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1. Summary

TR 25.85 [9] offers an initial description (based on [7]) of an uplink channel quality measurement based on the highest rate Transport Format Resource Combination (TFRC) that can be received under current channel conditions. This contribution shows that while the measurement report described in TR 25.858 offers a useful basis for progress, further examination of the underlying channel quality indicator used to drive the required TFRC mapping indicates that measurement report could be simplified to eliminate the private UE-specific TFRC mapping function, the reference power offset P_{hs} and the $BLER_{threshold}$ parameter. Importantly, 'advanced' receivers can still be accommodated by the simplified scheme, which also addresses some of the concerns expressed in [8] concerning the definition, openness and testability of the uplink measurement report. A specific text proposal is made for incorporation into TR 25.858.

2. Introduction

A series of contributions ([1]-[6]) has outlined the potential of combining a UE-reported C/I metric with downlink associated DPCH transmit power control (TPC) as a means of optimizing the Node-B scheduling function. In summary, these contributions suggest that UE C/I reporting metric rate should be variable (under RNC control), and that UE's in the DPCH soft-handoff (SHO) state should offer C/I reports at a relatively high rate, while UE's in the non-SHO state should report less frequently and permit the MAC-hs function to use downlink DPCH TPC to compensate for C/I reporting delay and to optimize HS-DSCH scheduling between C/I reports. Notably, [5]-[6] observed that a 4-bit uniform quantization of the C/I metric should be sufficient.

In response to the requirement for a C/I report, [7] proposed the scheme illustrated in Figure 1. As described in TR 25.858, under this proposal the UE is required to:

1. measure a downlink channel quality indicator[1] from the P-CPICH (alternatively S-CPICH),
2. take into account P_{hs} and $BLER_{threshold}$ – respectively, an *a-priori* signaled reference ratio of the power transmitted by the Node-B on the P-CPICH (or S-CPICH) and HS-PDSCH, and an *a-priori* signaled required block error rate for the HS-DSCH transport channel – to look up a private and UE-specific table of receiver performance data based on an *a-priori* specified set of TFRC's, and

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[1] *The precise nature of the downlink channel quality indicator has not yet been defined, and is presumably at the discretion of the UE manufacturer.*

3. report the maximum-rate TFRC that can be received at the specified $BLER_{threshold}$, or alternatively, a power-offset for the HS-PDSCH above the reference power level P_{hs} at which the lowest *a-priori* specified TFRC can be received at the specified $BLER_{threshold}$.

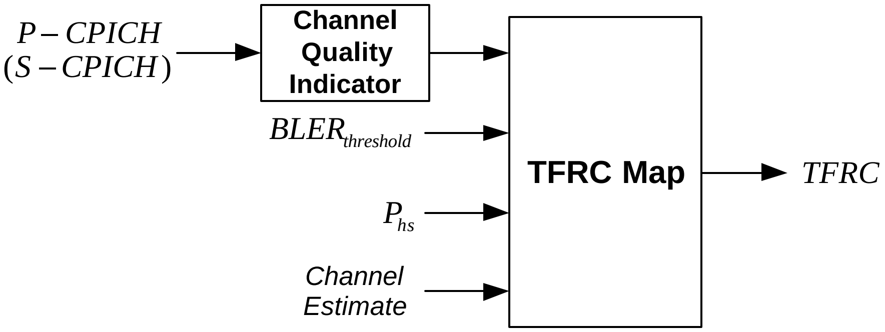

Figure 1. TFRC mapping and reporting process of [7].

While this proposal has been helpful in making progress on the topic of C/I reporting, there are a number of aspects to it which seem unclear or potentially problematic. Many of these issues are outlined in [8] (e.g. the use of non-open and UE-specific receiver performance data to perform mapping table lookup), other aspects include:

a) the P(S)-CPICH-derived “channel quality indicator” is not defined. It is not completely clear, for example, whether the multipath channel estimate or estimated Doppler frequency are required as separate inputs to the TFRC mapping process of Figure 1,[2]

b) the proposed TFRC mapping table supports only up to 5 HS-PDSCH codes, and does not address the requirement to support, for example, 10 or 15 codes (both values are defined in the UE reference capability combination discussion),

c) as pointed out in Section 2.3 of [5] (in reference to the proposal that no C/I report should be transmitted by the UE,) the association of the DPCH TPC data with the UE-reported TFRC may not be obvious to the Node-B, since the mapping table used by the UE as the basis for the report is UE-specific and not observable by the Node-B.

