schemes. Such capabilities will offer smoother video and better-synchronized audio over digital phone lines and LANs.

The following recommendations revolve around sound instructional design and assume that no video is used when hosting an online course using inexpensive collaboration software.

B. Instructional improvements

Superior technology is not the only ingredient in a robust instructional experience. Technology needs to be balanced by solid instructional design theory and principles and it needs to match the instructional goal that a class is set to accomplish.

To overcome instructional limitations of low-cost collaboration software, both instructional designers and instructors need to attend specialized training for learning how to create and deliver training delivered via such media. It is essential that instructors and instructional designers know how to best choreograph an entire classroom event using new technology, from figuring out the right proportion between lecturing, application sharing, to offering students ample 'question and answer' opportunities as well as chances to effectively use available interactive features.

Following are several suggestions for overcoming technical and instructional limitations of low-cost collaboration software. These suggestions assume that during class event students and instructors connect via a separate audio bridge (conference call) and no video conferencing is being used.

When using inexpensive collaboration software for providing synchronous training, the instruction should be divided into the following (Wells, 2004):

- Activities led by the instructor, which include clear visuals, brief presentations, prepared questions.
- Activities initiated by participants, which include questions and discussions.
- Activities practiced by the group, which include case studies, role-plays, and collaborative application of ideas to real job issues.

Each training segment provided via collaboration software should be kept relatively short (no more than 1-2 hours). Students grow weary of just watching the screen while listening to a "disembodied voice." In addition, participants learn and retain more when training is scheduled in small chunks rather than in day-long sessions (Wells, 2004). Keep students to no more than 15 per session.

Due to the fact that students and facilitators cannot "see" each other, emphasizing the relevance of the course materials to recipients becomes even more critical than in traditional instruction. Course relevance is inherent to instigating and sustaining student motivation. In an environment where students cannot get a visual of others, it becomes even more important to keep them motivated. The course design should contain frequent references as to how materials can be easily and immediately transferable to students' jobs or real life situations. Including student-suggested activities is also a great idea for maintaining their motivation and ensuring course relevance.

Instructors should clearly organize and streamline course discussions. In an electronic learning environment, students may become quickly overwhelmed by too much information. Clear organization of course materials eliminates confusion and builds students' confidence.

Classes delivered via collaboration software should provide structured activities (e.g., courses should provide guidelines for posting material, how often to comment, length of comments and what to say in them). This will avoid the situations when students may be stumped by online tasks, may lack Web expertise, misunderstand directions or are unsure what is expected of them.

To overcome the lack of quiz abilities in low-cost collaboration software, the course could point to independent online quizzes for practice and to final reviews that are developed via tools that enable a link to an LMC. This way, at the conclusion of a NetMeeting-based course, for instance, students may be asked to access a URL to a final review that has the ability to submit results to an LMC.

One of the reoccurring complaints from students when using inexpensive collaboration software for training is that peer camaraderie is lacking. Students
tend not to reach out to each other online as fully online as they do face-to-face. To overcome this complaint, teachers should assign online buddies and pair up students to help each other troubleshoot software problems and respond to questions about course content.

Another difficulty that stems from using inexpensive collaboration software is the inability to form "community of learners" online. Because students cannot see each other, it takes time for them to build trust and speak freely. Instructors should encourage students to interact casually and enable them to create discussion forums or areas for hanging out and hold personal introductions.

The course design should ensure that instructors cannot fall into lecture mode. Instructors should be required to ask students to initiate discussion topics and take turns in running discussion threads. They should also stop regularly during the presentation to ask if there are any questions since the presenter has no visual clues for judging whether students understand the content.

Instructors also need to work on their facilitation skills. Given the fact that, when using collaboration software, students cannot be seen most of the times, they have the tendency to ask more questions and comment on other participant’s suggestions (Jones, 2004). An instructor should be prepared to balance such interaction and fit it within the class schedule and flow. Instructors should also know that preparing for delivering synchronous online courseware may require 20-30% more time than preparing for a traditional class (Jones, 2004).

Inexpensive collaboration tools may often be based on shaky technology. This is why instructors should be prepared for technical errors. Students’ computers or Intranet connections may malfunction, or glitches may plague online discussion software. Instructors should check in regularly to see whether students need help using the discussion software or whether you need to call technology support personnel about more serious software problems. Instructors should also have a backup machine ready to deliver instruction in case of a computer crash.

V. REMARKS

If these suggestions are taken into consideration, using inexpensive collaboration tools for synchronous training may be a sound solution to fixing performance problems. Compared to standalone Web-based training for instance, a synchronous session is scheduled as part of a student’s day (thus guaranteeing commitment) and it also offers personal contact with peers and students. When used effectively, it can change attitudes, motivate mastery, and encourage more effective behavior on the job. IT can also ensure thoroughness of material coverage and spontaneity of ideas, which feeds creativity; Socratic questioning, considered one of the most effective teaching strategies in leader-led situations (Stamps, 2004).

REFERENCES

Abstract - The Electrocardiogram (ECG) is the most important biosignal used by cardiologists for diagnoses purposes, which provides key information about the electrical activity of the heart. Detection of abnormalities in ECG signal is a critical step in health care. This paper briefly introduces theory of wavelet transform and shows a few promising applications in ECG signal processing, as noise suppression, baseline wandering removal, ECG characteristic points detection. Wavelets provide efficient localization in both time and frequency (or scale) due to the multiresolution analysis (provided by the wavelet basis).

Key words: Wavelet Transform, multiresolution analysis, ECG signal processing, QRS detection

1. INTRODUCTION

The goal of wavelet research is to create a set of basis functions and transforms that can give an efficient and useful description of a signal (or function). Wavelets provide efficient localization in both time and frequency (or scale). Wavelet based analysis of an ECG signal is an exciting new tool for both scientists and engineers. The discretized wavelet analysis fits naturally with the digital computer with its basis functions defined by summations not integrals or derivatives. Unlike most traditional expansion systems, the basis functions of the wavelet analysis are not solutions of differential equations. To analyze any finite energy signal \( f(t) \in L^2(\mathbb{R}) \), the continuous wavelet transform (CWT) uses the dilation and translation of a single wavelet function \( \psi(t) \) called mother wavelet [3]. The continuous wavelet transform \( (W_\psi f)(s, \tau) \) of the signal \( f(t) \in L^2(\mathbb{R}) \) [1] is defined as:

\[
(W_\psi f)(s, \tau) = \int_{-\infty}^{\infty} f(t) \frac{1}{\sqrt{s}} \psi\left(\frac{t-\tau}{s}\right) \, dt \tag{1}
\]

where, we have used \( \overline{\psi} \) to denote the complex conjugate of \( \psi \). The function \( \psi \in L^2(\mathbb{R}) \) is an oscillatory function with zero mean. This last condition allows for the inversion of the wavelet transform. In particular the function \( f(t) \in L^2(\mathbb{R}) \) can be recovered from its transform \( (W_\psi f)(s, \tau) \) by the inverse formula:

\[
f(t) = C_\psi^{-1} \int_{-\infty}^{\infty} \int_{-\infty}^{\infty} (W_\psi f)(s, \tau) \cdot \overline{\psi}\left(\frac{t-\tau}{s}\right) \, ds \, d\tau \tag{2}
\]

Since the scale factor \( s \) is proportional to the inverse of the frequency \( \omega \), the value \( (W_\psi f)(s_0, \tau_0) \) exhibits the frequency content of \( f(t) \) in a frequency interval centered around \( \omega_0 = \frac{1}{s_0} \) at the time interval centered around \( t = \tau_0 \). The continuous wavelet transform maps a signal of one independent variable \( t \) of two independent variables \( s, \tau \). The scale factor and/or the translation parameter can be discretized. The usual choice is to follow a dyadic grid \( (s_j, \tau_{j,k}) \) on the time-scale plane with \( s_j = 2^j \) and \( \tau_{j,k} = k \cdot 2^j \). The transform is then called the dyadic discrete wavelet transform:

\[
(W_\psi f)(2^j, 2^k) = < f(t), \psi_{j,k}(t) > \tag{3}
\]

where \(< \bullet, \bullet >\) denotes the inner product in \( L^2(\mathbb{R}) \).

The dyadic sampling is a very natural choice for computers. We can construct functions [11]:

\[
\psi_{j,k}(t) = \frac{1}{\sqrt{2^j}} \psi\left(\frac{t - k \cdot 2^j}{2^j}\right) \tag{4}
\]

to form an orthonormal basis for signal representations. The decomposition process can be iterated, with successive approximations being decomposed in turn, so that one signal is broken down into many lower resolution components [3]. For discrete-time signals, the dyadic discrete wavelet transform (DWT) is equivalent, according to Mallat’s algorithm [1] to an octave filter bank, and can be implemented as a cascade of identical cells (low-pas and high-pass finite impulse response (FIR) filters). These filters split the signal’s bandwidth to half. Using downsamplers after each filter, the redundancy of the signal representation can be removed. This is called the wavelet decomposition tree.

