Title: Submission of Composite Rho Proposal to 3GPP RAN WG4

Source: Hewlett Packard Ltd

Preface

For the purpose of migrating work in progress from SMG2 L1-EG to 3GPP RAN WG4, this paper re-presents material that was previously introduced in ETSI SMG2 L1-EG Tdoc SMG2 L1 622/98[1] and Tdoc SMG2 L1 61/99[2].

NOTE
For the benefit of delegates who have read the earlier documents we point out that the content of the two previous Tdocs has been merged and edited, but there is **no new content** in this paper.

Introduction

Setting levels of Adjacent Channel Protection for the UTRAN directly affects the acceptable level of Adjacent Channel Power (ACP) that may be produced by a transmitter, but before we can talk meaningfully about acceptable levels of ACP from a transmitter we must set some measurable bound on modulation accuracy. Without this additional constraint then it is possible, at least in principle, to construct a signal that will meet an arbitrarily stringent ACP requirement in the frequency domain but would be unacceptably distorted in the code domain.

This paper proposes two measurements of modulation accuracy for the downlink. The first we have called *Composite Rho* because of its evident relationship to the IS-95 *Rho* measurement. The second is called *Composite Rho Estimate*.

HP is currently working on a definition for the *Composite Rho* measurement, which is a direct development of the existing *Rho* metric. We believe it is important to define *Composite Rho* in a theoretically rigorous manner, but our initial work suggests that it will be a computationally intensive measurement and it may transpire that *Composite Rho Estimate,* is the more practical measurement. Although *Composite Rho Estimate* is not a direct measure of modulation quality (it is based on the distribution of power in the code domain), it is a considerably less complex measurement and simulations suggest that it is in many cases a good estimate of the actual *Composite Rho*.

In this document we define *Code Domain Power, Composite Rho Estimate* and, for comparative purposes, we outline the definition of *Composite Rho*. This latter proposal is <u>not in final form</u>, it should be regarded as work in progress.

A descriptive narrative on *Code Domain Power* and *Composite Rho* can be found in Annex 1. A derivation of the relationship between Rho and EVM can be found in Annex 2. An investigation of the statistics of various traffic mixes with a view to defining a test signal can be found in Annex 3. Finally, some results on the code-domain effects of non-linearities can be found in Annex 4.

Preamble

In the following description of the measurement procedures we assume that the measuring instrument has demodulated the received W-CDMA signal to produce a time-record of chip-aligned complex valued samples representing the received $I+jQ$ baseband chip stream. We further assume that this chip stream has been descrambled.

For the less computationally intensive procedure of computing the code domain power coefficients, composite Rho estimate, chip rate EVM and de-spread EVM, the frequency and time alignment is achieved prior to the measurement processes. In the more computationally intensive procedure of computing true *Composite Rho*, frequency and time alignment is acquired as part of the reference signal estimation process.

Definition of Code Domain Power

Code Domain Power is the distribution of signal energy on the set of orthogonal code channels, normalized by the total signal energy. The distribution of energy is calculated for each orthogonal code interval, and then averaged over a number of code intervals. Since the set of orthogonal codes is complete, all of the signal energy projects on the set of code-channels, whether the signal has error or not. [See Annex 1 for a slightly more "friendly" description].

In general the vector (Z) of samples of the received, descrambled chip stream can be regarded as comprising $N \times M$ samples, where N is the number of symbol periods in the measurement interval and M is the spreading factor (M chips per symbol, with one sample per chip).

For the purpose of computing code domain power coefficients a measurement interval of one timeslot (625µs) is proposed. We believe that this makes the measurement applicable to both FDD and TDD modes though this is still being studied. Within this constraint it is evident that the values of N and M depend upon which layer of the OVSF code tree is being evaluated.

For the proposed *Composite Rho Estimate*, the code domain power coefficients are always computed at the C(8) layer so M = 256. Note that with a measurement interval of 625µs, N should, in principle, equal 10. In practice the samples constituting the symbol coincident with SCH1 are excluded and so N equals 9 . This aspect of the proposed measurement would benefit from further discussion.

Code Domain Power Coefficients (ρ_i)

Equation 1 is applied for $i \in \{0,1,2,\ldots,M-1\}$ to generate a vector (ρ) of code domain power coefficients.

$$\rho_i = \frac{1}{\sum_{k=0}^{M-1} \left| R_{i,k} \right|^2} \cdot \frac{\sum_{h=0}^{N-1} \left| \sum_{k=0}^{M-1} Z_{(h \cdot M + k)} \cdot \overline{R_{i,k}} \right|^2}{\sum_{h=0}^{N-1} \sum_{k=0}^{M-1} \left| Z_{(h \cdot M + k)k} \right|^2}$$

Equation 1 - Code Domain Power Coefficient Calculation

In this case the reference signal $R_{i,k} = C_{i,k} + jC_{i,k}$ and $C_{i,k}$ is the k^{th} chip of the i^{th} spreading code (the i^{th} row of the Hadamard matrix). $\overline{R_{i,k}}$ is the complex conjugate. The vector of ρ_i values may be plotted as a histogram to give a display of code domain power.

Since, as was mentioned earlier, the set of OVSF spreading codes is complete, and all the energy must be accounted for, we can state that,

$$\sum_{i=0}^{M-1} \rho_i \equiv 1$$

Equation 2 - Sum of Code Domain Power Coefficients Equality

Definition of Composite Rho Estimate

From a theoretically rigorous perspective, a measurement based on code domain power is a questionable measure of modulation accuracy, since it is insensitive to long term incoherence of the measured signal. For example, as we will see in the comparative section, when the measurements are made over a large number of orthogonal code intervals, *Code Domain Power* (and thus *Composite Rho Estimate*) is relatively insensitive to frequency error or (perhaps more practically) phase jitter, when compared to *Composite Rho*.

