Wireless LANs

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Abstract

Wireless LANs based on the IEEE 802.11 standard have achieved wide customer acceptance in the enterprise environment. They are expected to continue to expand in popularity and become ubiquitous communication systems even in private and public places. This paper discusses the basics of the wireless LAN physical layer, focusing on radio transceiver specifications and design options.

1. Introduction

The introduction and proliferation of data network wireless access is a natural evolution in modern communication systems, stimulated by the promise of user mobility and freedom from wires and greatly encouraged by recent advances in portable wireless terminal technology. Just as already is the case of cellular voice communications, wireless data access is on the way of becoming a universal modern capability. For example, small and inexpensive PCMCIA modules, which readily attach to laptop computers, are available to make multi Mb/s wireless connections with access points strategically located within enterprise buildings, which further connect the users to wired LANs, intranets, etc. Likewise, in the home or in public places, wireless LANs will increasingly provide valuable data communication channels.

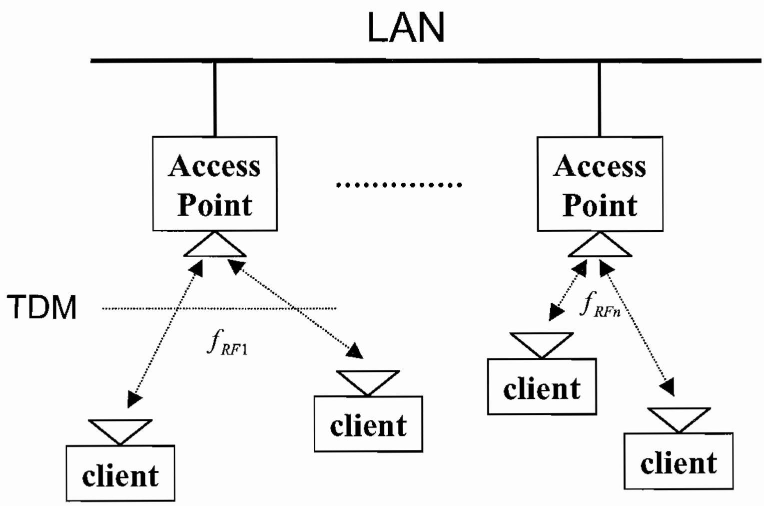
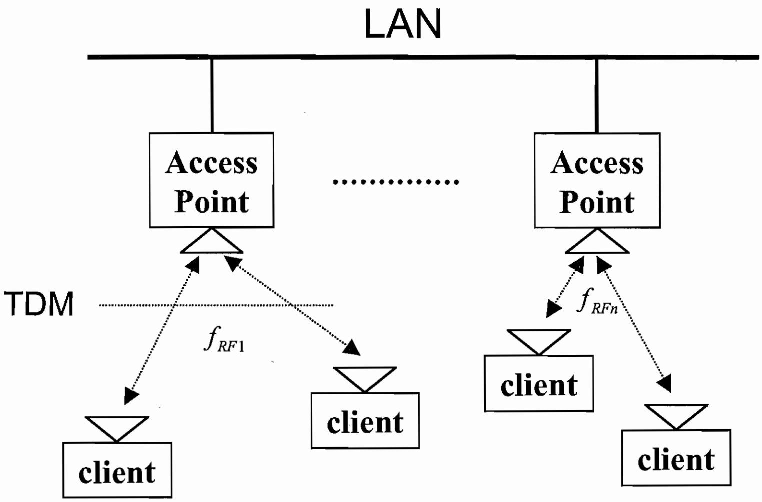


Figure 1: Typical Wireless LAN configuration

Typically, such networks are configured as in Fig. 1 with multiple users connected to each access point via a carrier sensing multiple access with collision avoidance (CSMA/CA) protocol. Naturally, the steady state data throughput for each user is the total system data rate after subtracting the network overhead divided by the number of users connected to one access point.

The uncontested success of the current generation wireless LAN technology with speeds up to 11 Mb/s based on the IEEE 802.11b standard [1-2] has been impressive but, it could be argued that it is based on previously established digital cellular and cordless technology. For example, the substantially higher data rates and channel-bandwidth of wireless LANs compared to cellular systems are balanced by lower sensitivity and blocking requirements, yielding similar transceiver design strategies, integration level, etc. However, for speeds in excess of 50 Mb/s as specified by the 802.11a standard [3], new and difficult transceiver design challenges arise, especially in the context of low power



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dissipation and low cost. In this paper we discuss the current wireless LAN technology and the new challenges designers will face for higher speed systems.

2. Current Systems based on the 802.11b standard

2.1 Modulation and radio specifications

Originally, the 802.11 standard was written for 1 Mb/s and 2 Mb/s data rates in the 2.4 GHz -2.5 GHz ISM band, possibly using direct sequence code division multiplexing in combination with DBPSK and DQPSK modulation, respectively. An eleven-chip long Barker sequence provides processing gain, which relaxes the required SNR to below 0 dB. The channel bandwidth of 14 MHz placed anywhere in the band on a 5 MHz grid allows network configurations with 3-4 access points in close physical proximity. The maximum allowed RF transmitting power is 30 dBm but typically, 15 dBm is used in existing systems.

The 802.11b standard option enhances the wireless LAN data rate to a maximum of 11Mb/s by Complementary Code Keying (CCK) modulation [4]. While still using the same chip rate in order not to change the RF signal bandwidth, a much-reduced processing gain accommodates the higher data rate to the expense of approximately 10 dB higher SNR requirements. Practically, at 11 Mb/s CCK is equivalent in almost all respects to regular DQPSK.