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II. THE MAIN ECG SIGNAL PARAMETERS

The electrocardiogram, is a time-varying signal that measures the electrical activity (on the surface of the human body) of the heart. Each heartbeat is a complex of distinct cardiological events, represented by distinct features in the ECG waveform. These features represent either depolarization (electrical discharging) or repolarization (electrical recharging) of the muscle cells in particular regions of the heart. Figure 2 shows a (human) ECG waveform and the associated parameters (features).

Fig. 2. The most important parameters of an ECG signal

The standard parameters of the ECG waveform are the P wave, the QRS complex and the T wave. But most of information lies around the R peak. Additionally a small U wave (with an uncertain origin) is occasionally present.

The cardiac cycle begins with the P wave, which corresponds to the period of atrial depolarization in the heart. This is followed by the QRS complex, which is usually the most relevant (recognizable) feature of an ECG waveform. The QRS complex corresponds to the period of ventricular depolarization. The start and end points of the QRS complex are referred to as Q and J (or very often S) points. The T wave follows the QRS complex and corresponds to the period of ventricular repolarization. The end point of the T wave represents the end of the cardiac cycle (presuming the absence of U wave). The durations (time between the onset and offset) of particular parameters of the ECG (referred as time interval) is of great importance since it provides a measure of the state of the heart and can show the presence of certain cardiological conditions. In practice, interval measurements, wave interpretations are carried out manually by ECG specialists.

III. MATERIALS AND METHODS

In all applications were used signals from MIT-BIH database with annotations from specialists (cardiologists), all methods were developed under Matlab (and Wavelet Toolbox). At first, the signals from database were filtered, denoised and after that baseline variation was removed, using wavelet approximation. The main idea of wavelet analysis is to find a function (a basis function) which properties are appropriate to the analysed signal, to obtain maximum of information with less coefficients. In this work, biorthogonal wavelet were used, because after many experiments they gave the best results in signal reconstruction from approximation coefficients [5,12]. These wavelets have a minimum number of sign changes which simplify the steps in the parameter estimation algorithm. The next figure presents how the wavelet analysis based feature extraction is carried out.

Fig. 3. Wavelet analysis based feature extraction

At first, in the preprocessing stage, the signals from database are filtered, denoised and baseline wander removed, using wavelet approximation. The basic denoising concept is to decompose the signal at different scales (the largest quantity of information about noise is contained by the first scales detail coefficients [4]), to threshold the detail coefficients and after that to reconstruct the signal, using the original approximation coefficients and the modified detail coefficients.

The first step in the baseline wandering removal is to identify the low frequency (large scale) components in the ECG signal. The typical baseline variation means 15 percent of peak-to-peak ECG amplitude variation of 0.15 to 0.3 Hz. This variation can be identified as the 8th level approximation for a signal sampled with 128 Hz.

Fig. 5 Wavelet decomposition, spectral thresholding for baseline wander removal

The proposed algorithm (baseline variation identified as the lowest frequency component) was experimented on test signals (weighted sum of sinusoidal functions). The lowest frequency components were identified and extracted from the original signal [6]. Figure 6 shows this procedure in the frequency domain.
This algorithm applied to real ECG signals gave good results. Results can be seen below:

The main advantage of this algorithm is, that it can be applied again to the already filtered signal.

In the processing stage, the main ECG signal parameters are identified, following the algorithm presented below:

Steps:
1. The segment of ECG signal considered for analysis consists of 60 s (meaning 7680, 128 Hz sample rate)
2. A 4 level wavelet decomposition performed, using biorthogonal wavelet functions (the reconstruction and decomposition filters are implemented as quadrature mirror filter of FIR type)
3. Determination of the R wave location (as local maxima) on first level approximation (first scale) Additionally, in this step the maxima can be identified as zero crossing points of the first derivative. An adaptive threshold is used (related to the maximum and mean values of the signal), to find the points over this value. After that, the R peaks are selected.
4. Determination of R-R intervals, as R-R distances
5. Determination of Q, S points as the first zero crossing or local minimum point before and after R wave. QRS complex’s area can be calculated from the Q-S duration and the value of the R peak.
6. Elimination of the QRS complex from the signal to obtain the other parameters
7. Determination of the P wave location (as maxima) (scales 3, 4) (the same procedure as in step 2), and the P-Q distance
8. Elimination of the P wave from the signal (same as 5)
9. Determination of the T wave location (as the remained maxima) (scales 3, 4) and S-T segments durations

IV. RESULTS

This algorithm leads us to determine the main parameters of an ECG signals. Were used over 27 files from the MIT-BIH database (free available on the Internet for algorithm test purposes), signals containing normal sinus rhythms and signals with abnormalities in order to find the main parameters. The results obtained (processing mainly ECG signals from normal sinus database) were compared with
annotated files from ECG databases, and gave very promising results: R wave detection around 98%, QRS complex detection over 95%, T wave detection/localization 88%, P wave detection/localization 88%.

<table>
<thead>
<tr>
<th>action</th>
<th>results</th>
</tr>
</thead>
<tbody>
<tr>
<td>R peak identification</td>
<td>98.2 %</td>
</tr>
<tr>
<td>QRS complex extraction</td>
<td>95.4 %</td>
</tr>
<tr>
<td>T wave identification</td>
<td>89.5 %</td>
</tr>
<tr>
<td>P wave identification</td>
<td>88.2 %</td>
</tr>
</tbody>
</table>

V. CONCLUSIONS

The present study is based on biorthogonal wavelets. It has been shown [2] that these wavelets are ideally suited for the purpose since they excite the various morphologies of the ECG signals at different scales. With the multiscale feature of WT’s, the QRS complex can be distinguished from P or T waves, noise, baseline drift, and artifacts [5]. Various morphologies are excited better at different scales. From these scales various segments, time widths as signal parameters can be determined more accurately. As a further work, an artificial neuronal network will be implemented (trained with a set of normal sinus beats) for ECG events analysis and abnormalities detection.

REFERENCES

Speech enhancement using Enriched Diversity Wavelet Transform and Bishrink filter

Cedric Clolus

Abstract - This paper presents a new denoising method for speech signals corrupted by additive noise. This method is based on the use of a special wavelet transform, called enriched diversity wavelet transform and of a special MAP filter, called composed bishrink. Some simulations are presented. The results obtained are compared with the results of other denoising methods.

Keywords: wavelets, denoising, soft thresholding, bishrink.

I. INTRODUCTION

A lot of applications in the domain of signal processing require voice activity detection: speech coding, echoes reduction in telephony domain for example. Some solutions for systems with a voice activity detector have been standardized, but these solutions lack of robustness. It is why other solutions for voice activity detection are searched. Those one try to integer in the structure of the voice activity detector methods of denoising speech signal. In this paper, we consider a method of denoising based on two new tools of signal processing, the enriched diversity wavelet transform and the Bishrink filter. Section II deals with some already existing methods of denoising speech signal. In Section III, the proposed method is presented. Section IV focuses on the results obtained by simulation of the preceding methods. The last section is dedicated to some concluding remarks.

II. DENOISING METHODS

In this section, we will focus on two main methods to denoise speech signals, the spectral subtraction and a method based on statistic properties of the signal.

1. Spectral Subtraction [1] is a method for restoration of the power spectrum or the magnitude spectrum of a signal observed in additive noise, through subtraction of an estimate of the average noise spectrum from the noisy signal spectrum. The noise spectrum is usually estimated from the periods when the signal is absent and only the noise is present. For restoration of time-domain signals, an estimate of the instantaneous magnitude spectrum is combined with the phase of the noisy signal, and then transformed via an inverse discrete Fourier transform of the time domain. In terms of computational complexity, spectral subtraction is relatively inexpensive. The block diagram of this method is presented by the figure 1.

2. The second method for speech enhancement on which we will focus is the system presented by Y.Ephraim and D.Malah [2][8] based on the utilization of a Minimum Mean-Square Error Short-Time Spectral Amplitude Estimator (MMSE STSA). The model used is a statistical model which utilizes asymptotic statistical properties of the Fourier expansion coefficients and more specifically, the fact that the Fourier expansion coefficients of each process can be modeled as statistically independent Gaussian random variables. The mean of each coefficient is assumed to be zero, since the processes involved are assumed to be zero mean. The variance of each speech Fourier expansion coefficient is time-varying, due to speech nonstationarity. This Gaussian statistical model is motivated by the central limit theorem, as each Fourier expansion coefficient is a weighted sum (or integral) of random variables resulting from the process samples. The MMSE STSA estimator depends on the parameters of the statistical model it is based on. In the algorithm proposed in [2], these are the a priori SNR of each spectral component and the variance of each noise spectral component. The a priori SNR was found to be a key parameter of the STSA estimator. Their method consists in combining a “decision-directed” method for estimating the a priori SNR with the MMSE STSA estimator which takes into account the uncertainty of signal presence in the noisy observations.