This perspective is further commented upon in the conclusions. For the present, we proceed on the assumption that *Composite Rho Estimate* is a useful result.

Given a vector of code domain power coefficients (`codePower[]`), the following code fragment demonstrates how *Composite Rho Estimate* can be computed.

```
usedCodesPower = 0;
codesUsed = 0;
unusedCodesPower = 0;
codesUnused = 0;
usedCodeThreshold = 1 / spreadingFactor;
for (i = 0; i++ ;i<spreadingFactor) {
       if rho[i] > usedCodeThreshold {
              usedCodesPower += codePower[i];
              codesUsed++;
              }
       else {
              unusedCodesPower += codePower[i];
              codesUnused++;
       }
}
averageNoisePower = unusedCodesPower / codesUnused;
compRhoEst = usedCodesPower - usedCodes * averageNoisePower;
```

IS-95-A Rho

Looking at the established definition of the *Rho* modulation accuracy measurement[3][5], we see that a reference signal is established by estimating the parameters of the reference signal to minimize the mean-square-difference between the reference signal and measured signal, over the measurement interval. This definition of *Rho* can be applied to any modulated signal, regardless of whether or not it is a CDMA signal, or whether or not it is a CDMA signal composed of only a pilot or a set of code-channels. The computation of *Rho* can be expressed by Equation 3,

$$\rho = \left[\frac{\left| \sum_{k=0}^{N \cdot M - 1} Z_k \cdot \overline{R_k} \right|^2}{\sum_{k=0}^{N \cdot M - 1} \left| R_k \right|^2 \cdot \sum_{k=0}^{N \cdot M - 1} \left| Z_k \right|^2} \right]$$

Equation 3 - Rho Calculation

Where Z_k is the received signal but R_k is now the estimated ideal reference signal (i.e. the spread data symbols, not just the channel spreading code).

For the "pilot only" case in IS-95-A, the data symbols are all 1 and so the ideal reference signal is constant over the whole measurement interval (R_k = 1+j) and it is possible to simplify the procedure by assuming, rather than detecting, R_k.

However, in general, to compute *Rho* on an arbitrary code channel, it is necessary to know the original symbol stream in order to synthesize the ideal reference signal. This can be achieved by employing a test signal with known data content, or by detecting the symbols in the measuring receiver. The reason for this (in comparison with the *Code Domain Power* calculation) is that the summation in the numerator is performed over whole measurement interval, rather than a single symbol period.

At the most general level, and assuming that the ideal reference signal can be created, then *Composite Rho* is the summation of the individual *Rho* values for the active channels.

Outline Development of Rho to Give Composite Rho

The possible computation of *Composite Rho* in a manner developed from IS-95 *Rho*, including the steps necessary to acquire frequency and timing (which were omitted from the *Composite Rho Estimate* description), is outlined below;

1. Pre-estimate the timing epoch and frequency error.
2. Calculate code domain power and test statistics that are compared to thresholds to detect active code channels.
3. Detect data bits of active code channels.
4. Generate an ideal reference signal composed of the detected, active code channels.
5. Estimate the parameters of the reference signal to minimize the mean-square-difference between the measured signal and the reference signal, using the detected data bits found in step 3. The estimated parameters are (i) timing, (ii) phase, (assuming all code-channels are time and phase aligned, (iii) frequency, and (iv) amplitudes of code channels.)
6. Calculate *EVM*, *Composite Rho* and *Code Domain Power*.

This is clearly a theoretically rigorous, but also computationally intensive definition *Composite Rho*. Estimating the reference for a multi-channel signal may require manipulation of a large parameter matrix; and it does not address, for example, how to identify active code channels in a mixed rate signal. These issues are currently under study.

Comparison of Composite Rho Estimate and Composite Rho

To develop an understanding of *Composite Rho Estimate* as a measurement relative to "true" *Composite Rho*, a simulation of the impact of both Gaussian Noise and Phase Jitter was conducted. In this comparison, *Composite Rho* was calculated based on an exact statement of the ideal signal, not an estimate. The test signal comprised 100 adjacent C(8) codes from C8(16) to C8(115) each carrying randomly generated symbols (an approximation of the test signal suggested in Annex 3). The chip rate EVM was computed using Equation 4,

$$EVM = \sqrt{\frac{\sum_{k=0}^{N\cdot M-1} e_k \cdot \overline{e_k}}{\sum_{k=0}^{N\cdot M-1} R_k \cdot \overline{R_k}}}$$

Equation 4 - General Equation for EVM

A re-statement of *Composite Rho Estimate* and *Composite Rho* as approximate EVM values using Equation 10 (derived later) is included for comparative purposes.

In the presence of Gaussian Noise

A Gaussian interferer was added at level that resulted in approximately 1%, 2%, 5% and 10% EVM as computed on the chip rate signal.

Chip Rate EVM	Composite Rho Estimate	EVM estimate (CRE)	Composite Rho	EVM estimate (CR)
0.986%	0.99990123	0.994%	0.99990623	0.968%
1.989%	0.99960309	1.993%	0.99962143	1.946%
5.002%	0.99749098	5.015%	0.99759949	4.905%
9.925%	0.99015111	9.973%	0.99064393	9.718%

Table 1 - Comparison of Composite Rho Estimate and Composite Rho with Gaussian Noise

In the presence of Phase Jitter

Sinusoidal phase jitter (5º RMS) was added at rates of 150 Hz, 1500Hz and 15kHz which in this case is equivalent to approximately 0.09, 0.9 and 9 cycles per measurement period.