The primary benefit of the TFRC-based proposal, however, is that it offers a potential means of dealing with differing UE performance levels associated with so-called ‘advanced’ receivers (taken here to mean any receiver not employing a RAKE architecture).

3. Channel Quality Indicator Definition

A further examination of the Channel Quality Indicator (CQI) function required in Figure 1 suggests that a simpler approach to UE measurement reporting might be possible, depending on the definition of the CQI.

Leaving aside second-order effects such as interleaver gain (which would, in any case, be limited by the 2ms HSDPA TTI), the effect of Doppler frequency on channel estimate SNR etc., the primary determinant

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[2] *Presumably, any such requirement would depend on the performance requirements agreed by RAN-WG4.*

of HS-DSCH link performance appears to be the signal-noise ratio at the output of the UE despreaders processing the received HS-PDSCH code set. This seems, therefore, to be a reasonable basis for the definition of the CQI function in Figure 1.

Section 7.1 of Appendix A below describes in detail how – in this case for the RAKE receiver – the HS-PDSCH despreader output SNR can be estimated based solely on observations of the received P-CPICH (or S-CPICH) signal. In summary, the UE forms a single multipath-combined estimate p_1 of the P-CPICH pilot symbol process (equation (2.6)), and then – by forming simple first and second-order statistics – estimates the associated signal-noise ratio SNR_{CPICH} of p_1 according to equation (2.13). Next, as discussed in Section 7.1.2, this metric can be used to estimate the desired HS-PDSCH despreader output $SNR_{HS-PDSCH}$ according to:

$$\frac{SNR_{HS-PDSCH}}{SNR_{CPICH}} = \frac{E_{cd}}{E_{cp}} \frac{s_{p_1}^2}{s_{d_m}^2} = \frac{E_{cd}}{E_{cp}} \frac{C_d}{C_p} \tag{1.1}$$

where $C_d = 16$ is the HS-PDSCH OVSF code length, $C_p = 256$ is the CPICH OVSF code length, E_{cd} is the received energy per chip per code for the HS-PDSCH, and E_{cp} is the received energy per chip for the CPICH (primary or secondary).

4. Discussion

Under the procedure of Figure 1, the UE would use the resulting $SNR_{HS-PDSCH}$ estimate in combination with the UE's internal performance lookup table, plus the parameters P_{hs} and $BLER_{threshold}$ to determine the reported TFRC.[3]

It is obvious, however, that the Node-B has knowledge of the constant ratio C_d / C_p, and so given a reported estimate of SNR_{CPICH} from the UE, the Node-B can readily predict the resulting HS-PDSCH despreader output $SNR_{HS-PDSCH}$ for any ratio E_{cd} / E_{cp} of HS-PDSCH code power to CPICH code power selected by the Node-B. Accordingly, at least for the RAKE receiver, direct reporting of SNR_{CPICH} offers a simple and direct method for the Node-B to predict receiver performance based on a hypothesized allocation of number of codes and per-UE HS-PDSCH code power allocation.

A key question is whether the same approach can be applied to advanced receiver architectures. In addition to the RAKE receiver, two classes of advanced receivers have been discussed in WG1 in the context of HSDPA – specifically, equalizer-based architectures [10] and parallel interference cancellers such as MPIC [11].

Section 7.2 shows that a relationship similar to that of equation (1.1) between $SNR_{HS-PDSCH}$ and SNR_{CPICH} can be defined for chip-rate equalizer receivers. In this instance, it is worth noting that – for the same environment that a RAKE receiver might observe – one would expect the measured SNR_{CPICH} to be greater in the case of the equalizer (provided I_{or} / I_{oc} was sufficiently large), and therefore the $SNR_{HS-PDSCH}$ estimate would be larger for a given Node-B selected ratio E_{cd} / E_{cp}. Nevertheless – at least for this

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[3] *Note that Section 7.1.4 discusses the computational complexity associated with generating* SNR_{CPICH} *; it can be seen to be low and unlikely to present implementation difficulties to HSDPA-capable UE's.*

particular 'advanced' receiver architecture – the Node-B can infer the $SNR_{HS-PDSCH}$ given an estimate of SNR_{CPICH} using the same relationship (equation (1.1)) as for the standard RAKE receiver.