Fig.1. Block diagram of a spectral subtraction system
III. SPEECH ENHANCEMENT USING ENRICHED DIVERSITY WAVELET TRANSFORM AND BISHRINK FILTER

In recent years, the techniques that use multiscale and local transform-based algorithms have become popular in noise filtering applications. In particular, the use of non-linear filters in the DCT domain was studied. In this section, we propose to present a denoising method based on the enriched diversity wavelet transform (EDWT) [5] and on the Bishrink filter [6] [7]. The block diagram of this method is presented in figure 2.

For a given signal, in using different mother wavelets, different concentrations of energy are obtained. Thus, for a given input signal, there is a mother wavelet which maximizes the concentration of energy. The idea of the enriched diversity wavelet transform, proposed in [5], is obtaining a discrete wavelet transform which is less sensitive to the choice of the mother wavelet. The construction of the wavelet transform is based on the increase of the diversity. The parameters of the discrete wavelet transform are the mother wavelet and the number of iterations. Thus, the diversity can be increased in calculating for a same signal, \( x(t) \), several discrete wavelet transforms. For each one, a different mother wavelet is used. So, the enriched diversity wavelet transform is obtained. It is a redundant discrete wavelet transform. This transform carries out association between the vector \( x(t) \) and a matrix \( EDWT[t,.] \). Each column of this matrix represents one of the discrete wavelet transform of the signal \( x(t) \). This transform can be inversed. For each columns of the matrix \( EDWT[t,.] \), the corresponding inverse discrete wavelet transform is calculated. So, a new matrix is obtained. Each column of this matrix contains the signal \( x(t) \). In calculating the mean of the columns of this matrix, the vector \( x(t) \) is obtained. The filtering in the domain of the enriched diversity wavelet transform is obtained by filtering each component of the transform.

In [6], Sendur and Selesnick proposed a wavelet-based denoising filter which considers the dependencies between the wavelet coefficients and their parents in detail. In fact, there are strong dependencies between neighbor coefficients such as between a coefficient, its parent (adjacent coarser scale location), and their siblings (adjacent spatial locations). From this observation, Sendur and Selesnick built four jointly non-Gaussian models to characterize the dependency between a coefficient and its parents, and derived the corresponding bivariate MAP estimators based on noisy wavelet coefficients in detail. From each model, a denoising method is proposed in taking into account the dependencies between the wavelet coefficients and their parents in detail.

As said previously and in [7], there is an important correlation between a wavelet coefficient at a given scale and the same coefficient situated in the same position at the next scale. This correlation can be exploited to construct adaptive filters acting at a given scale and using for the estimation of their parameters information obtained at the next scale. Using the parent and child wavelet coefficient of the input signal it is possible to estimate the child coefficients of the discrete wavelet transform of the useful part of the input signal, with the aid of a bishrink filter.

Concerning the model of the DWT of the useful component, in the case of our method’s filter, named composed Bishrink filter, for the first iterations (for scales with a number of coefficients superior with 16), a Laplace distribution will be considered (like in the case of the bishrink filter). For the last iterations, a soft thresholding with a threshold equal to 3 \( \sigma_n \), where \( \sigma_n \) is the noise variance (the rule of the 3 sigma), is used.

IV. SIMULATION RESULTS

This section will present the simulation done to test the new method and the comparison done with two others methods. The simulation have been carried out with two different noises, first a Gaussian white noise and secondly with a F16 noise, what is to say a noise recorded at the co-pilot’s seat in a two-seat F-16, traveling at a speed of 500 knots, and at an altitude of 300-600 feet. These noises have been added to a clean speech signal (figure 3) with different Signal-to-Noise Ratios. The simulations have been carried out with four different denoising methods. The first two methods used for the simulation are the spectral subtraction proposed by Kamath [8] and the statistical method using a Minimum Mean-Square Error Log-Spectral Amplitude Estimator (MMSE log-STSA) proposed by Ephraim and Malah [9]. The denoising has been also carried out with the method proposed in section III (EDWT). The final method used for the simulation is derived of the precedent method. Indeed, after the denoising of the noisy signal by the enriched diversity wavelet transform, a wrap detection is done on the denoised signal in order to obtain a voice.
activity detection. The result is multiplied with the denoised signal to obtain the fourth result of the simulation (EDWT + VAD). The figures 4 and 5 represent the relation between the input SNR and the output SNR for the two different noises. So, with a signal corrupted by a Gaussian white noise, the proposed method obtained better results than the others, for any input SNR. However, for the F16 noise, the proposed method obtained best results only for an input SNR superior to 3 dB. Thus, the denoising by using Enriched Diversity Wavelet Transform and Bishrink filter is, in the most part of
cases, more efficient than the others methods, but for some particular noise and for low input SNR, it obtains less good result.

V. CONCLUSION

In this paper is proposed a new denoising method based on the use of several discrete wavelet transforms and of the bishrink filter in the wavelet domain. This method takes into account also the statistics of the useful part of the input signal. That makes that this method to perform better in most cases than the denoising method based on spectral subtraction and some denoising method based on the statistical properties of the signal. Although the Bishrink filter was originally envisaged for denoising signals corrupted by white noise, it allows having good results with all kind of noise, even if the best results remain obtained with signals corrupted by white noise. A problem of the proposed method is its computational complexity more important than the computational complexity of the other methods used for the test presented in this article. Finally, the results obtained during the simulation showed that the proposed methods for denoising can be easily integrated in a voice activity detection system and allow obtaining good results.

REFERENCES

Performance analysis of spatial diversity algorithms on an 802.11a PHY

Ligia Chira¹, Ruxandra Dumitrescu

Abstract – Simulations performed on an 802.11a PHY platform have proven the necessity of introducing some spatial diversity schemes on the transmission chain. Basic antenna diversity designs – ThC (Threshold Combining), SDC (Selection Diversity Combining), MRC (Maximum Ratio Combining) and EGC (Equal Gain Combining) – have been tested. They clearly improve bit rate and packet error rate performance, and EGC provides higher antenna diversity gain than SDC. MRC achieves the highest antenna diversity gain compared to the other techniques.

Keywords: spatial diversity, receiver diversity, channel conditions, diversity combiner.

I. INTRODUCTION

Our analysis was performed in the context of a study of adaptive radio techniques, in the attempt to improve capacity performance of 802.11a WLANs. We have established our focus on spatial diversity techniques, and this paper presents the results of some receiver diversity implementations.

Both LOS (Line-of-Sight) and NLOS (Non Line-of-Sight) propagation conditions were simulated, as well as flat fading and dispersive fading conditions.

II. 802.11a PHY PERFORMANCE

The first step consisted of sets of simulations on the 802.11a PHY Simulink platform [1], aiming to establish the limitations of this scheme in terms of received SNR, bit rate, packet error rate, and number of propagation paths. The main blocks are: a data source, the modulator bank, the OFDM (Orthogonal Frequency Division Multiplexing) transmit assembly, the channel, the OFDM reception assembly, a frequency domain equalizer, and the demodulator bank. The code employs link adaptation scheme wherein we select the best coding rate and modulation scheme based on channel conditions.

We have simulated both LOS and NLOS propagation by combining and adapting channel blocks available in the Simulink Communications Blockset: e.g. Rician Fading Channel and Multipath Rayleigh Fading Channel. We have varied the number of paths in the case of the Rayleigh channel, and modeled indoor and outdoor environments by changing the delay spread according to standard values [2], table 1.

Table 1. 802.11a standard values for indoor and outdoor environments [2]

<table>
<thead>
<tr>
<th></th>
<th>802.11a OFDM</th>
<th>Delay spread</th>
<th>Max Path Length Difference</th>
</tr>
</thead>
<tbody>
<tr>
<td>Outdoor</td>
<td>1us – 20us</td>
<td>300m – 6km</td>
<td></td>
</tr>
<tr>
<td>Indoor</td>
<td>40ns – 200ns</td>
<td>12m – 60m</td>
<td></td>
</tr>
</tbody>
</table>

Table 2 synthesizes the main limitations of 802.11a PHY operation, obtained from our simulations. As the number of propagation paths is increased the overall performance degrades. Spatial diversity techniques are efficient in fighting multipath propagation effects and we will show how this can be achieved in the particular case of the 802.11a PHY. Incorporating any kind of spatial diversity scheme into an 802.11a system will obtain a high-rate packet transmission system suitable for high throughput applications like videoconferencing and multimedia [3].