Jitter Freq	Chip Rate EVM	Composite Rho Estimate	EVM estimate (CRE)	Composite Rho	EVM estimate (CR)
150 Hz	3.675%	0.99999462	0.232%	0.99967185	1.812%
1500 Hz	9.113%	0.99973637	1.624%	0.99181491	9.084%
15000 Hz	8.767%	0.9875203	11.242%	0.99235469	8.777%

Table 2 - Comparison of Composite Rho Estimate and Composite Rho with Phase Jitter

It is clear that *Composite Rho Estimate* is not sensitive to low frequency jitter and is degraded by higher frequency jitter. The breakpoint depends on the length of the measurement interval.

It should be noted that while the *Composite Rho Estimate* does not properly detect this impairment, it would be visible as phase jitter in a vector display of the de-spread symbols as could be quantified by a simple EVM-like calculation performed over the de-spread symbols.

Spurious Codes Measurement

There is the possibility that system distortions will lead to unwanted leakage of signal power into a specific unused code or set of codes. See Annex 4 for an example of this effect. The Rho measurement does not explicitly highlight this sort of distortion effect. If the spur codes were relatively low power then the *Composite Rho Estimate* value would be degraded but it would not be apparent that it was due to spurious power growth in a few specific codes. If the spur codes were sufficiently large they could even be conceivably be treated as "used codes" by the *Composite Rho* (and *Estimate*) measurements.

To address this second aspect of modulation quality we propose a test limit for spurious codes. If we use a standard test signal, then we know a-priori which codes should be occupied and it is then possible to set a level a few dB above the expected noise floor. It would not be acceptable for any spurs to exceed this threshold.

The level of this second spurious limit is **for further study**, taking into account the interaction of the code domain noise floor and the downlink dynamic range. Values of -50 dB relative to total signal power, or -30 dB relative to the largest channel power have been suggested.

Conclusions and Further Work

It can be argued that the specific impairments that cause *Composite Rho Estimate* to mis-report the true *Composite Rho* are to some extent pathological, for example, in a real receiver low frequency phase jitter would be tracked.

When this is weighed against the relative complexity of the two measurement approaches it is suggested that *Composite Rho Estimate* is a worthwhile and useful approximation which should be pursued, but not to the exclusion of true *Composite Rho*.

It is also worth recalling that modulation defects that are not properly detected by *Composite Rho Estimate* will be reported by a simple EVM-like calculation performed over the de-spread symbols.

HP are continuing to investigate the relative merits of both measurement approaches. We are, in particular, investigating the robustness of *Composite Rho Estimate* when measuring "Real World" signals, and the computational complexity of *Composite Rho*.

Annex 1 - Overview of Code Domain Power and Composite Rho Estimate

To introduce the basic concept of *Composite Rho Estimate*, we first consider a simplified CDMA system. In this system there are only eight possible spreading codes, and the spreading factor is fixed.

Figure 1. shows an example of the code domain display when three codes are used. Note that in order to introduce the concept clearly, the noise level present in the example signal is exaggerated.

In general, a code domain display is created by correlating the received signal with each of the spreading codes and recording the correlated power for each code as a histogram. Some proportion of the input signal power will correlate with each one of the spreading codes and, since the set of spreading codes is complete, and all the power must be accounted for; the heights of the histogram bars will sum to 1.0.

If the input signal has no structure in the code domain (for example, a CW signal) then the signal will correlate approximately equally with each of the spreading codes and for a system of N codes we will have approximately ($1/N$) of the total signal power in each histogram bar.

In this case, with 8 codes, all the bars would be approximately 0.125 high, relative to a full scale of 1. This, is shown as the dot-dash line in Figure 1.

We note that this serves as a natural "mid-line" for code domain measurement. Any code structure present in the signal will tend to increase the amount of power in the used code bars and consequently decrease the power in the unused code bars, as exemplified by Figure 1.

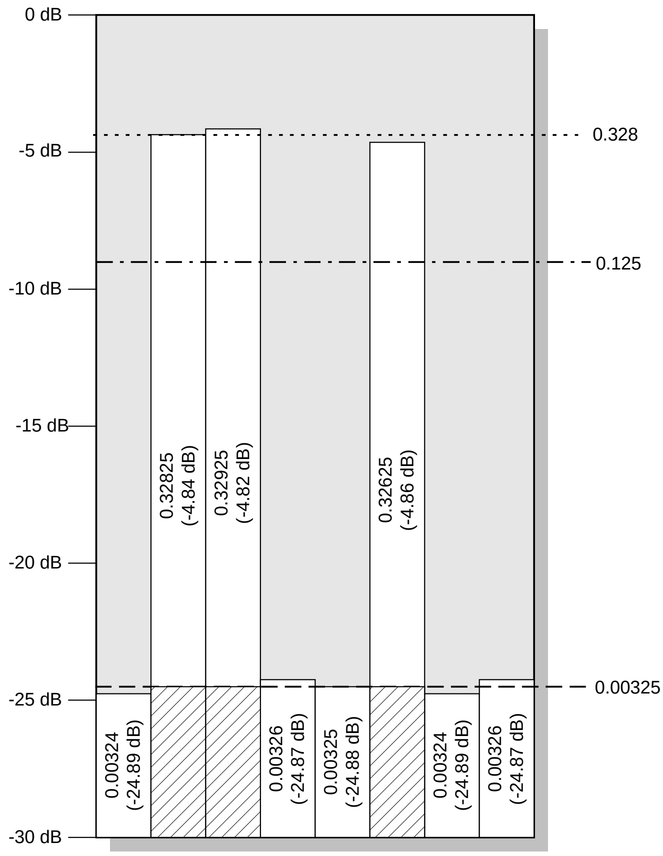

Figure 1 Code Domain Power of Simplified System

The code domain display of an ideal signal would have all the signal power correlating with the used codes and none with the unused codes. In reality, signal and system imperfections will result in a certain amount of the signal power correlating with the unused codes and the received signal will also have a certain noise level which will correlate across all codes.

