Feasibility of Multi-Bandwidth Transmission in Future Broadband Wireless Systems

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Abstract—**This paper[1] explores the feasibility of multi-band transmission, that would be used in future generation broadband wireless systems that use cognitive radio principles of spectrum-sharing. Transmitted signals' spectra are split into several disjoint sub-bands, which can quickly be shifted to adapt to the current multi-user environment without causing interference. Transceivers can use efficient frequency domain digital signal processing and analog RF techniques to generate and receive multi-band signals and separate them from unwanted signals. Both OFDMA and serial modulation classes of signals are considered. For RF signal processing, the zero-IF architecture appears promising and feasible, given its lower parts count and complexity, and recent advances in fabrication technology. An issue was identified regarding possible constraints on the maximum OFDM block length imposed by phase noise. Two technologies for analog band-selection filters are considered – switched capacitor and switch resistor.**

I. Introduction

A significant contributor to overall spectrum efficiency in future generation wireless systems is the ability of different users and service providers to share spectrum with one another, on an as-needed basis. This spectrum-sharing will be made possible by the concept of "cognitive radio" [FCC03]; entities wishing to use part of a frequency band for transmission first scan to determine currently unused portions of the band, quickly establish temporary occupancy of suitable sub-bands and transmit on them, and finally vacate the sub-bands when the transmission is finished, or hop to another set of sub-bands if a higher-priority user requires the original set. This paper is concerned with transmitter and receiver digital signal processing (DSP) and radio frequency (RF) technologies which accommodate signals whose spectra are split into several disjoint *sub-bands*, and which can quickly change their transmitted spectra to adapt to the current multi-user environment without causing interference.

If the width of the common sharable spectrum block lies within the capabilities of the users' transceivers' analog to digital and digital to analog converters and digital signal processing, then most of the multi-band transceiver signal processing can be done digitally. In this case the transmitter and receiver signal processing architectures can advantageously be based on efficient fast Fourier transform operations. This is the main topic of Section 2. If the total extent of the available band or bands exceeds state of the art digital transceiver capabilities, analog processing at radio frequencies is necessary for generating and extracting user signals. Section 3 deals with the impact of multi-band scenarios on the radio front ends. Section 4 discusses the important topic of the transmitted power spectrum and nonlinear power amplifier effects on it. Section 5 contains summary and conclusions.

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II. Multi-Band Transmissions

Multi-band transmission allows users to split their signals into several sub-bands which can be transmitted in the vacant portions of the system band. In this section, we assume that the users' transmitters and receivers have RF front ends and ADC and DAC converters that can accommodate up to the entire common band. Data is transmitted in fixed-length blocks, separated by cyclic prefixes, which facilitate efficient transmitter and receiver frequency domain processing, using a generalized frequency domain approach [FK04]. Assume a discrete Fourier transform (DFT) block size of N_c, and that we have to transmit M_N ($< N_c$) data symbols using S separate, disjoint frequency "chunks", or <u>sub-bands</u>, labelled by i=1, 2, …S, as illustrated in Figure II-1. Initial spectral analysis of the frequency band would be used as a basis to decide on the number and allocation of the sub-bands. Receivers will include wideband DFT operations on incoming signal blocks.

The number of sub-bands necessary to accommodate a given block of data would depend on how crowded is the band with other users, as well as on the number of data symbols, M_N, to be transmitted in the block. We anticipate that the number of sub-bands will typically be small – say on the order of S=2 or 3. The total bandwidth occupied by the transmitted signal is equal to the bandwidth of the M_N sub-bands, but the span of frequencies, including gaps, may be larger than this, but less than N_c. For example, N_c = 2048, corresponding to a bandwidth of about 100 MHz, and an inter-frequency spacing of about 50 KHz. M_N could perhaps range from about 16 to 1024.

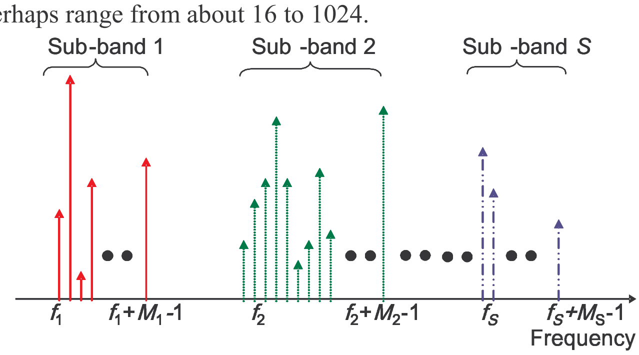

Figure II-1 Illustration of a multi-band line spectrum

For a general number S of sub-bands, the M_N data symbols are transmitted on a set of M_N discrete frequencies denoted by $\{f_1, f_1+1, ... f_1+M_1-1,\ f_2, f_2+1, ... f_2+M_2-1,$ $...\ f_S, f_S+1, ... f_S+M_S-1\}$, where the set $\Im_i = \{f_i, f_i+1, ... f_i+M_i-1\}$ forms the ith sub-band, and transports M_i data symbols, and