The non-linear architecture of the MPIC receiver makes it more difficult to prove that a relationship such as that of equation (1.1) is applicable. However, analysis of the MPIC architecture outlined in Figure 2 of [11] (especially, assessment of the channel estimation function of element 'h' of that figure) suggests that formulation of a post-cancellation combined pilot statistic p_1 analogous to equation (2.6) is feasible, and that this would permit direct estimation of the HS-PDSCH output SNR post-cancellation.

5. Conclusions

If the channel quality indicator is defined in an appropriate way, it seems possible to simplify the TFCI-based UE measurement report to permit direct reporting of the SNR_{CPICH} metric in such a way that advanced receivers can dealt with.[4]

This approach would have the following benefits:

a) the quality measurement SNR_{CPICH} is an 'open' and well-defined statistic,

b) SNR_{CPICH} is commonly applicable to RAKE and advanced receivers of the types so far discussed in RAN1 with respect to HSDPA,

c) the Node-B retains 'responsibility' for predicting receiver performance for a given HS-PDSCH resource allocation (code set, power level),

d) there is a clear relationship between the Node-B utilization of the UE measurement report and the DPCH TPC information,

e) the need for generation and storage by the UE manufacturer of non-standard internal performance lookup tables for TFRC selection can be avoided,

f) the need to report performance for a particular $BLER_{threshold}$ can also be avoided, and

g) no particular restrictions need be defined on the number of codes supported.

Proposed text for incorporation into TS25.858 is included in Appendix B below.

6. References

[1] Sony, "Delay on control information for HS-DSCH", TSGR1#17(00)1378, Stockholm, Sweden, November 21-24, 2000

[2] Sony, "Use of TPC for DL channel quality estimation", TSGR1#18(01)0074, Boston, U.S.A, January 15-18, 2001

[3] Sony, "Reduction of DL channel quality feedback rate for HSDPA (revision of R1-01-231)", TSGR1#19(01)0338, Las Vegas, U.S.A, February 27-March 2, 2001

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[4] *Note that at least one proposal [12] on HSDPA receiver performance assessment in RAN4 has suggested that RAKE receivers be used as the basis for testing.*

[4] Sony, “Follow up on variable C/I feedback rate proposal (rev. of R1-01-0512)”, TSGR1#20(01)0643, Busan, Korea, May 21-25, 2001

[5] Sony, “Downlink channel quality estimation for HS-DSCH”, (HSDPA Ad-Hoc) TSGR1(01)0693, Espoo, Finland, June 26-28, 2001

[6] Sony, “Influence of C/I quantization on HSDPA system throughput”, (HSDPA Ad-Hoc) TSGR1(01)0694, Espoo, Finland, June 26-28, 2001

[7] Nokia, “UL signaling for TFRC selection”, (HSDPA Ad-Hoc) TSGR1(01)1024, Sophia Antipolis, France, November 5-7, 2001

[8] Vodafone Group Plc., “Questions on UL signaling TFRC selection”, TSGR1#22(01)1261, Jeju, Korea, November 19-23, 2001

[9] 3GPP TR 25.858v1.0.0 (2001-12), “High Speed Downlink Packet Access: Physical Layer Aspects”

[10] Motorola, “Pilot power ratio signalling (corrected)”, (HSDPA Ad-Hoc) TSGR1#xx(01)1087, Sophia Antipolis, France, November 5-7, 2001

[11] NTT DoCoMo, “Complexity analysis on MPIC for HSDPA”, TSGR1#19(01)0329, Las Vegas, U.S.A, February 27-March 2, 2001

[12] Nokia, “Assumptions for HS-DSCH demodulation performance tests (FDD)”, TSGR4#20(01)1452, East Brunswick, U.S.A, November 12-16, 2001

7. Appendix A

7.1. HS-PDSCH SNR Estimation – RAKE Receiver

7.1.1. Theoretical Analysis

The purpose of the SNR estimation function is to estimate the symbol SNR at the output of the HS-PDSCH despreaders, for subsequent reporting to the MAC-hs function located in the HSDPA-serving Node-B.

