Joint Channel and Mismatch Correction for OFDM Reception with Time-interleaved ADCs: Towards Mostly Digital MultiGigabit Transceiver Architectures[1]

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Abstract— **Time-interleaved (TI) analog-to-digital converters (ADCs) are a promising architecture for realizing the high-speed ADCs required to implement "mostly digital" receivers for emerging multiGigabit communication systems. Mismatch between the parallel ADCs comprising a TI-ADC is a fundamental performance bottleneck. While there exist mismatch correction techniques for generic applications of ADC, we illustrate in this paper that mismatch compensation can be subsumed within the overall receiver design for communication applications. We consider an Orthogonal Frequency Division Multiplexing (OFDM) link with TI-ADC used after downconversion. We show that mismatch results in frequency selective interference across subcarriers that can significantly degrade the performance of a standard OFDM receiver. However, this performance degradation can be alleviated significantly by joint channel and mismatch estimation and compensation, leveraging already available training or pilot sequences. This eliminates the necessity for dedicated hardware for mismatch correction. Specifically, we present an algorithm for estimating the channel gains and the mismatch, followed by low-complexity linear equalization to suppress the inter-subcarrier interference resulting from the mismatch. Our simulations show that the error floor due to mismatch-generated interference can be eliminated, permitting bandwidth-efficient operation with large constellations.**

I. Introduction

"Mostly digital" transceiver architectures have been critical to the economies of scale underlying the proliferation of modern cellular and wireless local area network (WLAN) technologies. Key to these architectures is the accurate digital representation of analog signals using Analog-to-Digital Converters (ADCs). The trend of increasing communication speeds towards the multi-Gbps range is driving the need for ADCs with correspondingly higher sampling rates. For instance, the transmit power restrictions in the recently opened UWB and 60 GHz mm-wave bands make it necessary to employ power-efficient signaling strategies, which require sampling rates of multi-GHz for Gbps data rates [1],[2]. The design of ADCs, with a specific focus on their impact on the overall communication system, is therefore a critical issue for such systems.

A popular approach for building high-rate ADCs is to employ a Time-Interleaved (TI) ADC architecture, multiplexing several lower-speed ADCs in parallel [5]. Since these parallel ADCs can operate at lower speeds, it is possible to realize them with low-power architectures such as Successive Approximation [7] or Pipelined [5], instead of incurring the high power consumption of a single flash ADC, which is typically the choice for accurate conversion at high sampling rates. However, the performance of TI-ADCs is limited by the gain and timing mismatches among the parallel ADCs. There are many approaches for mismatch compensation to alleviate this performance degradation, including digital correction at the ADC output, as well as direct adaptation of the ADC hardware. Scaling in digital technologies favors digital correction compared to hardware adjustments [6],[7],[8]. Digital correction techniques can be blind or training-based, with the latter providing better performance, while incurring training overhead [6].

The key idea behind the current paper is that, for TI-ADCs used in a communication receiver, training overhead for ADC calibration can be eliminated by leveraging the available training in the communication protocol. Rather than compensate for mismatch at the ADC output, we view the non-ideal ADC as part of the overall channel impairment, and consider joint mismatch and channel compensation at the output of the demodulator. We illustrate our ideas for Orthogonal Frequency Division Multiplexing (OFDM) over a dispersive channel, and investigate the effect of gain and timing mismatch in TI-ADCs employed after downconversion. We characterize the inter-subcarrier interference caused by mismatch, and demonstrate that uncompensated mismatch leads to an error floor, which can be accurately characterized by a Gaussian approximation. We then consider the problem of training-based estimation of the channel taps and the mismatch. Since optimal joint estimation of channel taps and mismatch is intractable, we propose an iterative algorithm which yields good performance for the typical levels of mismatch that we consider. We then show that zero-forcing equalization is an effective means of suppressing inter-subcarrier interference with reasonably low complexity.

Prior work on using communication training symbols for ADC mismatch correction includes [7], where the pilot signals from the multi-band OFDM proposal for UWB [11] were used to estimate the voltage offset mismatch. However, timing mismatch, which was neglected in [7], becomes more significant at GHz sampling frequencies, due to the quadratic increase in mismatch-induced interference power with sampling rate [13]. Our focus in this paper, therefore, is on gain and timing mismatches.