2.2 Wireless transceiver solutions

The recent advances in RFIC and radio system technologies have provided ample opportunities for the realization of miniaturized and economically viable wireless LAN transceivers. Typically, these blocks are implemented as shown in Fig. 2 using a few ICs and several hundred passives (mostly by-pass capacitors), packaged tightly into small modules such as PCMCIA cards.

Usually the cost of such modules is well within the consumer electronics market demands.

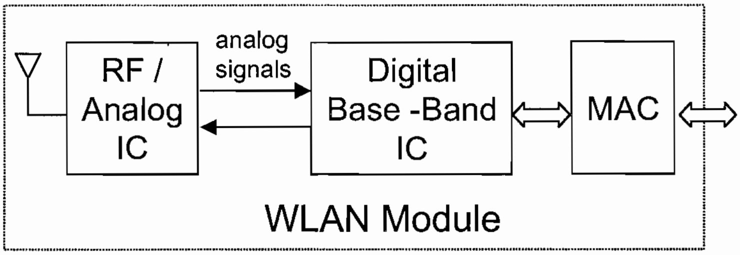

Figure 2: Typical Wireless LAN transceiver module

Focusing on the physical layer, notice that a radio chip and a base-band chip are typically used with analog I/Q transmit and receive interfaces. The base-band chip is mostly a digital circuit, containing only data converters as analog blocks. This system partitioning minimizes the digital switching noise coupling into the radio sections and provides low power chip-to-chip analog interfaces. The radio chip may be designed in different technologies such as Si bipolar, SiGe BiCMOS, or recently, even in straight CMOS. Typically, a –75 dBm sensitivity is accomplished for about 200 mW receiver power dissipation. The radio architecture has evolved from a conservative superheterodyne approach to less expensive direct down/up conversion. The efficiency of the linear power amplifier is limited by the signal peak-to-average ratio, which is moderate, allowing reasonable transmitter power dissipation, typically 500 mW.

3. Emerging Systems based on the 802.11a standard

3.1 Frequency bands, RF power levels, modulation formats, and data rates.

In order to enable data rates up to 54 Mb/s and to increase the number of channels for easier network planning, the 802.11a standard specifies three 5

GHz ISM band sections (in US, similar in other countries), each containing four 20 MHz channels. The first band is from 5150 MHz to 5250 MHz and it allows up to 16 dBm transmitting RF power. The second one is from 5250 MHz to 5350 MHz with 23 dBm maximum power, and the third one, mainly intended for outdoor applications, is from 5725 MHz to 5825 MHz with 29 dBm maximum power. Eight data rates are provided (only three are mandatory), supported by various modulation techniques and coding schemes. Since the realization of the highest rate performance, 54 Mb/s using OFDM/64-QAM modulation, is the most challenging design aspect of 802.11a transceivers, for most of the following considerations we will focus on this topic.

3.2 OFDM

Orthogonal Frequency Division Multiplexing (OFDM) [5], used in the physical layers of both 802.11a and HiperLAN/2 [6] is a special case of the classical frequency division multiplexing, in which the sub-carriers are orthogonal to each other in time domain, i.e., if any two are multiplied and integrated over a symbol period the result is zero. Therefore, an OFDM signal is a bank of narrow band-pass information-carrying sub-signals, placed very close to each other, as shown in Fig. 3. The time domain orthogonality is reflected in frequency domain as the property that each sub-signal spectrum is zero at the carrier frequencies of all the other sub-signals (the channel spacing is equal to the symbol-rate).

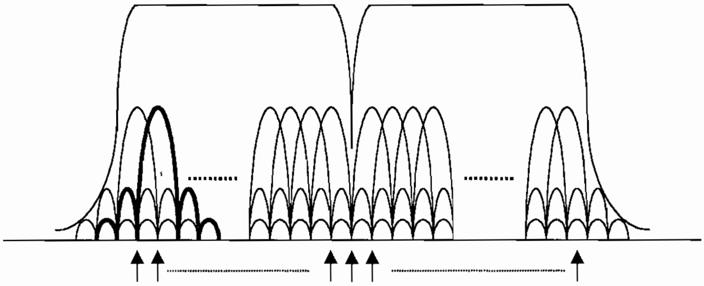

Figure 3: OFDM signal spectrum

As a result, the sub-signal center frequencies are very close to each other leading to high spectrum efficiency. An equivalent mathematical explanation of OFDM is based on the fact that a Discrete Fourier Transform uniquely relates two sets of N complex numbers: N time samples with N frequency samples. In fact, the usual way of demodulating an OFDM signal is by performing an FFT on N time samples resulting on N magnitude and N phase quantities, which represent the transmitted information. Compared to a single-carrier system, the symbol period increases while the overall data-rate remains unaffected. In addition, as the symbol-rate becomes N times longer, the guard time interval commonly introduced to avoid ISI (inter symbol interference) adds only a relatively small overhead. Nevertheless, this brings a considerable hardware saving, as a time-domain equalizer now becomes unnecessary. A frequency-domain equalizer is still indispensable for the purpose of compensating the channel frequency response, which may change from one sub-carrier frequency to another making correct data detection impossible otherwise. However, this frequency equalizer is simple: consisting of only a single multiplication of every sub-carrier with a complex number. In summary, the main advantage of OFDM over single carrier modulation techniques is that a hardware intensive time-

domain equalizer is replaced by an FFT-operation followed by a simple frequency-domain equalizer. This is especially advantageous in high-rate systems. The most important drawback of OFDM is the large peak-to-average ratio (PAR) of the signal, particularly detrimental to transmitter linearity requirements.