Future standards like 802.11n specify the use of multiple-antenna systems, MIMO (Multiple Input Multiple Output), to achieve high data rates (135 Mbps) and link reliability by using multipath to a benefit, unlike traditional systems. Until then, basic antenna diversity designs will help increase capacity and reliability for 802.11a systems.

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Table 2. 802.11a PHY operation limiting parameter values

<table>
<thead>
<tr>
<th>Parameters</th>
<th>Number of paths</th>
<th>2</th>
<th>3</th>
<th>4</th>
<th>5</th>
<th>&gt;6</th>
</tr>
</thead>
<tbody>
<tr>
<td>Rayleigh Channel</td>
<td>Dispersive fading</td>
<td>Out door</td>
<td>In door</td>
<td>In door</td>
<td>In door</td>
<td>In door</td>
</tr>
<tr>
<td>min SNR [dB]</td>
<td>11-15</td>
<td>10</td>
<td>17</td>
<td>23</td>
<td>35</td>
<td>Needs special solutions (e.g. spatial diversity)</td>
</tr>
<tr>
<td>max Bit rate [Mbps] for min SNR</td>
<td>6</td>
<td>12</td>
<td>18</td>
<td>18</td>
<td>36</td>
<td></td>
</tr>
<tr>
<td>max PER [%] for min SNR</td>
<td>10-12</td>
<td>12</td>
<td>18</td>
<td>14</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Flat fading</td>
<td>min SNR [dB]</td>
<td>8-12</td>
<td>7</td>
<td>15</td>
<td>20</td>
<td>33</td>
</tr>
<tr>
<td>max Bit rate [Mbps] for min SNR</td>
<td>6</td>
<td>6</td>
<td>6</td>
<td>12</td>
<td>36</td>
<td></td>
</tr>
<tr>
<td>max PER [%] for min SNR</td>
<td>12</td>
<td>10</td>
<td>4</td>
<td>18</td>
<td>12</td>
<td></td>
</tr>
</tbody>
</table>

III. RECEIVER DIVERSITY SCHEMES

Receiver diversity techniques are a means of implementing space diversity, and they involve the use of multiple antennas at the receiver. Our goal was to test and compare the performance of the 802.11a PHY with several receiver diversity configurations.

In receiver diversity, the independent fading paths associated with multiple receive antennas are combined to obtain a resultant signal that is then passed through a standard demodulator. Most combining techniques are linear: the output of the combiner is just a weighted sum of the different fading paths or branches [4]. Fig. 1 illustrates the 3-receive antenna case. We have considered dispersive fading conditions and one can notice that the SNR is estimated on each of the three branches.

There are several types of diversity combiners, with different implementation complexity and overall performance. We have tested four of them: Threshold Combining (ThC), Selection Diversity Combining (SDC), Maximum Ratio Combining (MRC), and Equal Gain Combining (EGC).

In Threshold Combining (ThC) only one receiver is needed. The received signals are scanned in a sequential order, and the first signal with a power level above a certain threshold is selected. This signal is used as long as its power is higher than the threshold value. When it falls below the threshold the selection process is reinitiated. With only two-branch diversity this is equivalent to switching to the other branch when the SNR on the active branch falls below $SNR_{th}$. This method is called switch and stay combining (SSC) [4]. Since the SSC does not select the branch with the highest SNR, its performance is between that of no diversity and ideal Selection Combining [4].

In pure Selection Diversity Combining (SDC), the received signals are continuously monitored so that the best signal can be selected. Since only one branch is used at a time, SDC often requires just one receiver that is switched into the active antenna branch. However, a dedicated receiver on each antenna branch may be needed for systems that transmit continuously in order to simultaneously and continuously monitor SNR on each branch [4]. The output of the combiner has an SNR equal to the maximum SNR of all the branches. SDC does not require co-phasing of multiple branches since only one branch output is used. To work properly each antenna branch must have relatively independent channel fading characteristics. To achieve this, the antennas are either spatially separated, use different polarization, or a combination of both [5].

In Maximum Ratio Combining (MRC) each signal is given a gain proportional to the ratio between the fading amplitude and the noise power. Since the signals are summed they must have the same phase to maximize performance. This requires not only separate receivers but also a co-phasing and summing device.

Equal Gain Combining (EGC) is a type of MRC. The weighting is equal, the weights are all set to the same value and are not changed after that. Then the signals are co-phased before the summation process just like in MRC.

IV. IMPLEMENTATION

Fig. 2 shows the threshold combing technique implemented with three receive antennas. All the received branches are scanned sequentially and the receiver outputs the first branch that has a SNR.
higher than the chosen threshold. The ThC block has a dialog window that allows the SNR threshold to be set by the user. The instantaneous SNRs of the received branches are obtained via the From blocks from the Multipath Channel block. The Constant block that can be seen in the figure below is the threshold SNR to which the SNRs of the three branches are compared. Only one of the If block’s conditions can be true at a given moment of time, and therefore only the signal of one of the receiver antennas is passed to the output port.

![Fig. 2. ThC scheme for a 1Tx-3Rx configuration](image)

Selection Combining takes the idea of threshold combining one step further by selecting the branch that has the highest SNR out of the received branches. This algorithm is implemented in Simulink by comparing the instantaneous SNRs of the incoming signals and outputting the signal from the branch with the highest SNR.

Similar to Threshold Combining, the conditions of the If blocks cannot be fulfilled simultaneously and therefore the signal from only one of the branches is passed to the output port at a given moment of time. The block diagram is too large to be presented here.

In the Equal Gain Combining technique the signals from the receive antennas are weighted by a gain that is preserved constant throughout the reception process. We have chosen the gain equal to 1/NRx, where NRx is the number of receive antennas. The resulting signals are summed and yield an output that is a combination of all the received branches.

![Fig. 3. EGC scheme for a 1Tx-3Rx configuration](image)

Maximum Ratio Combining technique proposes a means of combining the signals from all receiver branches, so that signals with a higher received power have a larger influence on the final output. The design depicted in Fig. 4 computes the gain of each branch according to the relation:

\[ a_i = \frac{SNR_i}{\sum_{i=1}^{N_{Rx}} SNR_i} \]  

where: \( a_i \) is the gain of branch \( i \); \( SNR_i \) is the instantaneous SNR of branch \( i \); \( N_{Rx} \) is the number of receive antennas.

![Fig. 4. MRC scheme for a 1Tx-3Rx configuration](image)

V. SIMULATION RESULTS

First we should mention our simulation settings. The symbol period was chosen 0.08us, the number of OFDM symbols per block – 20 symbols, the number of OFDM symbols in the training sequence - 4 symbols, low SNR thresholds: \{10 11 14 18 22 26 28\}, hysteresis factor for adaptive modulation - 2 dB, Viterbi traceback depth – 34. The channel fading mode was chosen dispersive fading and the maximum Doppler shift, 10 Hz. The receiver SNR threshold was set to 20 dB.

ThC as presented in the previous section, chooses from the received branches, the first one that has a SNR above a specified threshold. Its major drawback however is that some other branches may have an even better SNR than the chosen branch, and, still, they are suppressed. To mitigate this disadvantage, an optimised threshold has to be found. On the other hand, ThC technique has the advantage that, should a branch have a very low SNR, it does not influence the output of the receiver, as only one branch at a time is selected.

Table 3 shows the results of simulations carried out by tuning the average received signal to noise ratios SNR1, SNR2 and SNR3. Observe that these values represent the average SNR of the received signal, while the instantaneous SNRs that are involved in taking the decision on the branch to be selected at a given moment of time are displayed underneath the Multipath Channel’s block mask. These different
notions can be observed at the result of the simulation with the settings SNR$_1$=18 dB, SNR$_2$=25 dB and SNR$_3=30$dB, where we have an estimated SNR of 17.55 dB and a PER of 62 %. These poor results occurred because the ThC receiver has chosen the first branch with the instantaneous SNR exceeding the threshold value of 20 dB.

Also, the settings SNR$_1=5$ dB, SNR$_2=5$ dB and SNR$_3=30$dB do not illustrate the advantage that the poor SNR of some paths do not affect the output of the receiver. The artificially high values are chosen for the sake of exemplification and are unlikely to be encountered in real life transmissions.