The rest of the paper is organized as follows. In Section II,

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we provide a simple model of a non-ideal TI-ADC with gain and timing mismatches, and derive a model for the resulting inter-subcarrier interference. Section III provides a Gaussian approximation for the distribution of interference for a large number of parallel ADCs for the TI -ADC. Sections IV and V discuss mismatch estimation and compensation, respectively. Section VI summarizes the performance of the algorithm using numerical simulations. Section VII concludes the paper.

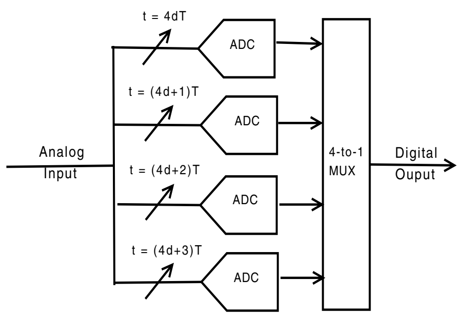

Fig. 1. Ideal time-interleaved ADC (d =integer)

II. System model

In this section we first present a simple model of a non-ideal TI-ADC and then use this model to quantify the effective interference level at the output of the OFDM receiver employing the non-ideal ADC.

A. ADC model

Fig. 1 shows an ideal TI-ADC with 4 parallel ADCs. Each parallel ADC operates at one-fourth the total sampling rate. The ADC outputs are digitally multiplexed to produce the final output. In practice, there may be mismatches between the parallel branches.

In a TI-ADC with L parallel ADCs, let g_i represent the total gain and δ_i represent the timing error of the i^{th} parallel ADC. Ideally, $g_i = 1$ and $\delta_i = 0$. Here, $i = \{0, 1, \cdots, L-1\}$. We use $i \bmod n$ to represent the remainder of i after division by n. Since, the m^{th} sample is taken by the ADC with index $m \bmod L$, the output of the multiplexer can be written as [9]

$$r[m] = g_m r(mT + \delta_m) \tag{1}$$

where T is the sampling interval and (g_m, δ_m) suggests $(g_{m \bmod L}, \delta_{m \bmod L})$.

B. Communication model

The ability of OFDM to collect the multipaths combined with the low cost implementation make it a popular design choice for many wireless channels including the indoor UWB channel [3]. An OFDM transmitter and receiver are shown in Fig. 2. Symbols are encoded onto the frequency domain sub-carriers by using the IFFT operation at the transmitter. This is undone by the use of FFT operation at the receiver. Cyclic prefix is conventionally added before each OFDM frame so as to make the linear convolution between the channel and the transmitted symbols into a circular one. This converts the channel equalization into a scalar equalization, thus reducing the complexity. The received signal corresponding to one OFDM frame after down-conversion and receive filtering is given as

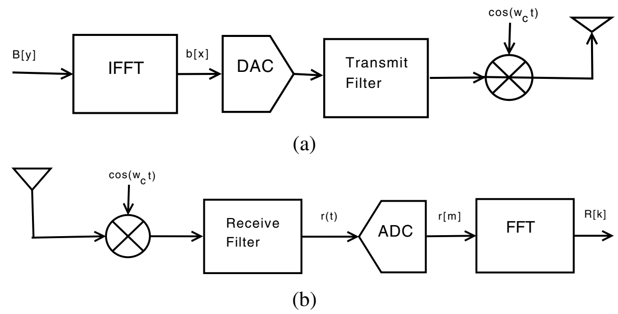

Fig. 2. (a) OFDM Transmitter (b) OFDM Receiver

$$r(t) = \sum_{k=-N}^{M-1} b[k \bmod M] p(t - kT) + \eta(t) \tag{2}$$

where $\{b[x]\}, x = 0, \cdots, M-1$ are the OFDM symbols in time domain, $p(t)$ is the combined impulse response of the transmit, channel and receive filters, $\eta(t)$ represents the noise obtained by passing white noise through the receive filter. Also, $p(t)$ is assumed to have significant values only in $[0, NT]$ and hence, the length of the cyclic prefix is chosen to be N. The time domain symbols $b[x]$ and the frequency domain symbols $B[y]$ are related as $b[x] = \sum_y B[y] e^{j2\pi xy/M}/\sqrt{M}$. This scaling preserves the total average energy before and after the transformation. The symbol energy $\mathbb{E}[|B[y]|^2]$ is denoted as E_s and the noise spectral density by $N_o/2$ in each dimension of the complex plane. Now, samples of (2) are taken at $t = mT$ with $m = \{0, \cdots, M-1\}$. The samples can be obtained using the non-ideal ADC model of (1) in (2). Further, we use the substitution $n = m - k$ to obtain the samples as