In the 802.11a 54 Mb/s mode, there are 52 sub-carriers; 48 of which are modulated with 64-QAM. There are four pilots, i.e., sub-carriers without any modulation, which enable coherent detection. Based on this simple description of OFDM it is apparent that designing 54 Mb/s 802.11a wireless LAN transceivers raises new difficulties compared to previous generation systems. It will be shown that the minimum required SNR of the received OFDM/64-QAM signals is about 30 dB, substantially higher than in other digital wireless systems. In addition, the presence of narrow-band sub-carriers across the channel implies accurate processing of the whole channel spectrum. For example, simple circuit solutions for broadband receivers such as AC coupling in direct conversion stages are not appropriate. Finally, the transmitter linearity requirements impose serious efficiency limitations to conventional power amplifiers.

3.3 Basic Transceiver Specifications

Using the standard, one can derive the basic transceiver specifications. The following approximate calculations are not intended to give precise design values but rather to indicate the rough figures for 802.11a radio systems.

Figs. 4 and 5 show the power spectral density (PSD), $L(f)$ and the power levels P of the desired and undesired (noise) signal components observed on the receive side of an OFDM communication link, under different limit conditions described below. The overall propagation loss is assumed such that the received

power level of the desired signal P_s be always –65 dBm. This is the receiver sensitivity level in 54 Mb/s mode, required by the standard for operation below a certain packet error rate. Figs. 4 and 5 illustrate how the radio channel affects the transmitted signal in receiver noise dominated and transmitter noise dominated conditions, respectively. In each case examples are shown for a frequency-flat (FF) channel and a frequency-selective (FS) channel.

The FF Channel Cases: In Figs. 4a and 5a the time delay spread of the channel is assumed significantly lower than the temporal resolution of the OFDM signal, i.e., the delay is smaller than 50 ns, the typical sampling period in 802.11a. Under this assumption, the wireless channel affects each sub-carrier of the OFDM signal in the same way, which leads to a flat PSD of the desired signal component at the receiver. In fact, the power level requirement in the standard refers to this type of channel.

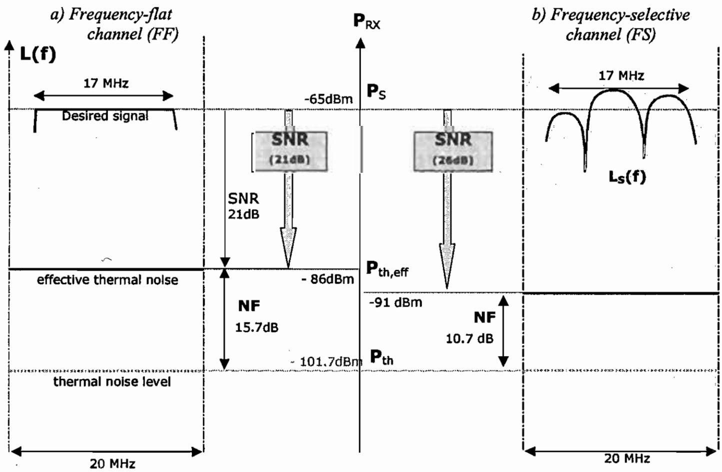

Figure 4: PSDs and signal powers at the receiver, assuming thermal *receiver* noise with given noise figure (NF) only. Shown for frequency-flat (FF), and frequency-selective (FS) channels.

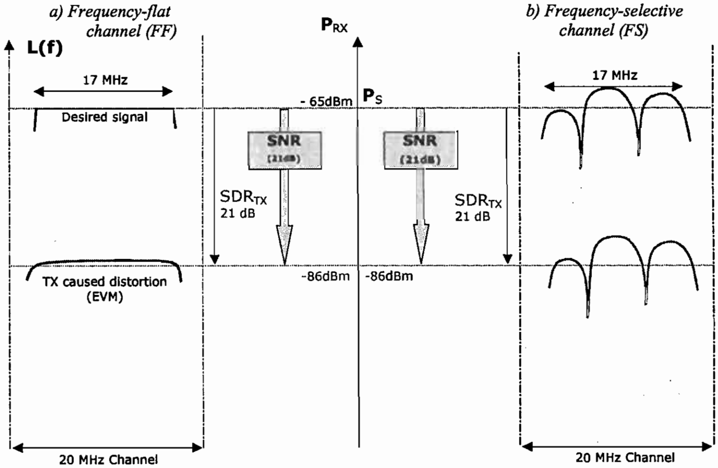

Figure 5: PSDs and signal powers at the receiver, assuming *transmitter* caused imperfections only. Shown for frequency-flat (FF), and frequency-selective (FS) channels.

The FS Channel Cases: In Figs. 4b and 5b the average time delay spread of the channel is assumed to equal or exceed the 50 ns time resolution of 802.11a OFDM. This leads to destructive and constructive multi-path interference, creating sub-carrier PSD levels above or below the average PSD. The actual indoor environment for typical practical applications of 802.11a systems is an FS channel. Note that many channel frequency response "snapshots" are equally valid, where the signal fading occurs at different frequencies than in Figs. 4b and 5b but having the same normalized integral power P_S. Next, we discuss the signal-to-noise and signal-to-distortion ratios (*SNR, SDR*) at the receiver A/D output, which are the primary overall design requirements.