To compute the *Composite Rho Estimate* measurement, we assume that all codes having a higher correlated power than the $1/N$ level are used codes, while all those having a lesser power are unused codes. We first average the power in all the unused codes (p_{CN}) to obtain an estimate of the average system noise level.

With reference to the simplified system in Figure 1, we have the estimated average noise power in each code (n_C),

$$n_C = \frac{1}{5}\left(p_{C0} + p_{C3} + p_{C4} + p_{C6} + p_{C7}\right) = 0.00325$$

We then subtract this estimated noise level from each of the codes that have been categorized as used codes to get the signal power (p_S) in each used code. The subtracted noise power is shown by the hatched areas in Figure 1.

Again with reference to the example system we have,

$$p_S = \left(p_{C1} + p_{C2} + p_{C5}\right) - 3 \times n_C = 0.974$$

This is then ratioed to the total received power (p_T), which is always 1, giving a composite Rho measurement of 0.974.

$$\rho = \frac{p_S}{p_T} = \frac{0.974}{1.000} = 0.974$$

Consideration of OVSF Spreading

In applying this measurement to the UTRAN we must consider the apparent extra complexity introduced by the use of variable spreading factor codes. In practice this does not present a problem. For the purposes of measuring modulation accuracy the signal is de-spread at the 256 chip spreading code (16 ksps) level regardless of the actual traffic mix. This is legitimate because the power attributable to a traffic channel using a higher rate spreading code will correlate with the block of *M* adjacent codes at the 16 ksps level for which the higher rate code is the parent (*M* is also the ratio between the used spreading code rate and 16 ksps).

Provided that all the codes in this block are identified as used codes then the aggregate power of the *M* codes in the block will equal the signal power of the higher rate code.

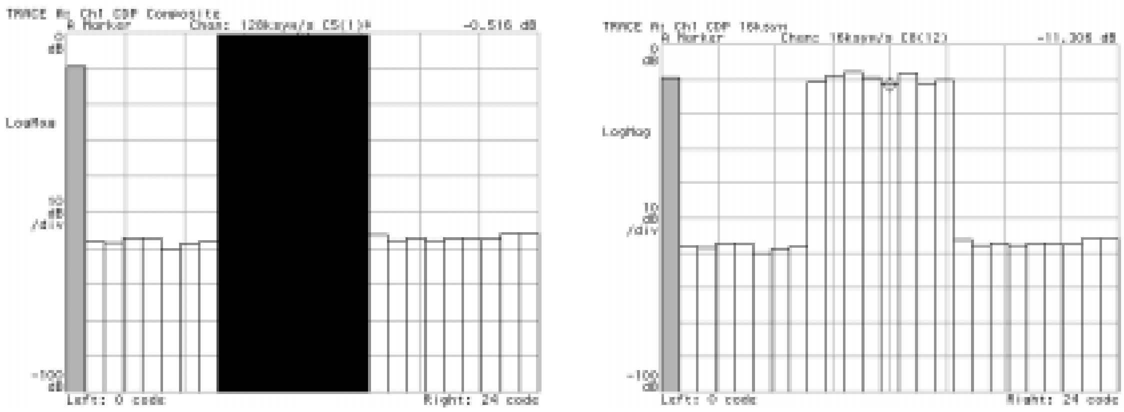

Figure 2 Example of De-spreading a 128 ksps Channel with 16 ksps Spreading Codes

Figure 2 shows an example of this. The (partial) composite code domain display of a 128 ksps channel is shown in the left hand plot. Its relative power is measured as -0.516 dB. The code domain display of the same signal de-spread at the 16 ksps level is shown in the right hand plot. This shows the corresponding block of 8 x 16 ksps codes, the measured relative powers in each code are (from left to right and in linear units);

0.08621301
0.1352388
0.1604759
0.111002
0.07403557
0.1479832
0.0741559
0.09894102

These values sum to give an aggregate relative power of 0.888045, or -0.51565 dB as expected.

The power is distributed reasonably evenly between the 16 ksps codes because of the randomizing effect of the traffic data.

One final point that must be considered to confirm that evaluation of *Composite Rho Estimate* at the 16 ksps level is legitimate, is to look at the used / unused code threshold level. If we further consider the case shown in Figure 2, the threshold - if we were to work at the 128 ksps level - would be 1/32 = 0.03125 = -15.05 dB (about 14.5 dB below the measured code power). At the 16 ksps level, the threshold is 1/256 = 0.0039063 = -24.08 dB which is also about 14.5 dB below the average power level of the 8 codes at 16 ksps (-9.55 dB).

To summarize the *Composite Rho Estimate* measurement algorithm;

1. The received signal is de-spread at the 256 chip spreading code level
2. A threshold level of 1/256 = -24.08 dB is applied to categorize codes as used or unused
3. The power in the unused codes is averaged to give an estimate of the code domain noise floor
4. The code power of the *N* used channels is summed
5. *N* times the estimated noise power is subtracted from this to give the signal power
6. The signal power is ratioed to the total power to give an estimated modulation accuracy measurement
7. This ratio is called *Composite Rho Estimate*.