Consider post-despreading processing of the pilot symbol observation process (assumed here to be based on P-CPICH, although S-CPICH processing is also possible) and HS-PDSCH data despreader outputs for purpose of HS-PDSCH despreader SNR estimation. The output $p(k,l)$ of the P-CPICH despreader corresponding to the k-th pilot symbol (where $k \in \{1,...,N_p\}$ over the estimation observation interval of length N_p pilot symbols[5]) and the l-th multipath component is approximately:

$$p(k,l) = a_l \sqrt{E_{cp}} + n_p(k,l) \tag{2.1}$$

where a_l is the complex coefficient associated with the l-th multipath component, E_{cp} is the received energy per pilot chip, and $n_p(k,l)$ is a zero-mean complex Gaussian noise process. Note that we assume the channel is invariant over the observation interval.

The variance $s^2_{n_p}(l)$ of $n_p(k,l)$ (where expectation is over k) can be approximated as:[6]

$$s^2_{n_p}(l) = \frac{I_{oc} + b_l I_{or}}{C_p} \tag{2.2}$$

where C_p is the pilot symbol length in chips ($C_p = 256$) and b_l represents the fraction of I_{or} not orthogonal to the l-th pilot despreader, i.e.:

$$b_l = \frac{P - |a_l|^2}{P} \tag{2.3}$$

where P is the instantaneous channel power (i.e. $P = \sum_{l=1}^{L} |a_l|^2$).

At the output of the despreader corresponding to the k-th symbol on the m-th HS-PDSCH code (where $k \in \{1,...,N_d\}$, $m \in \{1,...,M\}$ and $N_d = 16N_p$) at multipath l, the observed statistic $d(k,l,m)$ is approximately:

$$d(k,l,m) = a_l \sqrt{E_{cd}}\, s(k,m) + n_d(k,l,m) \tag{2.4}$$

where $s(k,m)$ is the associated modulated symbol (16-QAM or QPSK).

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[5] *The estimation interval is assumed to be the first timeslot of the HS-DSCH TTI, i.e.* $N_p = 10$ *– see [10].*

[6] *For simplicity, chip-rate processing is assumed.*

Assuming the noise is independent of the multicode index, the variance $s_{n_d}^{2}(l)$ of the noise process $n_d(k,l,m)$ is:

$$s_{n_d}^{2}(l)=\frac{I_{oc}+b_l I_{or}}{C_d} \tag{2.5}$$

where C_d is the HS-PDSCH data symbol length ($C_d = 16$).

Next, assume that a channel estimator (perhaps, a linear filter operating on a sequence (over k) of $p(k,l)$ values for each l) makes available a channel estimate $\hat{h}_l$ of each multipath component as $\hat{h}_l = a_l\sqrt{E_{cp}}$. Assume, for now, that this estimate is noise-free.

Next, the combined pilot observation statistic $p_1(k)$ is formed according to:

$$\begin{aligned} p_1(k) &= \sum_{l=1}^{L} p(k,l)\,\hat{h}_l^{*} = \sum_{l=1}^{L} p(k,l)\,\hat{a}_l^{*}\sqrt{E_{cp}} \\ &= E_{cp}\sum_{l=1}^{L}\left|a_l\right|^2 + \sqrt{E_{cp}}\sum_{l=1}^{L} n_p(k,l)\,a_l^{*} \\ &= E_{cp}P + \sqrt{E_{cp}}\sum_{l=1}^{L} n_p(k,l)\,a_l^{*} \end{aligned} \tag{2.6}$$