First, we consider the limit case of Figure 4 when the background thermal noise and the receiver generated noise, usually expressed as the input-referred noise figure (NF), are the only sources of disturbance in the communication link. Starting from the –174 dBm/Hz background thermal noise power and

adding 72.3 dB corresponding to 17 MHz noise-bandwidth, we obtain the effective antenna noise power $P_{th}= -101.7$ dBm, shown as the dashed line in Fig. 4. Subtracting this number from the required –65 dBm receiver sensitivity leads to an SNR at the input of the receiver of 36.7 dB. Simulations show that the static SNR at the receiver A/D output necessary to meet 10% packet reliability as required by the standard for an ideal additive-white-Gaussian-noise (AWGN) channel, is approximately 21 dB. Hence, for this *FF* limit case, we have about a 15-16 dB maximum receiver NF allocation. This is shown in Fig. 4a, resulting in an effective noise level of $P_{th,eff}$ = -86 dBm. In contrast, if frequency selectivity is present (*FS* case), the previous calculation must be amended by a "channel correction factor" of about 5 dB, increasing the necessary SNR to 26 dB, as illustrated in Fig. 4b. This is due to the fact that the sub-carriers in deep fade (requiring additional SNR) dominate the overall performance.

In a different limit case, shown in Figure 5, we consider the *FF* and *FS* channels under the assumption that the only relevant source of disturbance is transmitter imperfection, commonly referred to as *transmitter implementation noise*. Typical transmitter-related noise sources include the effects of transmitter non-idealities such as oscillator phase noise, finite linearity of the transmit chain, finite digital word length, and limited power amplifier (PA) back-off (see next subsection). These phenomena cause an error between the desired signal and the actual transmitted signal measured by the "error vector magnitude" (EVM). The standard specifies a maximum average RMS value for the EVM. The EVM-related noise process is proportional to the desired signal, and hence is specified by a relative dB number. For example, the standard specifies –25 dB EVM for the 54 Mb/s mode. In a first-order approximation, the in-band noise, caused by non-linear transmitter effects, is an AWGN process. Therefore, in the *FF* case an identical calculation with that of the previous

paragraph yields a SDR_{TX} value of 21 dB, necessary to meet the specified packet error rate. Since only transmitter imperfections are present, this number translates directly into 21dB receiver SNDR shown in Fig. 5a. In contrast to the situation of Figure 4, the same calculation is valid for the *FS* case shown in Fig. 5b, and the "channel correction factor" is zero! The reason for this property is that in the transmitter, all sub-carriers have equal power (before passing through the channel) and thus all are affected by transmitter noise equally. Hence, the sub-carrier signal-to-noise ratio remains unchanged during channel propagation.

A third limit case is when the only source of disturbance in the communication link comes from the receiver distortion, commonly known as *receiver implementation noise*. This type of disturbance is signal dependent and is produced by many non-idealities such as local oscillator noise, non-linearity in receiver chain, I/Q imbalances, DC offsets, A/D converter quantization noise, residual adjacent channels or blockers due to insufficient filtering, etc. The resulting interference is a near-Gaussian and frequency-flat noise signal, essentially directly related to the desired signal by some number SDR_{RX} in dB. For the *FF* case, the overall SNR requirement is 21 dB, as in the previous calculations. For the *FS* case, the "channel correction factor" is nonzero since the desired signal exhibits faded sub-carriers whereas the noise signal is flat and added *after* the channel propagation, hence affecting the weak sub-carriers. In actual systems, all noise/distortion processes caused by transmitter and receiver imperfections and the receiver thermal noise effectively add in the receiver. Depending on the actual spectral noise shape, the "channel correction factor" is between 0 and 5dB. Assuming an overall transmitter/receiver *implementation loss* of about 3-4 dB and -65 dBm sensitivity (corresponding to SDR_{TX} and SDR_{RX} values greater than 30 dB), the receiver NF must be 7dB or lower.

Notice that the only way the design methodology can make a difference in the transceiver performance is by minimizing the receiver NF and the various

practical errors mentioned previously. For this reason it is instructive to identify these errors and investigate the circuit blocks where they are produced in more detail.

3.4 Typical transceiver design issues

Before focusing on specific transceiver issues, we point out a clear distinction between the thermal noise and the implementation noise. The latter is expressed relative to the intended signal while the former is expressed as an absolute value. As a consequence, implementation noise is always important, independently of the distance between the transmitter and the receiver. On the other hand, bringing the receiver closer to the transmitter decreases the effect of thermal noise due to higher signal strength. In a PER (packet error rate) against SNR performance plot, the implementation noise appears as an impenetrable PER floor. A measure of the "implementation loss" is the amount of PER curve shift after applying implementation noise at the PER value of interest.

Summarizing the previous discussion, the total noise at the output of the receiver A/D converter is the result of contributions from three types error sources: transmitter noise with maximum level fixed in the standard, receiver thermal noise, and receiver implementation noise. The latter category can be divided into noise sources that are always present (e.g., integrated phase noise, quantization noise, DC offset, etc) and noise sources that are only present when a blocking signal is applied. During a gain/noise/linearity budget analysis it is important to make a distinction between these cases, as the thermal noise contribution is halved (intended signal is 3 dB above the sensitivity level) when a blocking signal is present.