Figure 3. shows an example of the 16 ksps code domain display of a Sync + 50 x 32 ksps DPCH signal. The measured *Rho Estimate* of this signal was 0.99364.

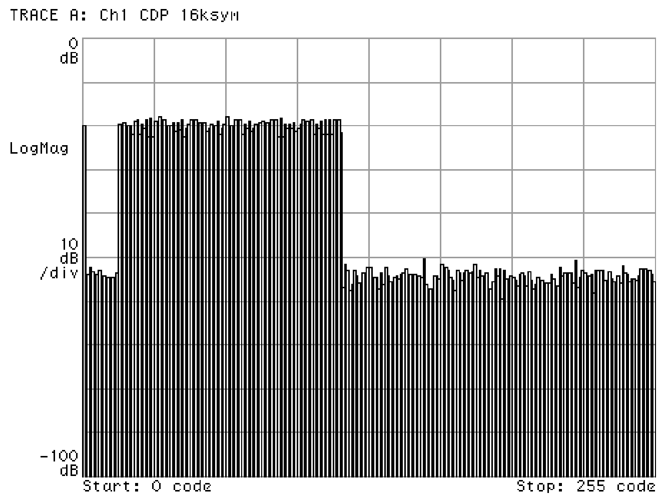

Figure 3 CDP Display of Sync + 50 DPCH

Annex 2 - Relationship Between Rho and EVM of the Spread Signal

Clearly *Error Vector Magnitude*, as a measure of the difference between the intended and actual trajectories of a vector modulated signal, is related to *Rho*. We now consider the relationship between these two measurements.

We start with the expression for *Rho*;

$$\rho = \left[\frac{\left| \sum_{k=0}^{N \cdot M-1} Z_k \cdot \overline{R_k} \right|^2}{\sum_{k=0}^{N \cdot M-1} \left| R_k \right|^2 \cdot \sum_{k=0}^{N \cdot M-1} \left| Z_k \right|^2} \right]$$

Equation 5 - Rho

Then re-express the received signal as a transmit signal plus an error signal,

$$\rho = \left[\frac{\left| \sum_{k=0}^{N \cdot M-1} (R_k + e_k) \cdot \overline{R_k} \right|^2}{\sum_{k=0}^{N \cdot M-1} R_k \cdot \overline{R_k} \cdot \sum_{k=0}^{N \cdot M-1} (R_k + e_k)\overline{(R_k + e_k)}} \right]$$

Equation 6 - Rho with the Received Signal Re-expressed as Transmit Signal plus Error Signal

Expand Equation 6,

$$\rho = \left[\frac{\left| \sum_{k=0}^{N \cdot M-1} R_k \cdot \overline{R_k} + \sum_{k=0}^{N \cdot M-1} e_k \cdot \overline{R_k} \right|^2}{\sum_{k=0}^{N \cdot M-1} R_k \cdot \overline{R_k} \cdot \left(\sum_{k=0}^{N \cdot M-1} R_k \cdot \overline{R_k} + \sum_{k=0}^{N \cdot M-1} R_k \cdot \overline{e_k} + \sum_{k=0}^{N \cdot M-1} \overline{R_k} \cdot e_k + \sum_{k=0}^{N \cdot M-1} e_k \cdot \overline{e_k} \right)} \right]$$

Equation 7 - Expansion of Equation 6

Now, if the ideal signal (R) and the error term (e) are un-correlated then we can simplify this expression by removing the sums of product terms representing the R to e correlation, and arrive at,

$$\rho \approx \left[\frac{\left| \sum_{k=0}^{N \cdot M-1} R_k \cdot \overline{R_k} \right|^2}{\sum_{k=0}^{N \cdot M-1} R_k \cdot \overline{R_k} \cdot \left(\sum_{k=0}^{N \cdot M-1} R_k \cdot \overline{R_k} + \sum_{k=0}^{N \cdot M-1} e_k \cdot \overline{e_k} \right)} \right]$$

Equation 8 - Approximate Expression for Rho

$$\rho \approx \frac{1}{1 + \frac{\sum_{k=0}^{N \cdot M - 1} e_k \cdot \overline{e_k}}{\sum_{k=0}^{N \cdot M - 1} R_k \cdot \overline{R_k}}}$$

Equation 9 - Re-arrangement of Equation 8 to Emphasize Relationship to EVM

By comparing Equation 9 with Equation 4, we can conclude that the *EVM* of the chip rate signal is approximately related to *Rho* (assuming that the error is un-correlated to the signal), by Equation 10.

$$\rho \approx \frac{1}{1 + EVM^2}$$

Equation 10 - Approximate Relationship Between Rho and EVM

Annex 3 - Composite Rho / Spur Code Measurement Test Signal

It is probably appropriate to define a standard test signal for use with these measurements in evaluating the modulation accuracy of a W-CDMA downlink transmitter.

The test signal should exercise the equipment under test, and it should as far as possible be representative of real-world signals. It should not be an unrealistically difficult signal.

To illuminate this last point, consider the case of a signal comprising 64 x 32 ksps traffic channels, where all the traffic channels have been encoded with the same T_{slot} assignment (e.g. zero). Such a signal can be legitimately constructed, but it has an extremely high peak / average power ratio (19.1 dB) as demonstrated by the CCDF plot in Figure 4. We do not anticipate this being a normal mode of operation, and so argue that this is an unrealistically demanding signal.

To gain some insight into what might constitute an appropriate test signal we have investigated the peak to average ratio (at 0.01% probability) for a number of traffic scenarios.