Table 3. Threshold Combining Simulations for a 1Tx-3Rx Configuration

<table>
<thead>
<tr>
<th>SNR$_1$ [dB]</th>
<th>SNR$_2$ [dB]</th>
<th>SNR$_3$ [dB]</th>
<th>PER$_{inst}$ [%]</th>
<th>SNR$_{est}$ [dB]</th>
<th>Data rate [Mbps]</th>
</tr>
</thead>
<tbody>
<tr>
<td>25</td>
<td>30</td>
<td>15</td>
<td>26</td>
<td>14.69</td>
<td>18</td>
</tr>
<tr>
<td>35</td>
<td>25</td>
<td>20</td>
<td>22</td>
<td>32.63</td>
<td>54</td>
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<tr>
<td>20</td>
<td>30</td>
<td>24</td>
<td>54</td>
<td>17.55</td>
<td>24</td>
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<tr>
<td>15</td>
<td>20</td>
<td>30</td>
<td>16</td>
<td>21.19</td>
<td>18</td>
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<tr>
<td>18</td>
<td>30</td>
<td>25</td>
<td>62</td>
<td>17.55</td>
<td>24</td>
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<tr>
<td>5</td>
<td>5</td>
<td>30</td>
<td>16</td>
<td>24.24</td>
<td>24</td>
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<tr>
<td>30</td>
<td>25</td>
<td>10</td>
<td>26</td>
<td>21.01</td>
<td>36</td>
</tr>
</tbody>
</table>

A SDC receiver compares the received branches among each other and outputs the one with the highest SNR. Compared to the ThC technique, it has the advantage that it selects the actual best signal in terms of received power, and not merely the first one that meets a certain criteria. Besides, it also maintains the property that branches with poor SNR do not influence the output of the receiver.

Table 4 provides proof of the latter statements. It can be noticed that, under the same simulation settings, the SDC receiver has provided much better SNRs, and thus, better data rates than the ThC receiver.

Table 4. Selection Combining Simulations for a 1Tx-3Rx Configuration

<table>
<thead>
<tr>
<th>SNR$_1$ [dB]</th>
<th>SNR$_2$ [dB]</th>
<th>SNR$_3$ [dB]</th>
<th>PER$_{inst}$ [%]</th>
<th>SNR$_{est}$ [dB]</th>
<th>Data rate [Mbps]</th>
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<tbody>
<tr>
<td>25</td>
<td>30</td>
<td>15</td>
<td>32</td>
<td>20.17</td>
<td>36</td>
</tr>
<tr>
<td>35</td>
<td>25</td>
<td>20</td>
<td>24</td>
<td>20.15</td>
<td>18</td>
</tr>
<tr>
<td>20</td>
<td>30</td>
<td>24</td>
<td>12</td>
<td>24.46</td>
<td>48</td>
</tr>
<tr>
<td>15</td>
<td>20</td>
<td>30</td>
<td>36</td>
<td>22.79</td>
<td>36</td>
</tr>
<tr>
<td>18</td>
<td>30</td>
<td>25</td>
<td>24</td>
<td>24.26</td>
<td>24</td>
</tr>
<tr>
<td>5</td>
<td>5</td>
<td>30</td>
<td>8</td>
<td>34.62</td>
<td>54</td>
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<tr>
<td>25</td>
<td>30</td>
<td>10</td>
<td>16</td>
<td>22.94</td>
<td>36</td>
</tr>
</tbody>
</table>

The average SNR gain increases with the number of receiver antennas, but not linearly. The largest gain is obtained by going from no diversity to two-branch diversity. Increasing the number of diversity branches from two to three will give much less gain than going from one to two, and in general increasing the number of receivers yields diminishing returns in terms of the SNR gain [4].

A major disadvantage of both ThC and SDC is that they only switch to another branch after fading has occurred.

EGC receiver outputs a signal that is a weighted sum of all the received branches. The gain of each branch is maintained constant throughout the reception. This method is designed on the premises that it is unlikely that the SNR of one branch would be significantly lower than the SNRs of the other branches. It is obvious that, as the signals from all receive antennas are combined to yield the output of the ECG receiver, a significant fading that affects one of the branches would diminish the quality of the output of the receiver. Such an unlikely situation is still taken into consideration in the simulations performed (Table 5).

Table 5. Equal Gain Combining simulations for a 1Tx-3Rx Configuration

<table>
<thead>
<tr>
<th>SNR$_1$ [dB]</th>
<th>SNR$_2$ [dB]</th>
<th>SNR$_3$ [dB]</th>
<th>PER$_{inst}$ [%]</th>
<th>SNR$_{est}$ [dB]</th>
<th>Data rates [Mbps]</th>
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<tbody>
<tr>
<td>25</td>
<td>30</td>
<td>15</td>
<td>65</td>
<td>8.625</td>
<td>12</td>
</tr>
<tr>
<td>25</td>
<td>35</td>
<td>20</td>
<td>6</td>
<td>15.65</td>
<td>24</td>
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<tr>
<td>30</td>
<td>24</td>
<td>28</td>
<td>15.65</td>
<td>20.13</td>
<td>36</td>
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<tr>
<td>15</td>
<td>30</td>
<td>34</td>
<td>14.68</td>
<td>12</td>
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<td>25</td>
<td>30</td>
<td>14</td>
<td>20.14</td>
<td>24</td>
<td></td>
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<tr>
<td>5</td>
<td>5</td>
<td>30</td>
<td>56</td>
<td>3.141</td>
<td>6</td>
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<tr>
<td>25</td>
<td>30</td>
<td>10</td>
<td>12</td>
<td>14.76</td>
<td>12</td>
</tr>
</tbody>
</table>

MRC receiver outputs a signal that is a weighted sum of all the received branches. But, unlike EGC, the gains of each branch are updated at the reception of every frame and they are proportional to the power of the received signal. Table 6 shows the results of the simulations performed on a 1Tx-3Rx configuration and by comparing them to those from Table 5 performed with the same simulation settings, with an EGC receiver, we can draw the conclusion that MRC receivers perform an improved combining of the incoming signals.

Table 6. Maximum Ratio Combining Simulations for a 1Tx-3Rx Configuration

<table>
<thead>
<tr>
<th>SNR$_1$ [dB]</th>
<th>SNR$_2$ [dB]</th>
<th>SNR$_3$ [dB]</th>
<th>PER$_{inst}$ [%]</th>
<th>SNR$_{est}$ [dB]</th>
<th>Data rates [Mbps]</th>
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<tr>
<td>25</td>
<td>30</td>
<td>15</td>
<td>26</td>
<td>10.53</td>
<td>12</td>
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<tr>
<td>25</td>
<td>35</td>
<td>20</td>
<td>22</td>
<td>18.22</td>
<td>18</td>
</tr>
<tr>
<td>30</td>
<td>30</td>
<td>24</td>
<td>31</td>
<td>15.78</td>
<td>12</td>
</tr>
<tr>
<td>15</td>
<td>20</td>
<td>30</td>
<td>18</td>
<td>18.22</td>
<td>24</td>
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<td>25</td>
<td>30</td>
<td>16</td>
<td>26.66</td>
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<td>5</td>
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<td>30</td>
<td>36</td>
<td>10.54</td>
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<tr>
<td>25</td>
<td>30</td>
<td>10</td>
<td>32</td>
<td>16.41</td>
<td>24</td>
</tr>
</tbody>
</table>

VI. CONCLUSIONS

A major drawback of ThC is that some other branches may have an even better SNR than the chosen branch, and, still, they are suppressed. To mitigate this disadvantage, an optimised threshold has
to be found. On the other hand, ThC technique has the advantage that, should a branch have a very low SNR, it does not influence the output of the receiver, as only one branch at a time is selected.

SDC has the advantage that it selects the actual best signal in terms of received power, and not merely the first one that meets a certain criteria. Besides, it also maintains the property that branches with poor SNR do not influence the output of the receiver.

A major disadvantage of both ThC and SDC is that they only switch to another branch after fading has occurred.

EGC provides higher antenna diversity gain then SDC. MRC realizes the highest antenna diversity gain compared to the other techniques.

The performance of the diversity combiners increases with the number of antennas, but not linearly, and will eventually stop growing beyond a certain number of antennas.

Until the 802.11n standard is ratified (at the earliest 2006) 802.11a products combined with spatial diversity techniques will offer a series of so-called “pre-n” products [6].

REFERENCES

Comb Generator for Measurement Receiver Test

Teodor Petriță¹, Adrian Mihăiuță²

Abstract – Quick testing of measurement receiver needs rich frequency spectrum signals, with each component known frequency and level. This can be done with comb generators having outputs in equal frequencies distances. They work using very sharp - and consequently very wide spectrum - pulses, distance between two harmonics being equal with repetition - or clock - rate of impulses.

This paper shows the performances of a portable comb generator working on this principle, used in verifying measurement receivers in 30MHz-2GHz areas, as well as experimentally obtained results.

1. INTRODUCTION

In the last years EMC extends its applicability area, leading to new standards and due these to new measurements requirements. Also EMC used equipments were modified in terms of performance and construction principle.

After their use, EMC equipments can be: measurement equipments (which are usually measuring the level of perturbations) and test equipments. First category means for instance measuring receivers, and the second consists of generators simulating all kinds of perturbations. This means generators like ESD (electrostatic discharge), burst generators or burst generators. Recently, new kinds of generators are appearing answering to standard's requests but allowing also checking of some specific measurement equipments.