As an illustration of typical transceiver design issues, next we will focus on several sources of implementation noise.

PA back-off: As mentioned in subsection 3.2, OFDM suffers from a high signal PAR. As a result, the necessary transmitter dynamic range is higher than that of the 802.11b case. The gains in the various blocks are set such that the average power level stays below the 1dB compression point by a certain dB amount called the back-off. As this value is usually smaller than the PAR for power efficiency reasons, signal clipping occurs and the corresponding interference produces implementation noise. The PA back-off value is extremely critical since the already low 10-15% efficiency of linear PAs [7] is easily decreased further. In order to determine the appropriate PA back-off, first, the EVM requirement should be met (-25dB for 54Mbps) and second, the spectral re-growth due to clipping should be limited within the specified spectral mask. Fig. 6 is an illustration on how clipping causes spectral re-growth.

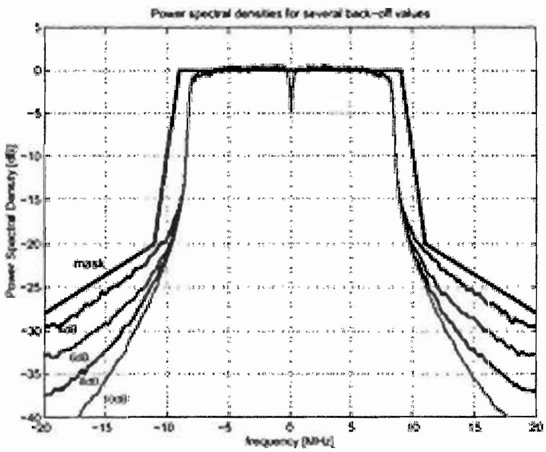

Figure 6: Transmitted signal spectrum for various PA back-off values

Transceiver Linearity: In OFDM, intermodulation of sub-carriers is of great concern, as the resulting products fall in channel exactly at the frequencies of other sub-carriers, corrupting the information carried by them. For example, the receiver maximum signal specified by the standard is –30dBm. Two

neighboring sub-carriers increased by sub-carrier PAR of 3dB (sub-carrier PAR is much smaller than the total OFDM signal PAR) intermodulate and corrupt other sub-carriers. Assuming a typical 5 dB margin allocated for other implementation noises, the resulting minimum receiver input IIP3 is about −10dBm. A similar analysis is valid for the second order intermodulation products, usually with less stringent effects but strongly coupled to the choice of receiver architecture (important in low IF and direct conversion).

Channel Selection Filtering: An important contributor to the receiver implementation noise is the residual blocker signal after channel selection filtering. As an example we consider a filter for an 802.11a low-IF receiver. This filter may be a complex band-pass continuous-time circuit. The passband is 17 MHz, which is one channel-width wide, and the center frequency is at the 10 MHz low-IF. The filter must attenuate the unwanted interferers/blockers in order to reduce the aliasing noise (produced by sampling before A/D conversion) to acceptable levels. The worst-case blocking specification for the 54 Mb/s data rate is –63 dBm level adjacent channel while the desired signal is at –62 dBm level (3 dB higher than the sensitivity level).

A conservative design uses a 6-order type-one Chebyshev filter with 0.5 dB pass-band ripple. The frequency response of this filter shown in Fig. 7 provides in excess of 30dB rejection for all sub-carriers in the blocking signal. In practice circuit imperfections may degrade the performance especially at the edges of the channels.

A source of blocker-dependent implementation noise, related to this low-IF filter, is the limited image-rejection due to circuit imbalance. Typically, an image-rejection of 30dB is achievable without compensation algorithms. As the

required SNR in 802.11a exceeds this number, a compensation algorithm is required and the resulting implementation loss has to be taken into account.

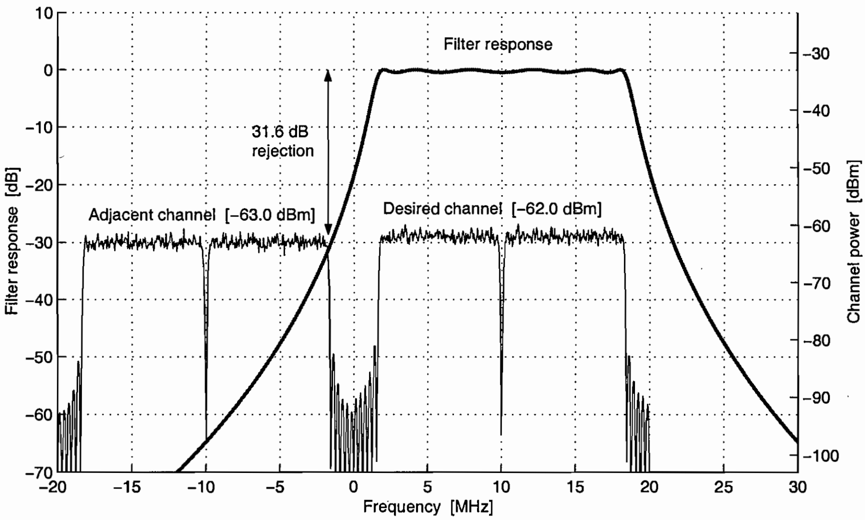

Fig. 7. Channel-selection filter and blockers

3.5 Transceiver Design Choices

In this subsection we discuss several design possibilities of key blocks, without attempting an exhaustive treatment of this topic. The receiver technology is always a prime concern in any RFIC design so various alternatives will be discussed with more details on a new low-IF sampling solution. The stringent specification of the power amplifier linearity is a major limitation to its power efficiency, ultimately resulting in high power dissipation. Possible alternative design methods are mentioned. Finally, the perennial question of which IC technology is best suited for this application will be addressed.