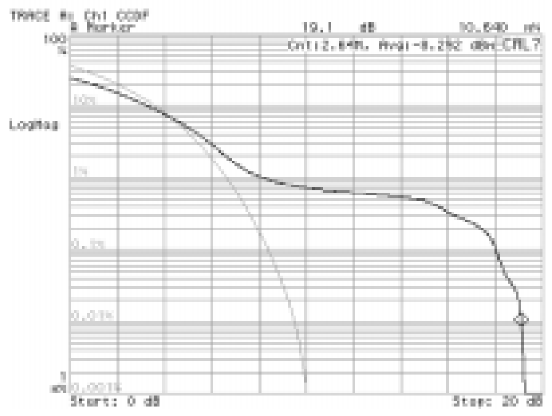

Figure 4 CCDF of Signal With T_{slot} = 0 for all Codes

Signal Generation and Measurement

For practical reasons we used signals generated according to the W-CDMA experimental system (NTT DoCoMo) definition, thus the synchronization channel was implemented as a "Perch 1" rather a SCH1 but as we were only investigating the Complimentary Cumulative Distribution Function (CCDF) of the signal, this difference is not considered to be significant.

The signals were generated using an HP ESG-D RF Signal Generator (option H97) and measured using both an HP89441A Vector Signal Analyzer and a Boonton Model 4500A Peak Power Meter / Analyzer. The two measuring devices were generally found to be in close agreement with the HP89441A giving results typically around 0.1 ~ 0.3 dB higher than the Boonton unit.

The results as presented here are rounded to the nearest 0.5 dB because experimental results suggest that the peak / average statistics of a given traffic mix are subject to a data dependent variation of approximately ±0.25 dB.

Test Results

Clearly there is an infinity of possible traffic mixes. The following scenarios were built according to the following principles,

- All signals include a Perch1 style 16 ksps synchronization channel on spreading code 0.
- 50% code space occupancy (except where higher loadings are investigated)
- Simple sequential assignment of non-zero T_{slot} values. Specifically, we repeat the assignment sequence $m \times \langle 0,9,1,8,2,7,3,6,4,5 \rangle$ where $m \in \{2,4,8,16,32,64\}$ for code rates 32 ksps to 1024 ksps respectively.
- Equal energy per bit power assignments

We have investigated representative examples of ;

- spreading codes allocated as a contiguous block
- spreading codes allocated at random across the whole code space
- single rate, two rate and multi-rate traffic mixes

– higher code space loadings than the nominal 50%

Single Rate Traffic (50% Code Usage)

The peak / average power ratio (0.01% probability) was recorded for each of the possible traffic channel symbol rates. In all cases 50% of the code space was used . This usage factor was chosen based on the observations that IS-95 CDMA typically achieves around 25% code usage and that W-CDMA is expected to improve upon IS-95 performance. Some other loading factor may be more appropriate, opinions vary.

Two code assignment models were investigated; the first allocated spreading codes adjacent to each other in a single block, the second allocated the spreading codes randomly across the code space. The actual spreading code assignments are tabulated in at the end of this Annex.

The results are given in Table 3. They show that there is clearly some advantage in employing random spreading code assignments, particularly when there are a lot of spreading codes in use.

Symbol Rate	No. Codes Used	Pk/Avg Power for Blocked Code Assignments	Pk / Avg Power for Random Code Assignments
1024 ksps	2	6	6
512 ksps	4	8.5	7.5
256 ksps	8	9.5	8.5
128 ksps	16	10	9.5
64 ksps	32	10.5	10
32 ksps	64	12	11

Table 3 Single Rate Traffic (50% Code Usage)

Single Rate Traffic (>50% Code Usage)

Noting from the first batch of single rate experiments that it appears to be the number of codes being used, rather than the proportion of the code space occupied that adversely affects the peak / average power ratio, and that a block code assignment is worse than a random code assignment, we investigated higher loadings of 32 ksps traffic channels.

The results are given in Table 4. They generally confirm that the trend of worsening peak / average power ratio with increasing numbers of traffic channels continues. Note also that the 14 dB peak / average (0.01% probability) was the worst case signal found in this investigation.

Symbol Rate	No. Codes Used	Pk/Avg Power for Blocked Code Assignments	Pk / Avg Power for Random Code Assignments
32 ksps	76 (60%)	12	not investigated
32 ksps	89 (70%)	12.5	
32 ksps	102 (80%)	13	
32 ksps	114 (90%)	13.5	
32 ksps	127 (100%)	14	

Table 4 Single Rate Traffic (> 50% Code Usage)

Dual-Rate Traffic

In these signals we are modeling a mixed voice / data traffic with various rates of data service. Again we have investigated both block assignment and random assignment of codes.

The results are given in Table 5. It is interesting to note that with an even mix of data rates and random spreading code assignments, the CCDF is approximately Gaussian.

Symbol Rates	No. Codes Used	Pk/Avg Power for Blocked Code Assignments	Pk / Avg Power for Random Code Assignments
32 + 64 ksps	32 + 16	10	10
32 + 128 ksps	32 + 8	10.5	10
32 + 256 ksps	32 + 4	10	10

32 + 512 ksps	32 + 2	10	10
32 + 1024 ksps	32 + 1	9.5	9.5

Table 5 Dual-Rate Traffic

Multi-Rate Traffic

Finally, we have modeled a mix of data services at different rates together with a significant loading of voice traffic. This signal occupies 87% of the code space.

The results are given in Table 6. It appears that the inclusion of higher rate data services actually improves the peak / average power ratio. Table 3 shows that 64 voice channels alone has a peak / average ratio approximately 1 dB higher than this case.