$$\sum_{i=1}^{S} M_i = M_N. \qquad \textbf{(II-1)}$$

Each block will be preceded by a cyclic prefix, and may also be preceded and followed by training symbol sequences, of length M_T symbols, where both the cyclic prefix and M_T exceed the maximum expected channel impulse response span, measured in symbol intervals.

A. Case of OFDMA (orthogonal frequency division multiple access)

The M_N data symbols are represented by $\{A(\ell)=a(\ell),\ \text{for}\ \ell=0,1,...M_N-1\}$. For OFDMA, the $\{A(\ell)\}$ are the coded data symbols themselves. This set of data is partitioned into S sets $\mathbb{A}_1 = \{A(\ell)\}_{\ell=0}^{M_1-1}$ … $\mathbb{A}_S = \{A(\ell)\}_{\ell=M_N-M_S}^{M_N-1}$. $\mathbb{A}_i$ is mapped onto frequency set $\mathbb{F}_i$. In the case of S>2, $\mathbb{A}_i$ would be mapped onto frequency set $\mathbb{F}_i$. Zeroes are mapped onto the remaining N_c-M_N frequencies constituting the block. The signal waveform is then generated from the N_c-point inverse DFT of the resulting block of data symbols and zeros where we have defined $M_0 = 0$. If more than one user's signal is to be frequency-multiplexed in the downlink, each user's signal would occupy and be generated in this way for a disjoint set of frequency sub-bands.

If the signal, with cyclic prefix, is considered as be periodic, with period N_c, it will have a line spectrum

$$S(f) = \begin{cases} A(f-f_i) & \text{for } f \in \Im_i \\ 0 & \text{otherwise.} \end{cases} \qquad \textbf{(II-2)}$$

The actual transmitted spectrum will be this line spectrum convolved with the $\sin(\pi f)/\sin(\pi f/N_c)$ spectrum of the rectangular time window of Nc samples. The resulting relatively high sidelobes can be suppressed by using a non-rectangular time window on the block plus its cyclic prefix, such as a raised cosine window.

For S sub-bands, the kth sample of the transmitted multi-band waveform is then

$$s(k) = \frac{1}{N_c}\sum_{i=1}^{S}\sum_{n=0}^{M_i-1} A(n+\sum_{i'=0}^{i-1} M_{i'})\exp(j\frac{2\pi k(f_i+n)}{N_c}),$$

$$\text{for } k=0,1,...N_c-1, \qquad \textbf{(II-3)}$$

At the receiver, after down-conversion and ADC conversion, the received block, not including the cyclic prefix, is processed by a N_c-point DFT, resulting in

$$R(f) = \begin{cases} A(f-f_i)H(f)+N(f) & \text{for } f \in \Im_i \\ N(f) & \text{otherwise.} \end{cases} \qquad \textbf{(II-4)}$$

where $H(f)$ is the channel's frequency response at frequency f, and $N(f)$ is the DFT of noise and other unwanted signals at frequency f. $R(f)$ is then sampled in the frequency sets $\{\mathbb{F}_i\}$; this results in M_N frequency domain samples, which are concatenated to form $\{Y(\ell),\ \text{for}\ \ell=0,1,...,M_N\}$, given by

$$Y(\ell) = A(\ell)H(\ell+f_i)+N(\ell+f_i) \ \text{for} \sum_{i'=0}^{i-1} M_{i'} \le \ell < \sum_{i'=0}^{i} M_{i'}$$

$$\text{and } i=1,...S \qquad \textbf{(II-5)}$$

These samples are equalized by the appropriate scalar multiplication of each frequency component, and are then passed on to the detector and decoder.