$$\begin{aligned} r[m] \quad = \quad & g_m \sum_{n=0}^{N} b[(m-n) \bmod M] p(nT + \delta_m) \\ & + \eta[m] \end{aligned} \tag{3}$$

where the samples of $p(t)$ outside $[0, NT]$ are neglected and the noise samples $\eta[m] = g_m \eta(mT + \delta_m)$ are distributed as zero-mean complex Gaussian with correlation denoted by $R_\eta(m_1, m_2) = \mathbb{E}[\eta(m_1)\eta^H(m_2)]$ given as,

$$\begin{aligned} R_\eta(m_1, m_2) \quad = \quad & N_o g_{m_1} g_{m_2} \times \\ & R_r[(m_1 - m_2)T + \delta_{m_1} - \delta_{m_2}] \end{aligned} \tag{4}$$

where, R_r is the autocorrelation of $p_r(t)$, the receive filter response normalized with $R_r(0) = 1$ so that in the no mismatch case, $R_\eta(m, m) = \mathbb{E}[|\eta(m)|^2] = N_o$.

Equation (3) represents the time-domain system equation. Since, the data symbols in OFDM are encoded onto the frequency domain sub-carriers, we need to derive the frequency domain counter-part of equation (3). First we express

$b[(m-n) \bmod M]$ of (3) in terms of $B[y]$ and then take the FFT of the samples $\{r[m]\}$ to obtain

$$R[k] = \frac{1}{M}\sum_{m=0}^{M-1}\sum_{y=0}^{M-1} g_m B[y] e^{-j2\pi m(k-y)/M} \times \sum_{n=0}^{N} p(nT+\delta_m) e^{-j2\pi ny/M} + \mathcal{N}[k] \quad (5)$$

where $\{R[k]\}$ and $\{\mathcal{N}[k]\}$ represent the FFT (energy preserving) coefficients of $\{r[m]\}$ and $\{\eta[m]\}$ respectively. It can be easily inferred that $\mathcal{N}[k]$ in (5) is distributed as zero-mean complex Gaussian with correlation given by

$$R_{\mathcal{N}}(k_1,k_2) = \frac{1}{M}\sum_{m_1=0}^{M-1}\sum_{m_2=0}^{M-1} R_\eta(m_1,m_2) e^{\frac{-j2\pi(k_1 m_1 - k_2 m_2)}{M}} \quad (6)$$

The summation over n in (5) represents the M-point FFT of the sequence $q[n] = p(nT+\delta_m)$ for $n = \{0,\cdots,N\}$. Since $p(t)$ has significant values only in $[0, NT]$, the sequence $q[n]$ corresponds to the sampled version of continuous time waveform $q(t) = p(t+\delta_m)$. The Fourier transforms of $p(t)$ and $q(t)$ are related as $Q(f) = P(f)e^{j2\pi\delta_m f}$. Due to the availability of guard bands, $p(t)$ is bandlimited to within $(-1/2T, 1/2T)$. This implies that the FFT of $q[n]$ is related to the FFT of $p[n]$ as $Q[y] = P[y]e^{j2\pi\delta_m\tilde{y}/(MT)}$, where $\tilde{y} = y,\ y \le \frac{M}{2}$ and $\tilde{y} = y - M,\ y > \frac{M}{2}$.

Using this result, (5) equals

$$R[k] = \sum_{y=0}^{M-1} B[y]P[y]\Delta_k[y] + \mathcal{N}[k] \quad (7)$$

where

$$\Delta_k[y] = \frac{1}{M}\sum_{m=0}^{M-1} g_m e^{j2\pi\tilde{y}\bar{\delta}_m/M} e^{-j2\pi m(k-y)/M} \quad (8)$$

where $\bar{\delta}_m = \delta_m/T$ represents the normalized timing offset. It can be seen using (8) in (7) that in the absence of mismatches, the model in (7) reduces to the standard equation for each subcarrier, $R[k] = B[k]P[k] + \mathcal{N}[k]$. But, in the presence of the gain and timing mismatches, the subcarriers of the OFDM system interfere with each other decreasing the Signal-to-Interference Ratio (SIR).