The estimated mean of $p_1(k)$ is then formed over the length-N_p observation interval according to:

$$mean(p_1)=\frac{1}{N_p}\sum_{k=0}^{N_p-1} p_1(k) \tag{2.7}$$

where, from equation (2.6), it can be noted that:

$$E[p_1]=E_{cp}P \tag{2.8}$$

Next, the estimated energy of the combined pilot symbol observation process can be formed as:

$$energy(p_1)=\frac{1}{N_p}\sum_{k=0}^{N_p-1}\left|p_1(k)\right|^2 \tag{2.9}$$

where it can be noted that (again, from equation (2.6)):

$$E[\left|p_1\right|^2]=E_{cp}^2P^2+s_{p_1}^{2} \tag{2.10}$$

where

$$s_{p_1}^{2}=E_{cp}\sum_{l=1}^{L}\left|a_l\right|^2 s_{n_p}^{2}(l) \tag{2.11}$$

In practice, the variance of the noise process associated with the combined pilot symbol statistic is estimated according to:

$$var(p_1)=energy(p_1)-\left|mean(p_1)\right|^2 \tag{2.12}$$

and the effective SNR of the combined pilot symbol statistic can in turn be estimated as:

$$\begin{aligned} SNR_{CPICH} &= \frac{E_{cp}^2 P^2}{s_{p_1}^2} \\ &\approx \frac{\left|mean(p_1)\right|^2}{var(p_1)} \end{aligned} \tag{2.13}$$

Importantly, this estimate can be used to compute the symbol SNR at each HS-PDSCH despreader output as follows (it is assumed that the radiated power assigned to each HS-PDSCH code is identical). From above, the combined despread data symbol statistic corresponding to the k-th symbol on the m-th code is:

$$\begin{aligned} d_m(k) &= \sum_{l=1}^{L} d(k,l,m)\,\hat{h}_l^* = \sum_{l=1}^{L} d(k,l,m)\,\hat{a}_l^*\sqrt{E_{cp}} \\ &= \sqrt{E_{cp}E_d}\sum_{l=1}^{L}\left|a_l\right|^2 s(k,m) + \sqrt{E_{cp}}\sum_{l=1}^{L} n_d(k,l,m)\,a_l^* \\ &= \sqrt{E_{cp}E_d}\,Ps(k,m) + \sqrt{E_{cp}}\sum_{l=1}^{L} n_d(k,l,m)\,a_l^* \end{aligned} \tag{2.14}$$

Over some subset of size M_1 of the M available HS-PDSCH codes, next form the estimate:

$$energy(d_m) = \frac{1}{M_1 N_d}\sum_{m=1}^{M_1}\sum_{k=0}^{N_d-1}\left|d_m(k)\right|^2 \tag{2.15}$$

where it can be noted (from equation (2.14), and assuming $E\left[\left|s\right|^2\right]=1$) that:

$$E\left[\left|d_m\right|^2\right] = P^2\,E_{cd}\,E_{cp} + s_{d_m}^2 \tag{2.16}$$

where

$$s_{d_m}^2 = E_{cp}\sum_{l=1}^{L}\left|a_l\right|^2 s_{n_d}^2(l) \tag{2.17}$$

However, from equations (2.2) and (2.5), it can be seen that:

$$s_{n_d}^2(l) = \frac{C_p}{C_d}\,s_{n_p}^2(l) \tag{2.18}$$

and so, from equations (2.11) and (2.16) it can be seen that:

$$s_{d_m}^2 = \frac{C_p}{C_d}\,s_{p_1}^2 \tag{2.19}$$

Accordingly, an estimate of the power P_N of the noise process associated with the HS-PDSCH despreader outputs can be formed according to:

$$P_N \approx \frac{C_p}{C_d}\,var(p_1) \tag{2.20}$$

Further noting that:

$$E\left[\left|d_m\right|^2\right] - \sigma_{d_m}^2 = P^2 E_{cd} E_{cp} \tag{2.21}$$

then the observed HS-PDSCH QAM symbol power P_S (which is equal to $P^2 E_{cd} E_{cp}$) can be estimated as[7]:

$$P_S = energy(d_m) - \frac{C_p}{C_d} var(p_1) \tag{2.22}$$

Finally, the estimated HS-PDSCH despreader SNR can be computed as:

$$SNR_{HS-PDSCH} = \frac{P_S}{P_N} \approx \frac{energy(d_m) - \frac{C_p}{C_d}\mathrm{var}(p_1)}{\frac{C_p}{C_d}\mathrm{var}(p_1)} \tag{2.23}$$