B. Case of serial modulation

For serial modulation, the data symbols are in the time domain, and so the frequency domain symbols $\{A(\ell),\ \text{for}\ \ell=0,1,...M_N-1\}$ are the M_N-point DFT coefficients of the data symbol sequence; i.e.

$$A(\ell) = \sum_{m=0}^{M_N-1} a(m)\exp(-j\frac{2\pi m\ell}{M_N}),\quad \ell=0,1,...M_N-1 \qquad \textbf{(II-6)}$$

for serial modulation where the $\{a(m)\}$ represent the data symbols. The other operations at the transmitter are identical to those described for OFDMA. The receiver operations are also the same, except that, after equalization, and before detection and decoding, a M_N-point inverse DFT operation is performed on the equalized frequency domain samples, to recover the time domain data. This approach, of transporting the DFT of a data symbol sequence on an arbitrary set of frequencies, has been called "carrier interferometry" in

[HNM+04].

It is easily shown by substituting the FFT expression (II-6) into (II-3), that the sampled waveform can be expressed as

$$s(k) = \frac{1}{N_c}\sum_{i=1}^{S}\exp(j\frac{2\pi k f_i}{N_c})\sum_{m=0}^{M_N-1} d_i(m) g_i(k-m\frac{N_c}{M_N}), \qquad \textbf{(II-7)}$$

where $d_i(m) = a(m)\exp(-j\frac{2\pi m}{M_N}\sum_{i'=0}^{i-1} M_{i'})$ for $m=0,1,2,...M_N-1,$

$$\textbf{(II-8)}$$

and

$$g_i(k) = \exp(j\frac{\pi(M_i-1)k}{N_c})\frac{\sin(\frac{\pi M_i k}{N_c})}{\sin(\frac{\pi k}{N_c})} \quad \text{for } k = 0,1,2,...N_c - 1.$$

(II-9)

For both OFDMA and serial modulation, essentially the same multi-band average power spectra will be produced. Figure II-2 illustrates the average power spectra for serial modulation and OFDMA for two sub-bands, with M_N=256 data symbols within a total bandwidth of N_c=2048. The other parameters for this figure are f_1=589, M_1=103, f_2=1390 and M_2=M_N – M_1 = 153. 5.5% rolloff raised cosine time windowing is used to suppress spectrum sidelobes.

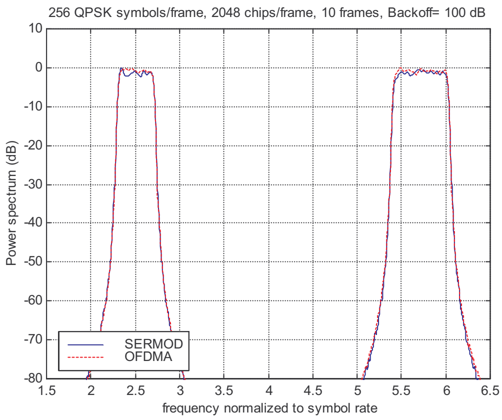

Figure II-2 Power spectra for S=2, N_c=2048, M_N=256 for OFDMA and serial modulation

III. Impact on Radio-Front end

A. Downconversion

One of the main constraints is the large bandwidth relative to the carrier frequency (4.5-6 GHz). This large bandwidth introduces constraints on the RF filters: components need to be linear over a very large range; rejection of aliases and interference becomes difficult with a single RF filter and a large percentage bandwidth (i.e. bandwidth normalized by the total system bandwidth). If the width of the total band to be decoded is higher than 100 MHz, several RF-front-ends with 100 MHz bandwidth each are recommended for implementation.

The widely used superheterodyne architecture has some drawbacks. The fact that there are two LOs, amplifiers and mixers, and complex filers for image and LO leakage rejection makes this structure complex, and thus costly and more power consuming, especially for signals whose bandwidth varies from zero to 100 MHz within a 4.5-6 GHz band plan.

Another option for the front-end is the Zero IF frequency or direct conversion architecture, requiring only one LO. The LO leakage and image frequency problems are eliminated. The simplicity of this topology makes it cheaper and less bulky than the superheterodyne. However this architecture is not suitable for an implementation of digital I and Q. The reason is that the need to have a low sampling clock results in a low-IF frequency. This means that the resulting image frequency is very close to the desired RF frequency and this makes it impossible to effectively filter any image frequency in the RF.

An undesired DC offset in IQ signal will shift the origin of the baseband signal and causes signal saturation and/or degradation in BER performance. Modulations which have signal content near DC, are more susceptible to this phenomenon. As a result of the effects of a multi-path channel and mobility, there can be abrupt variations in the channel conditions. This can imply a change in the DC offset in the receiver, requiring an AGC dynamic compensation.