III. Gaussian Approximation

To gain more insight into the effect of the mismatch on the system performance, we derive an approximate model for the interference in (7). From (7), we can observe that invoking the Central limit theorem on the interferers needs the assumption of a large number of non-zero $\Delta_k[y]$ and independent frequency domain symbols $\{B(y)\}$. Then, the total interference is approximately distributed as Gaussian and SIR can be used to predict the BER accurately. Assuming zero mean $\{B(y)\}$ and perfect channel knowledge, SIR for the k^{th} sub-channel is given as

$$\mathrm{SIR}(k) = \frac{|\Delta_k(k)P(k)|^2}{\sum_{y \in \mathcal{M}_{\bar{k}}} |\Delta_k(y)P(y)|^2 + (\sigma_k^2/E_s)} \quad (9)$$

where $\mathcal{M}_{\bar{k}}$ is all of $\{0,\cdots,M-1\}$ except k and $\sigma_k^2 = R_{\mathcal{N}}(k,k)$ can be obtained by (6). It can be seen from (9) that for large E_s/σ_k^2, the SIR is completely determined by $\{\Delta_k(y)\}$ implying an error floor. Larger constellations incur a larger BER for a fixed SIR, and hence exhibit a higher error floor.

IV. Estimation of mismatches

Since ignoring the mismatches leads to an error floor, estimation and correction of mismatches become inevitable. In this section, we try to use the channel estimation sequences of the OFDM system to jointly estimate for the channel and the mismatches.

In an OFDM system, significant gains in channel estimation performance have been demonstrated by using the time-domain techniques compared to direct estimation in the frequency-domain where the channel frequency correlation structure is not used [10]. Hence, we consider the joint estimation of channel and ADC mismatches using time-domain samples in (3).

Using the sampling theorem for bandlimited $p(t)$, we have

$$p(t) \approx \sum_{l=0}^{N} p(lT)\mathrm{sinc}(t/T - l) \quad (10)$$

where samples of $p(t)$ outside $[0, NT]$ are neglected. Using (10) in (3), we have

$$\mathbf{r} = B\mathbf{p} + \eta \quad (11)$$

where we denote the receive vector by $\mathbf{r} = \{r[m]\}$, the noise vector by $\eta = \{\eta[m]\}$ and the channel tap vector by $\mathbf{p} = \{p(lT)\}$. Also, the elements of the M-by-$(N+1)$ matrix B are given as

$$B(m,l) = g_m \sum_{n=0}^{N} b[(m-n) \bmod M]\mathrm{sinc}((n-l)+\bar{\delta}_m)$$
$$m = \{0,\cdots,M-1\} \ \&\ l = \{0,\cdots,N\} \quad (12)$$

where $\{b[m]\}$ represents the the time-domain training sequence.

Let $i = 0,\cdots,L-1$ denote the indices of the parallel ADCs in the TI-ADC. Now, we present an iterative algorithm with the k^{th} iteration given as

Step 1 : *Given the mismatch estimates at the* $(k-1)^{th}$ *iteration,* $(\hat{g}_i^{(k-1)}, \hat{\bar{\delta}}_i^{(k-1)})$*, find the ML estimate of the channel taps* $\hat{\mathbf{p}}^{(k)}$.

Given the mismatches, the estimation of $\mathbf{p}$ in (11) is a standard estimation problem with the ML solution given as

$$\hat{\mathbf{p}}^{(k)} = (B^H R_\eta^{-1} B)^{-1}(B^H R_\eta^{-1})\mathbf{r} \quad (13)$$

where R_η, the correlation matrix of the noise samples, is as given in (4).