Such a generator is the comb type, which generates a wide spectrum of equally distanced harmonics, so the output spectrums looks like the teethes of a comb - and here from derives the name.

Comb generators are used in wideband applications where lots of spectral components - correlated with a low frequency signal - are required. Having in mind their wide area of frequency and the fact that the level of harmonics is known, comb generators are being used more and more in EMC, especially in immunity precompliance tests. On the other side, they are usually portable and that makes them very useful in case of inter-site measurements comparison.

Advantages of comb generators can be itemized like this:
- They can be calibrated referring to known standards;
- They work from a few kHz to tens of GHz;
- Low influence of existing interferences in the measuring site;
- High immunity to other perturbations;
- Can generate conducted or radiated interferences;
- Lightweight, portable tools.

2. OPERATION PRINCIPLES

Comb generators are made either starting from a very short - very low width ratio -pulse, either with pronounced unlinear characteristics circuits. Originating pulse can be ready generated by a included pulse generator or shaped from a sine wave, which can come also from outside.

The comb spectrum is theoretically obtained from a scaled version of periodic Dirac distribution (see fig.1) of type:

$$\delta_f(t) = \sum_{n=-\infty}^{\infty} \delta(t - nT)$$

"A" value from figure 1 is not the impulse amplitude, but its area. It can be shown that the Fourier transform of Dirac distribution has a similar look:

$$F\{A\delta_f(t)\} = A\Omega\delta_{\Omega}(\omega) = A\Omega \sum_{k=-\infty}^{\infty} \delta(\omega - k\Omega),$$

where \( \Omega = 2\pi/T \).

Comb spectrum is shown in figure (2).
Obviously in practice we can't generate a Dirac pulse, which remains only of theoretical value, and thus we cannot obtain a comb spectrum. But it can be generated a small width factor rectangular signal with finite amplitude, as shown in fig.3.

Developing its exponential Fourier series we get:

\[ x(t) = \sum_{k=-\infty}^{\infty} a_k \exp(j k \Omega t), \quad \Omega = \frac{2\pi}{T} \]  

(3)

where:

\[ a_k = \frac{A \tau}{T} \sin\left(\frac{k \Omega T}{2}\right) \exp\left(-\frac{k \Omega T}{2}\right) \]

(4)

For graphical representation of spectrum we will use the harmonic form of associate Fourier series:

\[ x(t) = c_0 + \sum_{k=1}^{\infty} c_k \cos(k \Omega t + \varphi_k) \]  

(5)

The connection between identities (3) and (5) is:

\[ c_0 = a_0, \]

\[ |c_k| = 2|a_k|, k \geq 1 \]  

And

\[ \varphi_k = \arg(a_k) \]

where \( c_k \) are the coefficients of harmonic Fourier series.

The representation of the coefficients of the harmonic Fourier as a function of frequency is shown in fig.4. The continuous component \( c_0 \), was not represented, being useful in this case.

We can see that the spectrum contains harmonic components positioned at integer multiples of fundamental frequency and the spectrum is having more lobes which are zero at integer multiples of \( \pi/\tau \).

In order to obtain a closer signal to the comb spectrum we will use a part of the first lobe. In conclusion, in order to have more harmonics in the first lobe, it is necessary for the pulse train to have a big period \( T \). On the other side, in order to have a wide bandwidth, \( \tau \) has to be as short as possible or the width factor \( \tau/T \) to be as short as possible. If the amplitude of pulses will remain constant, the energy of a pulse will be lower by lowering the width factor and thus a low width factor will impose high amplitude of pulses; this is increasing design complexity.

The frequency bandwidth of a comb signal is selected by an application that imposes a maximum delta between the amplitude of useful harmonics of the first lobe. Usually are accepted the harmonics bigger than \( 1/\sqrt{2} \) of the maximum of the main lobe envelope. The equation \( (\sin x)/x = 1/\sqrt{2} \) is having the root \( x_0 = 1.3916 \). The bandwidth will be then between 0 and \( x_0/\pi \).
3. OPERATION OF THE GENERATOR

The block schematic is shown in fig.5. It contains the power source SA that from portability reasons it is a battery. Since the battery voltage is not constant in time, there is a level detector DN that supervises that the voltage is in safe range - and that means that the comb it is in its intended parameters.

As long as the battery voltage is not decreasing under a certain level (approx. 6V) DN is driving the validating gate PV with a logical "1" and the signal generated by the quartz oscillator OSC is passing further to the other blocks.

The pulse shaper FI is making a short pulse, rich in harmonics. Further more, the distortion block BD is taking the harmonics of the pulse to the maximum of BFR96 capabilities. Resulting signal is amplified by the amplifier A and delivered to the output.

The electrical schematic is given in fig.6. The oscillator is realized with IC1A and IC2A gates of a F-series 7400 driven by Q1 20MHz quartz. The oscillator gives signal into IC1C gate, which is the validating gate. The validating signal is coming from the battery level control block. There is a 7805 IC that stabilizes the supply voltage of comb generator. The level detector is an Schmitt trigger made of T1 and T2 transistors. When battery voltage falls beyond 6.5 V it is locking the validating gate. The state of generator is signaled with two LEDS: LED1 on - means enough voltage; LED2 on means - not enough voltage and blocked output. The oscillator gives basically a 1/2 width factor. The oscillator establishes the step of the harmonic components of the spectrum. This signal it is not good enough for our application so it needs more harmonics. We will form a short pulse using the gates IC1C and IC1D. This block's operating principle is based on logic gate delay. The signal \( s_1(t) \) is going into the gate IC1C which outputs \( s_2(t) \). \( s_2(t) \) is going to one of IC1D inputs, \( s_1(t) \) being at the other input. Here comes an \( s_3(t) \) very short pulse with ns fronts. Further on, there is the distortion circuit of BFR96 with a RF diode. This stage is working in nonlinear area like a derivation. The diode is getting out faster the transistor from saturation. Last stage is also a BFR96. It is biased with R3 and amplifies the signal since the output power of the distortion stage is relatively small. The transistor is used to the maximum extent of its frequency bandwidth. (Cutoff frequency is about 5GHz). Its gain is about 10-15 dB, but the linearity is strongly affected upon 2GHz.
4. EXPERIMENTAL RESULTS

Here are the measurements made for this comb generator. There were used a spectrum analyzer Agilent E4406 VSA \( F_{\text{max}}=2.7 \text{ GHz} \) and a digital oscilloscope TDS3502 (with \( F_{\text{max}} = 500 \text{ MHz}, 5 \text{ Gs/s} \), was having an inferior bandwidth than our application since the useful spectrum shows usable harmonics over 2.5 GHz - at 2.5 GHz, there is 1 mV amplitude harmonic!).

Fig. 7 shows the wave shape of the obtained pulse at generator's output and also its Fourier spectrum made in Matlab. We can see a parasitical oscillation of obtained pulse on its negative crest. This happens due to the oscillating LC equivalent circuit that appears when Q2 is blocked.

The oscilloscope bandwidth spectrum is inferior to the one obtained from the spectrum analyzer due to the limited bandwidth of oscilloscope (Fig.8).

The conclusion is that the comb generator is producing a complex signal that has a rich spectrum of equally distanced harmonics of 20 MHz and it goes until 2GHz with useful values. The levels of spectral components in low frequency are about 0dBm (appreciatively 0,2 V), and it goes down to -26 dBm (about 1 mV), at 2 GHz (100 harmonic component).

The comb was made under contract INFRAS nr. 247/2004.

Bibliography

Algorithm for Frequent Pattern Recognition in Telecommunication Alarm Logs

Petru Serafin

Abstract – In telecommunication networks all the perturbations that influence the quality of telephony services must be presented to the network monitoring system by proper means that are generically called alarms. Alarms are registered in alarm logs. This paper presents a study of data mining over alarm logs in order to determine sequences of alarms that repeat themselves with a certain frequency. Such sequences of alarms constitute frequent patterns and may be of a certain interest for network monitoring systems.

Keywords: Alarm logs, pattern recognition, candidate patterns, frequent patterns.

I. INTRODUCTION

In the actual context of the development of telecommunications, the volume of information that is transported in telecommunication networks is continually increasing. Therefore, an important matter for the network monitoring system [2] is to be able to process the information with real-time constraints in order to determine the optimal functioning conditions for the network elements. Network monitoring systems are based on data acquisition of information provided within the network.

Generally, the information of notification about functional states of network elements at a given moment is called alarms. The information flow of all these alarms for the entire network is registered in alarm logs.