Using low integration multiple IF super-heterodyne receivers: This is the

most conservative design choice for 802.11a receivers. The required high performance as described earlier can be met readily if enough external filters and other precision RF components are used. Of course, the cost will almost surely be too high for this application.

Using highly integrated single IF super-heterodyne receivers: Having a single IF SAW filter in addition to an RFIC and few external components may be a proper compromise between cost and performance. However, the design is still challenging due to analog I/Q down-conversion from IF to base-band. In order to insure final 30 dB SNR, excellent image rejection and linearity must be accomplished in the presence of the usual phase and magnitude errors, A/D converter imbalances, DC offsets, etc. In addition, the large frequency error between transmitter and receiver synthesizers allowed in the standard could place an OFDM sub-carrier very close to DC after the final down-conversion. There, substantial offsets and 1/f noise may corrupt the information in the respective sub-carrier.

Using highly integrated super-heterodyne receivers with low-IF sampling: An attractive theoretical method for mitigating most problems related to analog conversion to base-band is by using IF sampling. Digitizing at IF removes any I/Q imbalance errors (down-conversion done digitally) and also avoids any DC offset or 1/f-noise problems entirely. Unfortunately, this approach requires two successive IF SAW filters in order to avoid aliasing of blockers close to the channel. Naturally, this is not attractive for cost reasons. However, it is possible to replace one SAW filter [8] by a combination of a fully integrated continuous-time complex filter at 10 MHz (second low-IF) with a complex (I/Q) sampling circuit [9]. The receiver block diagram is shown in Fig. 8. After a first IF conventional stage using a SAW filter, the signal is converted to a 10 MHz low IF and is processed through a continuous-time fully integrated complex band-pass filter. A similar filter for Bluetooth is described in

[10]. The frequency response of a three-pole Butterworth complex filter including practical component mismatches is shown in Fig. 9, including the leakage signal due to mismatches. A further 8 dB rejection of the adjacent channel is accomplished by the complex sampling operation, which exhibits a notch filter characteristic given in Fig. 10.

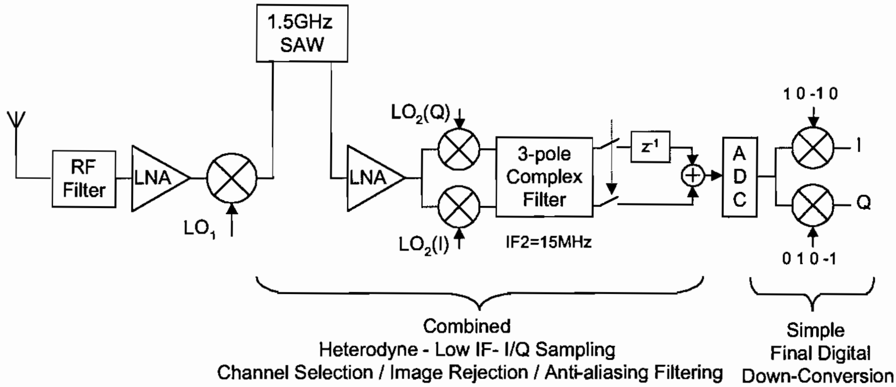

Figure 8: Superheterodyne receiver with low-IF sampling

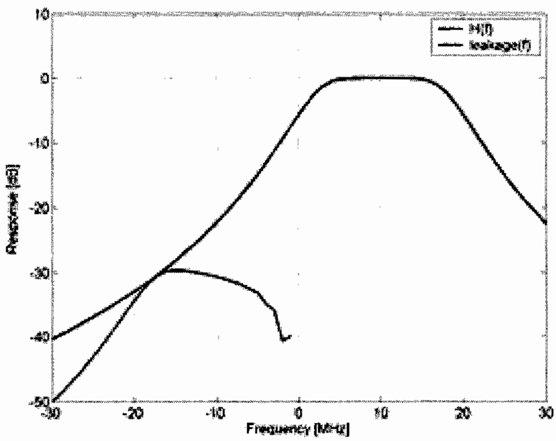

Figure 9: Frequency response of continuous-time complex filter including practical mismatches

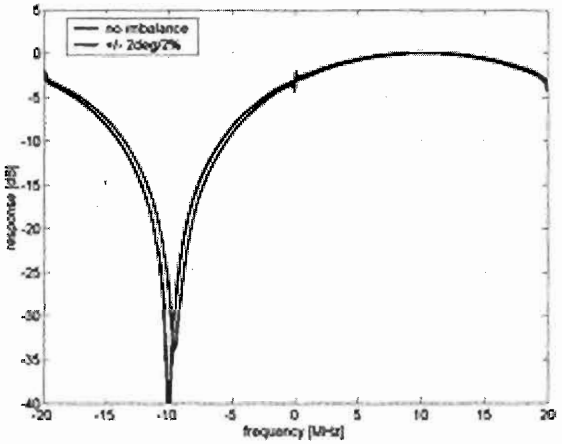

Figure 10: Frequency response of complex (I/Q) sampling with notch filtering

This method does not accomplish full integration, however, compared to the previous approach, it makes more efficient use of the hardware resources with very little if any power dissipation penalties and better performance. Notice that the sampling frequency is only twice the value of the base-band sampling case and a single A/D converter is used rather than two.