An imperfect LO phase shift leads to I and Q imbalance. This I and Q imbalance can cause asymmetry and rotation in the signal constellations. If the I and Q imbalance is perfectly known, it might be compensated for digitally. However, the effects being frequency dependent, and perhaps even dependent on the environmental conditions, complex calibration circuits for the I and Q imbalance will be needed. An attractive alternative is the generation of the in-phase and quadrature oscillators by division from a common higher frequency (18 to 24 GHz) oscillator with appropriate data latching. Such high frequency VCO's are both difficult to design with good performance and difficult to work with (i.e. clock rate of dividers). However, we believe that improvements in fabrication technologies can only make this design easier in the future, and so we believe that the Zero-IF solution is an attractive alternative for future broadband wireless systems.

B. Oscillators

Probably the most important parameter to consider when choosing the oscillator is its phase-noise. This is particularly true for OFDM modulation. Note that all values given below are based on one typical hardware implementation; they are not suitable to draw conclusions on the smallest possible carrier spacing, etc: we make the assumption that the C/I contribution from the oscillators should be kept under 35 dB. As there will be two 5 GHz oscillators, one in the transmitter and one in the receiver, the noise contribution of each should be kept under 38dB. Working on the basis of 2048 carriers in 100 MHz, we'll therefore assume a subcarrier spacing of 48.828 kHz, when calculating the C/I contribution of the phase noise.

The worst C/I contribution of the phase noise will be given when considering the central OFDM carrier. We use a formula from [Sto98] to calculate the contribution of the phase noise to the C/I and assume that the equivalent of today's high quality commercial oscillator will be available for use in future low cost applications. Such an oscillator might have a phase noise of -70 dBc/Hz (determined by the PLL) and of –115 dBc/Hz (determined by the VCO) at 1 MHz offset from the carrier. We find that the above specification achieves 32.4 dB of C/I contribution from a single oscillator, and thus the total noise contribution from the oscillators will be 29.4 dB. This is poor, relative to the nominal specification of 35 dB discussed above, and will add an unavoidable noise floor to the system. This problem can be addressed by using a larger inter-subcarrier spacing (reducing the number of subcarriers from 2048 to 1024, results in a doubling of the carrier spacing and in a

C/I contribution of a single oscillator of 36.4 dB, or 33.4 dB for the total oscillator contribution.), or by signal processing techniques to compensate for the phase noise at the receiver.

C. Filter Design: Feasibility of adjustable filters

If the required multi-band signal spans a frequency range beyond the capability of a single receiver down-converter and A/D converter, an analog channel selection filter must be used, with an adjustable bandwidth. We propose two possible structures for this filter, switched capacitor and switched resistor.A 50 MHz bandwidth tunable switched capacitor filter has been demonstrated by [Moo00]. So we believe that a 100 MHz bandwidth tunable filter must be possible to design if not now in the very near future.

Switched resistor filters are typically found integrated into ASICs. In such an environment the exact value of the passive components is difficult to precisely control, and can vary from their nominal value by 10 or even 20%. This large variation would be catastrophic for a filter design except for the fact that the largest part of this variation is systematic and is the same for all passives, with only a small residual difference between individual components. This is due to the fact that a single ASIC comes from a small portion of the wafer and suffers exactly the same process variations that resulted in the large difference from the passives nominal value. If all of the resistors or all of the capacitors in an RC filter vary by an equal percentage, the result is a variation in the corner frequency of the filter, while the shape of the filtering function remains the same. In many communications systems this variation in the corner frequency results in degradation in the performance, especially of the adjacent channel rejection. The means to alleviate this is with a switched resistor filter. If we know that the resistor values might vary by ±10%, but we can only accept a 5% error, we might replace every resistor in the filter by three resistors, one at the nominal value, and the other two at ±5% respectively. A calibration process can be used to select the appropriate resistor to use to maintain the specification on the corner frequency. In general all of the resistors vary by the same percentage. However, we aren't constrained to just small variations in the resistor values and can therefore envisage that such a design trick might be extended to allow multi-bandwidth filters to be created. The disadvantage is that a large number of different resistor values might be needed for both the process variation and the multi-bandwidth capabilities. This is in contrast to the switch-capacitor design where the corner frequency is altered with a varying clock signal.