The architecture of most network monitoring systems is based on a modular organization. The data acquisition process consists of collecting alarms into alarm logs. The main objective of the network monitoring system is to guarantee and increase the quality of telephony services. It is important to analyze alarm logs to determine eventual faults in the supervised system.

The analysis of alarm logs [1], [4], [7], [8] aims to diagnose the functioning state of the network elements to be able to provide data for the expert system to make decisions for operating and maintenance of the telecommunication network. The important flow of alarms transmitted to the network monitoring system reveals the problem of alarm processing. Alarm processing eventually correlates alarms into relevant categories and tries to eliminate non-relevant alarms that do not influence the quality of services.

One of the methods of alarm processing is to find sequences of alarms that repeat frequently in a telecommunication alarm log. These sequences of alarms are called frequent patterns and they may be of some importance for the network monitoring system since they express a correlation between alarms that the system has to further analyze.

II. ALARM LOGS

An alarm log is defined as a list of alarm that is ordered chronologically following their moments of appearance in the network. An alarm log contains registered information about the functional state of network elements, for a given time interval. This time interval is called the observation window for the alarm log.

For example, in Figure 1 it is presented an alarm log for an observation window of 15 minutes about two network elements, UNIT1 and UNIT2:

The mathematical expression of an alarm is a couple of elements (x',t0), where x' is the type of

Figure 1. Alarm log example

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the alarm referring to network element \( i \), and \( t_0 \) is the moment of appearance of the alarm in the network [3]. Using this notation, the alarm log represented in Figure 1 can be expressed using a set of alarms that concern \( \text{UNIT1} \), \( a^i \) for \text{IN SERVICE}, \( b^i \) for \text{OVERLOAD} and \( c^i \) for \text{OUT OF SERVICE}.

In the same manner, concerning network element \( \text{UNIT2} \), it can be expressed using the alarm set \( a^2, b^2 \) and \( c^2 \).

With the notations introduced above, the alarm log in Figure 1 can be expressed mathematically as follows:

\[
J = \{ a^1, b^1, c^1, a^2, b^2, c^2 \}.
\]

An important observation is that the alarms which compose the log in equation (1) are distinct one to another, because each alarm has its own type and moment of appearance in the network. Another observation is that the alarms referring to network element \( \text{UNIT1} \) are repetitive because they can be found more than one time in the alarm log. Therefore from the given alarm log we may be able to extract two sub-logs, each referring to a different network element.

For network element \( \text{UNIT1} \), the mathematical expression of the given alarm log is the following:

\[
J'_{\text{UNIT1}(1)} = \{ a^1, b^1, c^1, b^1, a^1, c^1, a^1, b^1, c^1, b^1, a^1, c^1, a^1, b^1, c^1, b^1, a^1, c^1 \}.
\]

For network element \( \text{UNIT2} \), the alarm log can be expressed as follows:

\[
J'_{\text{UNIT2}(1)} = \{ a^2, b^2, c^2, a^2, b^2, c^2 \}.
\]

In Figure 2 it is depicted the graphical implementation of the sub-log in relation (2) for network element \( \text{UNIT1} \):

![Figure 2: Moments in an alarm log](image)

An important remark about equations (1) and (2) is that they may express some simultaneous alarms. For example \( a^1, b^1, c^1 \) and \( a^2, b^2, c^2 \). In reality these alarms probably do not appear at the same time in the network but with very short delays between them. But because of the discrete timing of moments of appearance, these short delays seem as simultaneous moments. In order to keep the characteristic of simultaneous alarms in equation (2), but still not indicate the exact appearance moments, we can write the equivalent equation (4):

\[
J = \{ a^1, b^1, c^1, a^2, b^2, c^2 \}
\]

Equation (4) is also referred as expressing an alarm log without indicating the moments of appearance of alarms, thus meaning it does not contain temporal constraints, it contains only the ordering constraints.

### III. FREQUENT PATTERNS

The main idea to analyze alarm logs to find sequences of alarms that repeat themselves with a certain frequency, is to generate some possible sequences of alarm, that are called candidate patterns, and to retain only those patterns that are frequent. This means that a calculation has to be done to determine if a candidate pattern is frequent. This calculation can be done considering the moments of appearance of each alarm in the log at the given time of the calculation or considering the moments of appearance by sub-logs.

The algorithm for frequent pattern recognition hereby presented is based on a fundamental propriety of frequent patterns (see [2]): if a pattern in an alarm log is frequent then all its sub-patterns are necessary to be also frequent in that alarm log. Another expression of this theorem is that the necessary and sufficient condition for a pattern to not be able to be frequent is that at least one of its sub-patterns is not frequent. Of course, nothing can be implied for a pattern with all its sub-patterns frequent: even if all sub-patterns are frequent we can only assume that the pattern may be frequent, not that it should be frequent.

For example, considering pattern \( abc \) and all its sub-patterns \( a, b, c, ab, bc \) and \( ac \), one can clearly determine that if any of these sub-patterns is not frequent then surely the pattern \( abc \) itself is not frequent. On the other hand, even if all sub-patterns are frequent, one can only determine that pattern \( abc \) may be frequent. So, considering \( ab \) is frequent, we determine that \( a \) and \( b \) are frequent. The algorithm needs to check the frequency of \( bc \) and \( ac \), to be able to retain \( abc \) as a possible frequent pattern.

### IV. THE ALGORITHM

The algorithm can be written as a logical diagram as presented in Figure 3. At each iteration \( i \) of the algorithm, there are generated sub-patterns of dimension \( i+1 \) starting from the frequent sub-patterns of inferior dimension. The frequency calculation eliminates sub-patterns that are not frequent and therefore eliminates the whole branch of
patterns of superior order that can be deduced starting from these sub-patterns.

An important parameter of the algorithm is the minimal considered frequency $f_{\min}$ that allows more or less complexity of the construction of candidate sub-patterns.

The algorithm ends at iteration $i$ for which there are no more frequent patterns calculated. Of course, the maximal dimension of the solution is given by the maximal dimension of the sub-log $J$ that is being analyzed, $n \leq \text{dim}(J)$.

Given a sub-log of alarms $J$ and a process of pattern recognition we can note $\overline{A}(J)$ the set of solutions of dimension $i$ that are found in sub-log $J$. Evidently the solutions depend on the strategy of sub-patterns generation and frequency calculation that are chosen. Let us suppose the strategy means that any two sub-patterns are distinct one to another.

The algorithm may be written in programming pseudo-language as follows:

```plaintext
/* Calculate minimum frequencies and retain in list only frequent sub-patterns of dimension $n+1$ */
$A_{n+1}(J) =$ \begin{align*}
\bigcup_{a_i} & a_i \in A_{n+1}(J), f(a_i) \geq f_{\min} \bigg) \end{align*};
/* Increment dimension $n$ */
n \leftarrow n + 1;
/* End of sequential do while */
End Do;
/* Construct list of results as assembly of all frequent sub-patterns from dimension 1 to $n-1$ form $J$ log */
Return $A(J) \leftarrow \bigcup_{i = 1}^{n-1} A_i(J)$;
```

Because we need to avoid the problem of infinite multiplicity of superior dimension candidate patterns, we must allow a predefined order between the alarms. For example we may consider the space of parallel alarms, which means candidate patterns are generated from previous dimension patterns by adding alarms of superior or equal order in the pattern. For example, the pattern $ab$ generates candidate patterns $abb$ and $abc$, but we need no longer to generate candidate $aab$ because this pattern contains sub-pattern $aa$ which is not frequent and was not retained.

The procedure for determination of frequent patterns in parallel space can be written in programming pseudo-language as follows:

```plaintext
// Procedure candidate_parallel() Generates parallel candidate patterns
/* Generates candidate pattern using a parallel-style space for alarm assembly */
In $\overline{A}_n(J)$ assembly of frequent patterns of dimension $n$
Out $A_{n+1}(J)$ assembly of candidate patterns of dimension $n+1$
/* Initialization of candidate patterns of dimension $n+1$ */
$A_{n+1}(J) =$ NULL;
For $\forall a_i a_{i1} \ldots a_{ik} \in \overline{A}_n(J)$
For $\forall a_j \in J$ AND $a_j \geq a_i, \forall k \in [1..n]$
/* Adding alarms of superior or equal type */
Candidate-parallel-pattern = $a_i a_{i1} \ldots a_{ik}$;  
/* Verify sub-patterns of the candidate pattern */
If all sub-patterns $a_i a_{i2} \ldots a_{ik} \in \overline{A}_n(J)$
then $A_{n+1}(J) =$ $A_{n+1}(J)$ $\cup a_i a_{i2} \ldots a_{ik}$;
End If;
End For;
End For;
/* Solution in candidate patterns of dimension $n+1$ */
Return $A_{n+1}(J)$;
```