Using fully integrated low-IF receivers: This recently popular method is attractive because it accomplishes full integration and provides high tolerance to DC offsets, 1/f noise, and frequency errors. The most important challenge for this technique is the image rejection precision in high-order complex filters, driven very hard by the OFDM SNR demands. Notice that, there is no SAW filter to help the complex filters in this approach as in the IF-sampled technique.

Using fully integrated direct conversion receivers: Despite fundamental limitations, direct conversion has become an accepted and even preferred receiver method in many applications. The most attractive feature is its simplicity, although this may be deceiving when DC offset cancellation loops

and extra linearity demands are included in the picture. The application of direct conversion to 802.11a is possible but very difficult due to I/Q matching requirements and the large transmitter-receiver frequency errors. Notice that in this case the situation is much more serious than in the single-IF super-heterodyne approach because first, the DC offset is substantially higher and second, the desired signal is not amplified by an IF amplifier prior to DC conversion. Two options remain: narrow down the AC-couple cut-off frequency as much as possible and accept the associated loss, or correct the local oscillator synthesized frequency with very fine resolution. The latter may require the use of complicated fractional synthesizer techniques, which in turn introduce extra noise and spurious signals.

Using pseudo direct conversion receivers (wide/sliding IF): This techniques, used recently in a fully integrated CMOS 802.11a radio RFIC [11], converts the RF signal to base-band in two consecutive mixing operations. Some of the classical direct conversion problems, such as LO leakage in the RF band and DC offset from RF LO self-mixing, are avoided but the lack of an IF filter/AGC strip increases the receiver NF.

The power amplifier efficiency problem: As discussed previously, the OFDM signals have large peak-to-average ratios, which requires PA operation in class A with large back-off. This is reflected in lower power efficiency with serious repercussions in terms of total transmitter power dissipation. The application of PA linearization techniques could improve the efficiency. In addition, two methods are known to potentially use efficient nonlinear PAs and still achieve linear amplification. However, these techniques are yet to be widely introduced in RFIC products. The first method, known as the LINC technique [12], decomposes the band-pass signal into two or more constant envelope signals. These can be amplified by highly efficient switching PAs and then recombined before sending to the antenna. A research test chip

demonstrating this concept is described in [13]. The second method uses a polar signal representation. A frequency-modulated signal is first amplified with an efficient PA and then an amplitude modulation is added to the signal [14]. The actual effectiveness of these techniques remains to be demonstrated.

IC Technology choices: While CMOS is the universal technology for base-band and MAC processing, the proper technology choice for the radio RFIC depends on the availability to proprietary technology. From a purely technical perspective, it is clear that the challenges of the 802.11a transceiver design justify the use of a high-performance RFIC technology such as bipolar or BiCMOS. However, the consumer driven cost pressures of wireless LANs cannot be avoided. This will produce increasingly aggressive designs emphasizing low cost without compromising performance. Proprietary inexpensive SiGe BiCMOS technology such as in [15] is ideally suited for these developments. In addition, the standard CMOS widely available from foundries has been making considerable progress in RF capabilities driven by a large pool of talented designers with no access to other technologies. The main current limitation of CMOS RFICs for high-speed wireless LANs is more related to power supply voltage scaling rather than inferior transistor performance.

4. Future higher rate systems

Following the trends of wired data communications, wireless LANs are likely to evolve towards even higher rates, i.e., 100 Mb/s, 200 Mb/s, etc. The following question arises naturally: which technical solution would best match this trend? From the facts discussed in this paper it appears that a further increase of the modulation depth, i.e. using 256-QAM, etc., will pose tremendous transceiver implementation problems not easily solvable with inexpensive circuits. From a fundamental point of view, Shannon's famous

channel capacity theorem [16] clearly shows that increasing the information content per unit bandwidth is realized only with an asymptotically exponential SNR increase, which has dramatic cost and power implications in practice.

For instance, since for a static AWGN channel with given *SNR*, the capacity in bits per transmitted symbol is given by $C = log_2(1 + SNR)$, doubling the theoretical limit for the number of bits per symbol from 1 to 2 (e.g., BPSK to QPSK) requires an increase in transmitted energy by a factor of 3. Going from 4 to 8 bits (16-QAM to 256-QAM), brings a 17-fold increase the transmitter output power. For large spectral efficiencies, each additional bit transmitted in the same symbol requires a doubling of the transmit power. Clearly, the common sense compels us to try to transmit twice the number of bits for doubled transmit energy. This is possible but only by exploring other "dimensions" in the communication theory.

4.1 Doubling the Bandwidth

A brute force approach to accomplishing higher rates is by increasing the channel bandwidth. For example, 100 Mb/s could be realized through two present channel transmissions at 54 Mb/s. The current radio system and RFIC technology are suitable to accomplish this but the cost and power dissipation of the new system may not be as attractive as the current generation of wireless LAN products. Generally, there is a vast amount of bandwidth in the 5 GHz ISM band and hence, this approach is interesting due to its relative simplicity.