IV. Effect of HPA Nonlinearity on Transmitted Spectrum

Neighbouring desired and undesired user spectra may be received with large power variability due to differing path losses. Avoidance of adjacent channel interference then requires low transmitted power spectral sidelobes and rather stringent spectral masks. For example, allowable interference to adjacent-frequency receivers is usually specified in terms of maximum interference power at a certain distance and at a certain frequency offset from the interferer's carrier. Under typical transmitted power and path loss conditions, this may imply spectral masks with as much as 30 to 50 dB of attenuation of transmitted spectrum sidelobes at a normalized frequency offset of 1. Nonlinear distortion caused by the transmitter's high power amplifier (HPA) is a major factor influencing spectral sidelobe levels. Power backoff and transmitter signal processing techniques to reduce the dynamic range of the transmitted waveform can be used to reduce spectrum sidelobe levels. The serial modulation approach produces a transmitted waveform with a lower dynamic range than that of the OFDMA approach, and therefore requires a lower power backoff than that of a comparable OFDMA signal, resulting in improved high power amplifier (HPA) efficiency and higher average transmitted power for a given allowable level of nonlinear distortion. In practice it is usually found that the HPA nonlinearity has a more critical effect on the re-growth of power spectrum sidelobes than on bit error rate performance. Consequently, in this section, we focus on the effect of backoff and nonlinearities on power spectrum sidelobes in transmitted multi-band waveforms.

Figure IV-1 shows the instantaneous normalized transmitted power distributions for OFDM and serial modulated signals for S=2 bands, with the same transmission parameters as in Figure II-2. No explicit PAPR-reduction schemes were applied (unless the extra DFT operation in serial modulation is thought of a PAPR-reduction scheme for OFDMA). Spectral sidelobes were reduced by applying a 5.5 % raised cosine time window to each waveform. It is clear that the serial modulated waveform has a significantly lower dynamic range than the corresponding OFDMA waveform. This implies that the serial modulated waveform will require a lower power backoff at the input of a nonlinear transmitter power amplifier. This is confirmed by Figure V-1, which shows the power spectra of the two waveforms, with different power backoffs, at the output of an amplifier whose nonlinear characteristic is modeled by a AM/AM conversion Rapp model [RAP91], with parameter p=2. Comparison of these two cases shows that the spectral sidelobe levels for serial modulation with a power backoff of 8 dB is about the same as that for OFDMA with a power backoff of 10 dB. Similar curves (not shown) have been generated for examples of S=4 sub-bands. As expected, the difference in IAPR distribution and power spectrum sidelobe re-growth between OFDMA and serial modulated signals further narrows as the number of sub-bands increases. As S increases much beyond 4, serial modulations loses its PAPR reduction advantage over OFDMA.

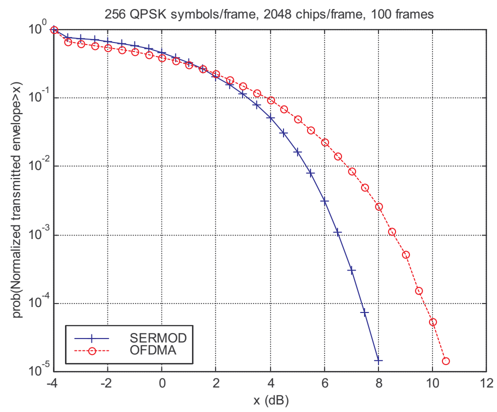

Figure IV-1 Distribution functions of instantaneous average power ratio (IAPR) of S=1 sub-band OFDMA and serial-modulated signals, N_c=2048, M_N=256, S=2.

V. CONCLUSIONS AND SUMMARY

Multi-band transmission is feasible and is readily implemented using a combination of digital and analog signal processing techniques. Transceivers can use frequency domain processing with efficient FFT and inverse FFT techniques to generate and receive multi-band signals and separate them from unwanted signals. The main differences from generation and reception of conventional signals which occupy a single spectral "chunk", are: (1) the mapping operation between data and subcarrier frequencies, and (2) some care in channel estimation - to interpolate measured channel frequency responses only within sub-bands. If the total range of available frequencies exceeds the ADC and DAC capabilities of individual transceivers, multi-band transmission can be implemented by a combination of RF analog and digital signal processing techniques.

Two main categories of signals are of interest for multi-band transmission: parallel modulated signals, which include OFDMA, and (2) serial modulated signals. Both categories transmit signals in blocks, separated by cyclic prefixes. Both categories can also be considered as forms of multicarrier modulation. For S sub-bands, serial modulation is equivalent to S single carrier signals transmitted in parallel, each up-converted by a different frequency. For moderate values of the number of sub-bands S, serial modulated multi-band waveforms have a lower dynamic range, which makes them less susceptible to impairment from nonlinear power amplifiers. In particular, these signals can tolerate lower transmit power backoff than comparable parallel modulated signals for the same spectrum sidelobe re-growth. This means that transmitters can use cheaper power amplifiers, a very important consideration for restraining the cost and power consumption of user terminals transmitting to base stations.