Symbol Rates	No. Codes Used	Pk/Avg Power for Blocked Code Assignments	Pk / Avg Power for Random Code Assignments
32 ksps + 64 ksps + 128 ksps + 256 ksps	63 + 8 + 4 + 2	11	not investigated

Table 6 Multi-Rate Traffic

Conclusion

The above experiments represent a very small data set and there is some risk in drawing strong conclusions from these findings. With that note of caution we make the following observations;

Single rate traffic has a higher peak / average than a mixed rate signal of similar code occupancy.

Single rate traffic assigned spreading codes in blocks has a higher peak / average than a signal with random code assignment.

Contiguous blocks of code channels at the same symbol rate constitute a stressful yet realistic signal.

A figure of around 12 dB for peak / average power ratio (at 0.01% probability) seems to encompass all the traffic mixes investigated here, except the >50% code occupancy cases.

Considering these observations we suggest that a signal comprising a sync channel plus a block of 50 x 32 ksps adjacent traffic channels assigned to spreading codes 8 through to 57 may serve as a suitable test stimulus for standardized *Composite Rho*, *Composite Rho Estimate* and *Spurious Code* measurements. A loading of 50 traffic channels gives a peak / average power ratio of approximately 12 dB.

Note that these signals also contain significantly larger peaks, but with lower probability. For example the suggested test signal has a peak / average of about 14 dB at the 0.001% probability level.

Appendix

The following tables show the spreading code assignments used in the experiments.

Single Rate Traffic			
Symbol Rate	**No. Codes Used**	**Blocked Code Assignments**	**Random Code Assignments**
Sync + 1024 ksps	1 + 2	0 1 to 2	0 1, 3
512 ksps	4	2 to 5	2, 3, 6, 7
256 ksps	8	4 to 11	1, 3, 4, 7, 10, 11, 13, 15
128 ksps	16	8 to 23	1, 4, 8, 9, 11, 12, 16, 17, 18, 19, 21, 24, 26, 27, 28, 30
64 ksps	32	8 to 39	1, 3, 4, 7, 8, 9, 11, 14, 18, 19, 22, 23, 24, 26, 27, 28, 32, 36, 37, 41, 43, 46, 47, 49,

			50, 51, 52, 56, 57, 59, 60, 61
32 ksps	64	8 to 71	3, 4, 5, 6, 7, 8, 12, 14, 15, 16, 19, 20, 21, 23, 27, 31, 33, 34, 37, 38, 40, 42, 43, 44, 45, 46, 47, 48, 49, 52, 62, 63, 64, 66, 67, 69, 72, 78, 80, 82, 83, 89, 91, 92, 95, 96, 97, 98, 101, 102, 103, 104, 105, 106, 109, 110, 112, 114, 118, 121, 122, 124, 126, 127

Dual Rate Traffic			
Symbol Rates	**No. Codes Used**	**Block Code Assignments**	**Random Code Assignments**
32 ksps + 64 ksps	32 + 16	8 to 39, 32 to 47	1, 3, 4, 7, 8, 9, 11, 14, 18, 19, 22, 23, 24, 26, 27, 28, 32, 36, 37, 41, 43, 46, 47, 49, 50, 51, 52, 56, 57, 59, 60, 61 and 33, 36, 40, 41, 43, 44, 48, 49, 50, 51, 53, 56, 58, 59, 60, 62
32 ksps + 128 ksps	32 + 8	8 to 39, 16 to 23	1, 3, 4, 7, 8, 9, 11, 14, 18, 19, 22, 23, 24, 26, 27, 28, 32, 36, 37, 41, 43, 46, 47, 49, 50, 51, 52, 56, 57, 59, 60, 61 and 17, 19, 20, 23, 26, 27, 29, 31
32 ksps + 256 ksps	32 + 4	8 to 39, 8 to 11	1, 3, 4, 7, 8, 9, 11, 14, 18, 19, 22, 23, 24, 26, 27, 28, 32, 36, 37, 41, 43, 46, 47, 49, 50, 51, 52, 56, 57, 59, 60, 61 and 10, 11, 14, 15
32 ksps + 512 ksps	32 + 2	8 to 39, 4 to 5	1, 3, 4, 7, 8, 9, 11, 14, 18, 19, 22, 23, 24, 26, 27, 28, 32, 36, 37, 41, 43, 46, 47, 49, 50, 51, 52, 56, 57, 59, 60, 61 and 5, 7
32 ksps + 1024 ksps	32 + 1	8 to 39, 2	1, 3, 4, 7, 8, 9, 11, 14, 18, 19, 22, 23, 24, 26, 27, 28, 32, 36, 37, 41, 43, 46, 47, 49, 50, 51, 52, 56, 57, 59, 60, 61 and 3

Multi Rate Traffic			
Symbol Rates	**No. Codes Used**	**Block Code Assignments**	**Random Code Assignments**
32 ksps + 64 ksps + 128 ksps + 256 ksps	63 + 8 + 4 + 2	1 to 63, 32 to 39, 20 to 23, 12 to 13	not investigated

Annex 4 - Effect of Non-Linearity on Code Domain

In considering the case for a spurious codes limit we briefly discussed the fact that system distortions will lead to unwanted leakage of signal power into a specific unused code or set of codes. If the spur codes are relatively low power then the *Composite Rho* (and *Estimate*) value would be degraded but it would not be apparent that it was due to spurious power growth in a few specific codes. This led us to propose a spur code limit.

A recent paper[4] argues that for a given set of codes, PA non-linearity will cause error energy to be mapped into other codes in a predictable manner. As an example of this effect, we investigated a simple case where a signal comprising two codes (in this case C7(8) and C7(14)) was passed through a deliberately overdriven PA. The gain compression is evident from Figure 5 which shows the signal CCDF before and after the amplifier.