Step 2 : *Given $\hat{\mathbf{p}}^{(k)}$, find the ML estimates of the mismatches, $(\hat{g}_i^{(k)}, \hat{\overline{\delta}}_i^{(k)})$ $\forall i$.*

Given the channel estimate, it can be seen from (3) that the information about $(g_i, \overline{\delta}_i)$ is contained only in $\mathbf{r_i}$, the set of all $r[m]$ with m mod $L = i$. The ML estimates of the mismatches are obtained as

$$(\hat{g}_i^{(k)}, \hat{\overline{\delta}}_i^{(k)}) = \text{argmin } ||\mathbf{r_i} - B_i \hat{\mathbf{p}}^{(k)}||_\eta^2 \quad (14)$$

where $||x||_\eta^2 = x^H R_\eta^{-1} x$ and B_i is the matrix formed by including only the rows of the form $\{i + 1 + dL\}$ for integer d. In general, a two dimensional search can be performed to find the mismatch estimates in (14). When the mismatches are small, we can use the following iterative search procedure. Using (12) in (14), it can be observed that the objective function is quadratic in g_i and hence an exact minimum can be found for each given $\overline{\delta_i}$. One-dimensional search can be performed to find the optimal value of $\overline{\delta_i}$. This procedure is then repeated for all i.

The iterations are performed for a nominal number of times or till a convergence criterion like $||\hat{\mathbf{p}}^{(k)} - \hat{\mathbf{p}}^{(k+1)}||$ being small is satisfied.

V. Compensation of mismatches

Converting the complex time domain equalization to scalar frequency domain equalization is the most important advantage of an OFDM system. Hence, unlike estimation, the equalization has to be performed in the frequency domain. Equation (7) is the standard Multi-user interference model and vast literature exists on low complexity and high performance equalization [12]. Since the mismatches are typically small, we use the conventional Zero Forcing (ZF) equalizer because of its low complexity. However, when all subcarriers interference with each other, the complexity of the ZF equalizer is $O(M^2)$, which is significant. Fortunately, we can drastically improve this situation by choosing the system parameters such that L divides M.

Consider an alternative representation of equation (8) given by

$$\Delta_k[y] = \frac{1}{M} \sum_{i=0}^{L-1} g_i e^{j2\pi \tilde{y} \overline{\delta}_i / M} \sum_{m \in M_i} e^{-j2\pi m(k-y)/M} \quad (15)$$

where $m \in M_i$ represents all m for which m mod $L = i$. For the special case when L divides M, evaluating the summation over m in (15), we have

$$\Delta_k(y) = \begin{cases} \frac{1}{L} \sum_{i=0}^{L-1} g_i e^{\frac{j2\pi(\tilde{y}\overline{\delta}_i - ik + iy)}{M}}, & y \in \mathcal{Y} \\ 0 & \text{otherwise} \end{cases} \quad (16)$$

where $\mathcal{Y}$ includes all y in $\{0, \cdots, M-1\}$ such that y mod $(M/L) = k$. For each k, there are L values of y for which $\Delta_k(y)$ is non-zero. This implies from (7) that each symbol is interfered by $L-1$ other symbols. The equalization complexity is $O(ML)$ which is small when $L << M$. In OFDM, to exploit the computational power of FFT, M is taken as a power of 2. Hence, we also need to take L as a power of 2.

TABLE I

Mismatch values considered

Level	Type	Values (in %)							
10%	$g-1$	3.3	-1.4	-9.5	-7.1	5.7	9.3	-4.9	1.1
	δ	4.6	-9.6	-1.3	-0.5	2.9	6.9	-0.4	-5.1
5%	$g-1$	0.1	4.5	-1.9	-4.0	4.7	1.8	3.2	1.6
	δ	4.9	-2.6	-3.7	1.2	-0.7	2.5	0.5	1.7
1%	$g-1$	0.6	0.8	-0.7	0.8	0.3	-0.8	-0.4	0.1
	δ	0.9	0.9	-0.7	0.9	0.9	0.0	0.6	-0.7

TABLE II

Dispersive channel taps considered

Re[h]	0.5	-1.8	1.7	-0.6	1.0	0.2	0.4	-0.6	0.3
Im[h]	-0.1	-1.8	3.8	1.8	-0.2	1.5	0.6	-0.1	0.0

VI. Numerical Results

We consider a 128 subcarrier OFDM system with a symbol rate of 512MHz. The number of parallel ADCs in the TI-ADC is assumed to be $L = 8$ with a combined sampling rate of 512MHz. The receive filter is taken to be the ideal low pass filter with two-sided bandwidth of 512MHz. Table I gives the mismatch values considered. By a level of 10%, it means that the corresponding mismatch set is drawn randomly assuming that the mismatches are uniformly distributed in the $\pm 10\%$ range from their nominal values.