4.2 Exploiting the Spatial Dimension (MIMO)

A fundamentally different system approach to obtaining higher rates is based on the MIMO concept. MIMO (Multiple-In-Multiple-Out) refers to a system in which there are both multiple transmit antennas transmitting simultaneously in the same bandwidth, and multiple receive antennas used to capture and recover

the data streams and reconstruct the desired information. A 2 x 2 MIMO structure is illustrated in Fig. 11. As demonstrated in [17], the Shannon capacity for MIMO structures is impressive. For a given overall power we can transmit significantly more bits per unit bandwidth than in the traditional single-antenna systems.

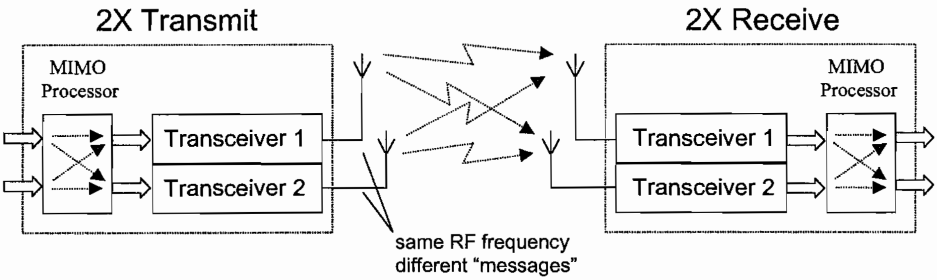

Figure 11: Conceptual 2x2 MIMO system

In general, it can be shown that for an *n x n MIMO system*, there is a *linear* relationship between the overall signal power and the channel capacity (in contrast to the logarithmic dependency in the classical case). However, notice that in order to exploit this concept, it is required that the wireless propagation channel be "favorable". In mathematical terms, this can be expressed by the condition that the channel matrix is of high rank. Only then we have parallel independent data pipes between transmitter and receiver which can be exploited in order to increase the data rate. In a typical indoor-situation, due to the rich multi-path fading environment, this condition is typically fulfilled provided that the antenna outputs are uncorrelated (for instance spaced at least _ to _ wave-length apart).

Note, however, that similarly to the original (one-dimensional) Shannon limit, the MIMO capacity theory gives only upper limits of the achievable

spectral efficiencies, and does not provide any guidelines as to how these limits are approached. Over the past few years, various signaling schemes have been developed for MIMO channels emphasizing various aspects of the new degree of freedom. The most prominent examples include the different variants of BLAST [17] or space-time coding [18], but due to the wide spatial-temporal design possibilities a large number of other advanced techniques have been proposed recently. The performance of MIMO enhancements in 802.11a OFDM WLANs with spatial maximum likelihood (ML) detection has been investigated for varying propagation environments in [19].

Table 1: Achievable spectral efficiencies in (bit/sec)/Hz and data rates in Mb/s for varying number of transmit antennas (i.e. spatial dimension), channel code rates, and number of constellation points (numbers extrapolated based on the IEEE 802.11a OFDM parameters).

Spatial Dim	*Code Rate*	*Num Const Pts*	*Eff b/s/Hz*	*Rate Mb/s*
1	1/2	16	*2.0*	24
1	1/2	64	*3.0*	36
1	3/4	16	*3.0*	36
1	3/4	64	*4.5*	54
2	1/2	16	*4.0*	48
2	1/2	64	*6.0*	72
2	3/4	16	*6.0*	72
2	3/4	64	*9.0*	108
3	1/2	16	*6.0*	72
3	1/2	64	*9.0*	108
3	3/4	16	*9.0*	108
3	3/4	64	*13.5*	162

Table I gives an overview of basic physical layer parameters for single and multiple antenna transmission and corresponding achievable data rates under the assumption that the general OFDM format remains unchanged (i.e., number of OFDM sub-carriers, number of pilot tones, duration of guard intervals etc.).

The columns contain the spatial dimension, i.e., number of transmit antennas, the effective code-rate (ratio of information-carrying bits to transmitted bits), the constellation depth, the resulting raw spectral efficiency in bits/s/Hz, and the achievable data-rate. Note that the spectral efficiency for 100 Mb/s is in the order of 9 bits/s/Hz, which is twice the number at 54 Mb/s and generally very large compared to other existing commercial wireless systems. Also, the system in the third row of Table I has been chosen by the 802.11a standard over the one in the second row due to better robustness under typical channel conditions, despite the fact that both systems provide the same data rate.

However, although it is hard to beat well-designed MIMO codes in terms of spectral efficiency, these schemes come with the burden of increased base-band signal processing needs and increased complexity of the RF circuitry. Moreover, the sensitivity of MIMO to co-channel interference still has to be assessed and may partly reduce the theoretical spectral efficiency gains of this approach.

The debate about next-generation WLAN standards is about to take a more concrete form in the standards bodies, and it will be interesting to see what technological features will eventually play a part in an ultra-fast WLAN air interface.

5. Conclusions

While wireless access to data networks has become a fully accepted capability in everyday life, enabled by the availability of inexpensive transceiver technology, the upcoming speed enhancement to 54 Mb/s through the 802.11a standard encompasses serious design challenges. Power dissipation and cost are the most important final features, of course, assuming a reasonable range performance. Increasing the data rate even further, as will surely be dictated by the market, will set new technical hurdles, which will require innovative system

and circuit solutions. The presentation in this paper has tried to illustrate that close synergy between systems and circuits is a necessary ingredient for future successful designs.

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