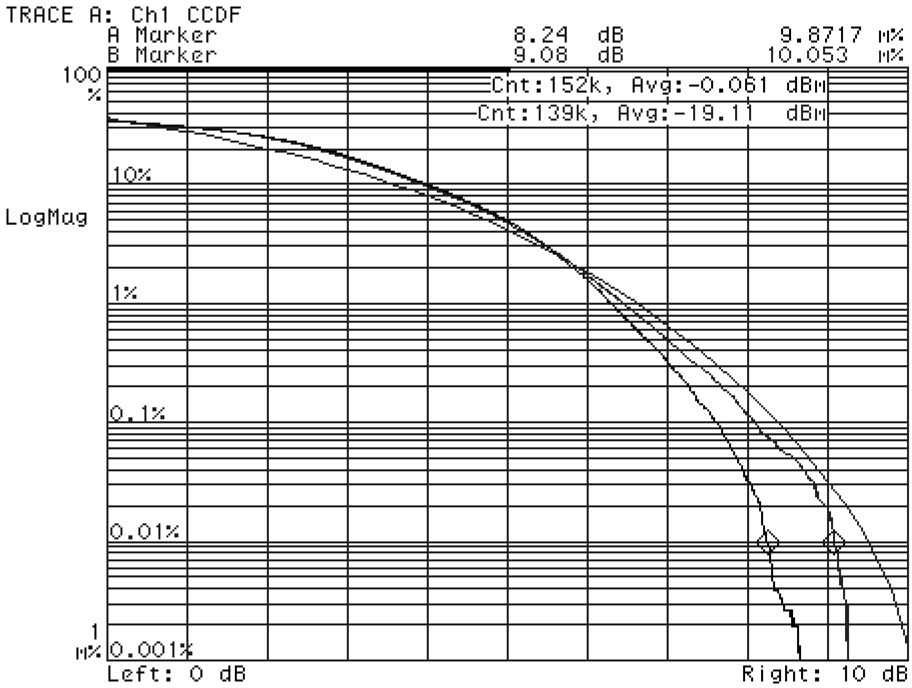

Figure 5 - CCDF Before and After PA

The (partial) Code Domain Display of the resulting PA output signal is shown in Figure 6.

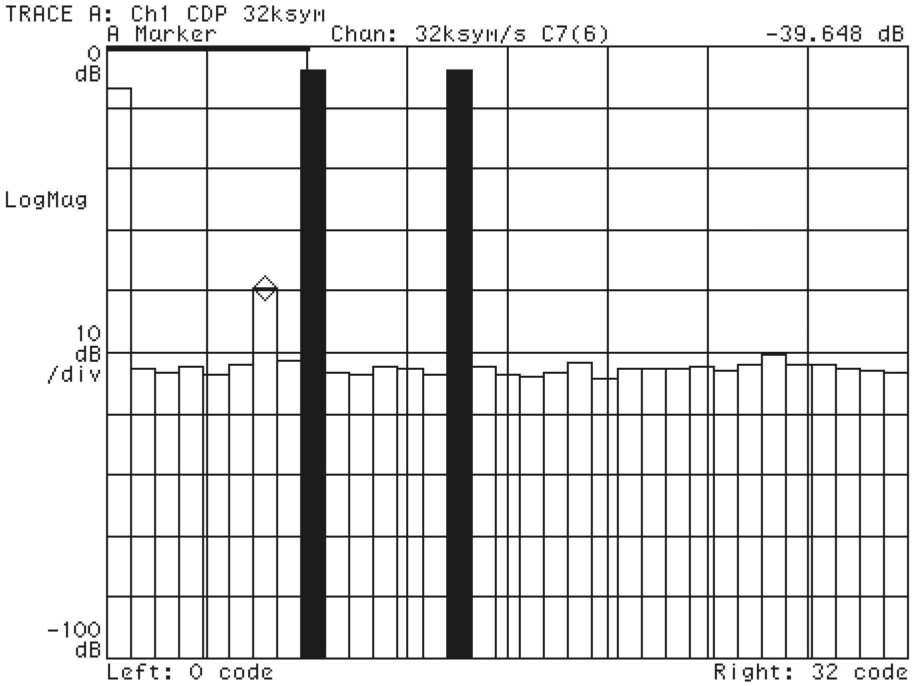

Figure 6 - Code Domain Display of Distorted Signal

It can be seen from Figure 6 that there is significant error energy mapped into C7(6). This code is the exclusive-OR of codes C7(8) and C7(14) as predicted by [4].

Annex 4 - References

[1] Hewlett Packard, "Downlink Modulation Quality Measurement", SMG2 UMTS L1-EG Tdoc 622/98.

[2] Hewlett Packard, "Further Work on Modulation Quality Measurement", SMG2 UMTS L1-EG Tdoc 61/99.

[3] R.A. Birgenheier, "Overview of code-domain power, timing, and phase measurements," HP Journal, pp. 73-93, Feb. 1996. http://www.hp.com/hpj/feb96/fb96a9.pdf

[4] R. N. Braithwaite, "Nonlinear Amplification of CDMA Waveforms: An Analysis of Power Amplifier Gain Errors and Spectral Regrowth", Proceedings of the 48th IEEE Vehicular Technology Conference, pp. 2160-2166, 1998.

[5] "Recommended Minimum Performance Standards for Base Stations Supporting Dual-Mode Wideband Spread Spectrum Cellular Mobile Stations", TIA/EIA IS-97-A, Chap. 12 pp 5-14, July 1996.