Table II gives the relative values of the first few significant channel taps of a bandpass filtered instance drawn from CM 1, the line-of-sight channel model proposed for UWB [3]. The filtering is done in the band $[3.704, 4.216]$GHz which is also the second band of Mode 1 UWB devices [11]. Then, the channel response is sampled at rate 512MHz to yield the tap coefficients. We use this channel for all the illustrations that follow.

Fig. 3 depicts the effect of mismatches on the BER performance for 16 QAM constellation. To consider only the effect of mismatch, we have assumed that the channel taps are known accurately. It can be seen that mismatch levels of 5% and 10% introduce error floors indicating the need for correction. No error floor is observed for 1% mismatch levels till a BER of 10^{-4}. Also shown is the BER calculated using Gaussian Approximation (GA) of interference from Section III. From Section V, the number of interfering symbols for $L = 8$ and $M = 128$ is $L-1 = 7$. It can be seen that GA yields results very close to the simulation results suggesting that the interference is indeed approximately distributed as Gaussian.

In Fig. 4, we illustrate for 16-QAM the performance of the estimation and compensation algorithms given in sections IV and V. We estimate the channel using the 6 OFDM-frame channel estimation sequence given in [3]. The results are averaged over 4 instances of received channel estimation

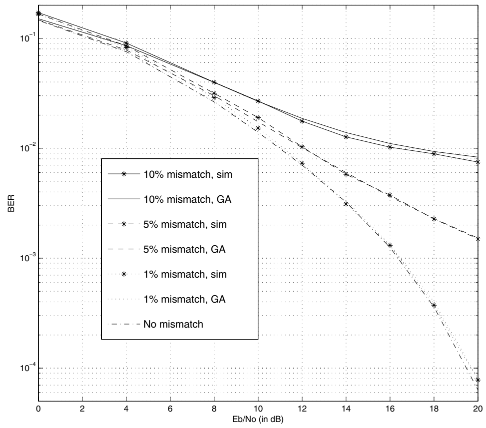

Fig. 3. Effect of mismatch for 16-QAM constellation. Also, shown is the BER calculated using Gaussian Approximation. Perfect channel knowledge is assumed.

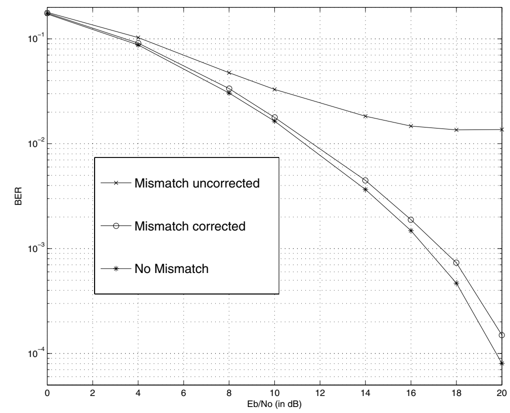

Fig. 4. BER after estimation and correction of mismatches for 16 QAM. Maximum mismatch level considered is 10%.

sequences. For the first iteration, the estimates of $(\hat{g}_i, \hat{\bar{\delta}}_i)$ are taken as $(1, 0)$ for all i. We used four iterations in the estimation algorithm given in Section IV. The post-equalization performance is close to that without mismatch, demonstrating the effectiveness of linear interference suppression for dealing with mismatch. Similar positive results have also been obtained for QPSK. Compared to [8], the proposed approach requires no additional training overhead for mismatch estimation. Also, for the 4-channel TI ADC considered, the proposed mismatch compensation has an effective complexity of a 4-tap filter compared to 21 taps needed in [8].

VII. Conclusion

Our results indicate the feasibility of "mostly digital" communication transceiver architectures even at multiGigabit speeds. With 8 parallel ADCs running at 500 MHz, for example, it is possible to get a sampling rate of 4 Gsamples/sec, which suffices for OFDM transmission spanning a 4 GHz channel. Given the small excess bandwidth required by OFDM, this means that data rates of more than 10 Gbps are achievable using constellations such as 16-QAM. Applying such techniques to the unlicensed 60 GHz band, or for very short-range UWB links (which are more severely power-constrained), implies that the economies of scale associated with digital electronics can be leveraged for proliferation of multiGigabit wireless personal area networks and wireless local area networks. Ongoing research focuses on the use of non-ideal TI-ADCs for "mostly digital" transceivers for singlecarrier modulation over dispersive channels.

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