For the determination of candidate patterns in the serial space alarms do not respect a certain order. New dimension candidate patterns are obtained just by adding alarms at the end of the assembly of previous patterns.
The procedure for determination of frequent patterns in parallel space can be written in programming pseudo-language as follows:

```plaintext
Procedure candidate serial() Generates serial candidate patterns /* Generates candidate patterns using a serial-type assembly relation for alarms */
In \( \mathcal{A}_n(J) \) assembly of frequent patterns of dimension \( n \)
Out \( \mathcal{A}_{n+1}(J) \) assembly of candidate patterns of order \( n+1 \)
* Initialization of assembly of candidate patterns of order \( n+1 \) */
\( \mathcal{A}_{n+1}(J) = \text{NULL}; \)
For \( \forall a_1 a_2 \ldots a_j \in \mathcal{A}_n(J) \)
For \( \forall a_j \in J \)
/* Adding alarms at the end of the assembly of candidate pattern */
Candidate-serial-pattern = \( a_1 a_2 a_j \); /* Verify sub-patterns for the candidate serial pattern */
If all sub-patterns \( a_1 a_2 a_j \) \( \in \mathcal{A}_n(J) \)
then \( \mathcal{A}_{n+1}(J) = \mathcal{A}_{n+1}(J) \cup a_1 a_2 a_j ; \)
End If;
End For;
End For;
/* Solution in candidate patterns of dimension \( n+1 \) */
Return \( \mathcal{A}_{n+1}(J) \);
```

IV. PARTIAL RESULTS

We can graphically represent the solutions for the frequent pattern recognition applied to alarm log from equation (1) in parallel-style assembly as follows:

Figure 4. Frequent patterns in parallel-style assembly

Also, in serial-type assembly the frequent patterns for the same alarm log in equation 1 can be expressed as follows:

Figure 4. Frequent patterns in serial-style assembly

The software implementation of the frequent pattern recognition algorithm was realized in the development environment OMNeT++ [6] as presented in Figure 5:

Figure 5. Module implementation

Some partial experimental results obtained in testing the algorithm over an alarm log of 10,000 alarms can be viewed in Table 1:

<table>
<thead>
<tr>
<th>Minimal freq.</th>
<th>Frequent alarms</th>
<th>Candidate patterns</th>
<th>Frequent patterns</th>
</tr>
</thead>
<tbody>
<tr>
<td>50</td>
<td>56</td>
<td>366</td>
<td>152</td>
</tr>
<tr>
<td>75</td>
<td>43</td>
<td>135</td>
<td>95</td>
</tr>
<tr>
<td>100</td>
<td>24</td>
<td>83</td>
<td>38</td>
</tr>
<tr>
<td>125</td>
<td>15</td>
<td>72</td>
<td>26</td>
</tr>
<tr>
<td>150</td>
<td>12</td>
<td>69</td>
<td>23</td>
</tr>
<tr>
<td>200</td>
<td>10</td>
<td>66</td>
<td>19</td>
</tr>
</tbody>
</table>

Table 1. Partial experimental results

For further extension of the algorithm by introducing Petri Nets formalism see [5].

IV. REFERENCES

Instructions for authors at the Symposium of Electronics and Telecommunications ETc 2004

Gheorghe I. Popescu

Abstract – These instructions present a model for editing the papers accepted at the Symposium of Electronics and Telecommunications, in view of publication in the "Buletinul Științific al Facultății de Electronici și Telecomunicații" ("Transactions on Electronics and Communications"). The abstract should contain the description of the problem, methods, solutions and results in a maximum of 12 lines. No references are allowed here.

Keywords: editing, Symposium, author

I. INTRODUCTION

The page format is A4. The articles must be of 6 pages or less, tables and figures included.

II. GUIDELINES

The paper should be sent in this standard form. Use a good quality printer, and print on a single face of the sheet. Use a double column format with 0.5 cm in between columns, on an A4, portrait oriented, standard size. The top and bottom margins should be of 2.28 cm, and the left and right margins of 2.54 cm. Microsoft Word™ for Windows is recommended as a text editor. Choose Times New Roman fonts, and single spaced lines. Font sizes should be: 18 pt bold for the paper title, 12 pt for the author(s), 9 pt bold for the abstract and keywords, 10 pt bold for the section titles, 10 pt italic for the subsection titles; distance between section numbers and titles should be of 0.25 cm; use 10 pt for the normal text, 8 pt for affiliation, footnotes, figure captions, and references.

III. FIGURES AND TABLES

Figures should be centered, and tables should be left aligned, and should be placed after the first reference in the text. Use abbreviations such as “Fig.1.” even at the beginning of the sentence. Leave an empty line before and after equations. Equation numbering should be simple: (1), (2), (3) … and right aligned:

\[ x(t) = \int_{-\infty}^{t} y(\tau - t)d\tau . \] (16)

IV. ABOUT REFERENCES

References should be numbered in a simple form [1], [2], [3]…, and quoted accordingly [1]. References are not allowed in footnotes. It is recommended to mention all authors; “et al.” should be used only for more than 6 authors.

<table>
<thead>
<tr>
<th>Table 1</th>
</tr>
</thead>
<tbody>
<tr>
<td>Parameter</td>
</tr>
<tr>
<td>I</td>
</tr>
<tr>
<td>U</td>
</tr>
</tbody>
</table>

V. REMARKS

A. Abbreviations and acronyms

Abbreviations and acronyms should be explained when they appear for the first time in the text. Abbreviations such as IEEE, IEE, SI, MKS, CGS, ac, dc and rms need no further explanation. It is recommended not to use abbreviations in section or subsection titles.
B. Further recommendations

The International System of units is recommended. Do not mix SI and CGS. Preliminary, experimental results are not accepted. Roman section numbering is optional.

REFERENCES

Instrucțiuni de redactare a articolelor pentru Buletinul științific al Facultății de Electronică și Telecomunicații

Gheorghe I. Popescu


I. INTRODUCERE


II. INSTRUCȚIUNI DE REDACTARE

Articolul trebuie să fie transmis în forma standard descrisă în acest material. Tipărirea se va face cu o imprimantă de bună calitate pe o singură față a paginii. Textul se va plasa pe două coloane de 8 cm cu spațiu de 0,5 cm între ele. Pagina A4 orientată pe înălțimea paginii are marginile de sus și jos de 1,78 cm, iar cele din stânga și dreapta de 2,54 cm. Prima pagină a articolului va avea margină superioară de 5 cm. Pentru editarea articolului se recomandă utilizarea procesorului de text Microsoft Word for Windows cu caractere Times New Roman dactilografate la un rând. Dimensiunile și stilul caracterelor sunt: Titlul articolului 18 pt îngrasat, autorul 12 pt rezumatul 9 pt îngrasat, cuvintele cheie 9 pt îngrasat, titlu paragraf 10 pt majuscule, titlu subparagraf 10 pt italic, distanța de la numărul de ordine la titluri va fi de 0,25 cm, textul normal 10 pt, afilierea autorului 8 pt, notele de subsol 8 pt, legendele figurilor 8 pt și bibliografia 8 pt.

III. FIGURI ȘI TABELE

Figurile și tabelele trebuie să fie inserate în text aliniate la stânga. Se recomandă evitarea plasării figurilor înainte de prima lor mențiune în text. Se va folosi abrevierea “Fig.1.” chiar și la începutul propozitiilor. Ecuatiile trebuie tipărite cu un rând gol deasupra și dedesubt. Numerotarea lor se face simplu în paranteze: (1), (2), (3) … Numerotarea va fi aliniată față de marginea dreaptă.

IV. REFERINȚE BIBLIOGRAFICE

Referințele bibliografice se numeroază consecutiv în forma [1], [2], [3]… Citările se fac simplu prin plasarea numărului corespunzător [5]. Nu sunt permise referințe bibliografice în notele de referință în subsol. Se recomandă scrierea tuturor autorilor și nu folosirea expresiei “și alții” decât dacă sunt peste 6 autori.

V. SFATURI UTILE

A. Abrevieri și acronime

Explicitați abrevierile și acronimele prima dată când ele apar în text. Abrevieri precum IEEE, IEE, SI, MKS, CGS, ac, de și rms se consideră cunoscute și nu mai trebuie explicat. Nu se recomandă utilizarea abrevierilor în titluri decât în cazul când sunt absolut inevitabile.

B. Alte recomandări

Se recomandă utilizarea unităților de măsură din sistemul internațional. Utilizarea unităților britanice poate fi făcută doar ca unități secundare (în paranteză). Se va evita combinarea unităților SI și CGS. Nu se admet rezultate experimentale preliminare. Numerotarea cu cifre romane a titlurilor de paragrafe este opțională. În cazul utilizării acestora se vor numerota paragrafelor proprie-ziște și nu “BIBLIOGRAFIA” sau “MULȚUMIRI”.

BIBLIOGRAFIE


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