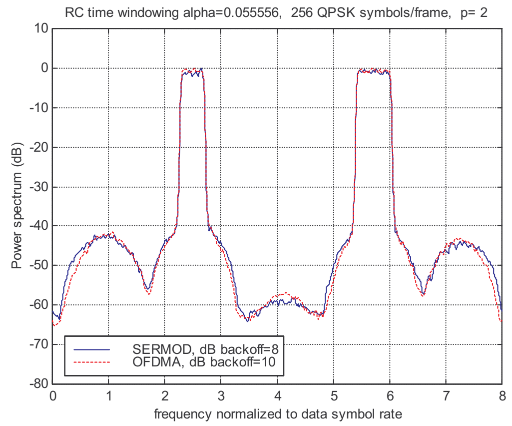

Figure V-1 Power spectra of OFDMA and serial modulated signals emerging from a RAPP model, *p*=2 nonlinearity, with 10 dB and 8 dB power backoffs, respectively

For the RF front end, the zero-IF option appears promising and feasible, given its lower parts count and complexity, and recent advances in fabrication technology. Phase noise was identified as a potential RF implementation issue for broadband multicarrier systems, solvable by judicious choice of the number of subcarriers or by signal processing. Two technologies for analog band-selection filters were considered – switched capacitor and switch resistor. The choice depends on the number of different bandwidths that need to be accommodated.

Further elaboration of the multi-band concept is under way in the IST WINNER project, including: elimination of the cyclic prefix through OFDM/IOTA [LJG02], [FAB95], application of pseudo-random postfix for channel estimation [MCD+03], and reduction of power backoff requirements through signal processing [BMH01], [MH97] and HPA pre-distortion [WP00].

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REFERENCES

[BMH01] M. Breiling, S. H. Muller-Weinfurtner, and J. B. Huber, "SLM Peak-Power Reduction Without Explicit Side Information," IEEE Communications Letters, vol. 5, pp. 239–241, June 2001.

[FAB95] B. Le Floch, M. Alard and C. Berrou, "Coded Orthogonal Frequency Division Multiplex", Proc. IEEE, Vol. 83, No. 6, June, 1995.

[FCC03] FCC web site on cognitive radio: www.fcc.gov/oet/cognitiveradio/

[FK04] D. Falconer and S. Kaiser, "Broadband frequency Domain-based Air Interfaces", presentation at WWRF meeting #11, Oslo, June, 2004.

[HNM+04] S. Hijazi, B. Natarajan, M. Michelini, Z. Wu and C.R. Nassar, "Flexible Spectrum Use and Better Coexistence at the Physical Layer of Future Wireless Systems via a Multicarrier Platform", IEEE Wireless Communications, April 2004, pp. 64-71.

[LJG02] D. Lacroix, J.P. Javaudin, and N. Goudard, "IOTA, an Advanced OFDM Modulation for Future Broadband Physical Layers", Wireless world Research Forum (WWRF) meeting #7, Dec. 2002.

[MCD+03] M. Muck, M. de Courville, M. Debbah and P. Duhamel, "A Pseudo Random Postfix OFDM Modulator and Inherent Channel Estimation Techniques", Proc. Globecom 2003, Dec., 2003.

[MH97] S. H. Müller and J. B. Huber, "A Comparison of Peak Power Reduction Schemes for OFDM," in Proc. IEEE Global Telecommunications Conference (GLOBECOM'97), Phoenix, AZ, USA, pp. 1–5, November 1997.

[Moo00] Un-Ku Moon, "CMOS high-frequency switched-capacitor filters for telecommunication applications", IEEE J. Solid State Circuits, vol. 35, pp. 212-220, Feb. 2000

[RAP91] C. Rapp, "Effects of HPA-Nonlinearity on a 4-DPSK/OFDM Signal for a Digital Sound Broadcasting System", Proc. 2nd European Conf. On Satellite Communications, Liege, Oct., 1991, pp. 179-184.

[Sto98] J. Stott, "The effects of phase noise in COFDM", European Broadcasting Union Review Technical, no 276, pp. 12-25, 1998.

[WP00] K. Wesołowski and J. Pochmara, "Efficient Algorithm for Adjustment of Adaptive Pre-distorter in OFDM Transmitter", Proc. of IEEE Vehicular Technology Conference, Boston, September 2000.