7.1.2. CPICH and HS-PDSCH SNR Ratio

From equations (2.8), (2.10), (2.16) and (2.18) it can be readily shown that:

$$\frac{SNR_{HS-PDSCH}}{SNR_{CPICH}} = \frac{E_{cd}}{E_{cp}} \frac{\sigma_{p_1}^2}{\sigma_{d_m}^2} = \frac{E_{cd}}{E_{cp}} \frac{C_d}{C_p} \tag{2.24}$$

Accordingly, given a reported estimate of the pilot SNR SNR_{CPICH} from the UE, the Node-B can infer the HS-PDSCH despreader output SNR for a given ratio E_{cd}/E_{cp} at the Node-B transmitter.

7.1.3. Algorithm Summary

In summary, the required steps in the algorithm for P-CPICH (S-CPICH) SNR estimation are as follows (the corresponding equations above are identified), where operations incurring *significant* additional computational cost *over and above* those operations required for conventional HS-PDSCH data recovery are identified in bold:

1. Despread N_p pilot symbols (2.1).
2. Form per-ray pilot-derived channel estimates $\hat{h}_l$.
3. **Form combined pilot observation** p_1 **over** l **multipath components for** N_p **symbols** (2.6)**.**
4. **Estimate** $mean(p_1)$ **over** N_p **symbols** (2.7)**.**
5. **Estimate** $energy(p_1)$ **over** N_p **symbols** (2.9)**.**
6. Compute $var(p_1)$ (2.12).
7. Compute SNR_{CPICH} (2.13).

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[7] *RAN-WG1's working assumption is that the pilot to HS-PDSCH data ratio need not be transmitted by the Node-B to the UE.*

7.1.4. Computational Complexity

Clearly, the effective computational complexity of the procedure will depend on the particular architectural partitioning in the UE. Some outline estimates of computational complexity can still be made, however. Step 3 above (pilot symbol combining) might well be performed in a chip-rate hardware accelerator, since it is entirely analogous to R'99 chip-rate data symbol combining procedures. Steps 4 and 5 can also be performed by the chip-rate processor. Assuming steps 4 and 5 *must* be executed in some general-purpose DSP core however, the total complexity can be estimated as follows.

Step 4 requires approximately $2N_p$ additions (assuming complex operands, and neglecting the one-time division operation). Step 5 requires approximately $2N_p$ additions and $2N_p$ multiplications ($4N_p$ ops in total). Summing steps 4 and 5 gives the total number of elementary operations N_{ops} per TTI as:

$$N_{ops} = 6N_p \tag{2.25}$$

Assuming, as stated above, CPICH SNR estimation processing takes place over the first timeslot of the TTI, then $N_p = 10$. Then, $N_{ops} = 60$ per 2ms TTI, which gives an average load of approximately 30kops/s when consecutive TTI's are addressed to a single UE. For the lowest HSDPA capability class (where 1 TTI in every 3 TTI's is active), the average computational load becomes approximately 10kops/s, which clearly is very small compared to contemporary signal processing resources.

7.2. HS-PDSCH SNR Estimation – Equalizer Receiver

There are few differences in SNR estimation between the single-antenna RAKE and equalizer receiver architectures. The required processing to estimate the HS-PDSCH P_S and P_N is outlined in [10]. Since a single pilot despreader follows the equalizer, no pilot symbol combining (as embodied in equation (2.6)) is required. Rather, the pilot symbol process $p_1(k)$ is immediately available post-despreading of the equalizer CPICH process Similarly, the per-multicode despread data symbol observations $d_m(k)$ are also available without combining.

By viewing the concatenation of the instantaneous channel impulse response with the equalizer impulse response as a single composite linear channel, the arguments applied above to show that the noise power at the output of the HS-PDSCH despreaders and pilot despreader are related by the ratio of the respective pilot and data spreading factors (equation (2.24)) can be re-deployed.

Denoting the amplitude of the selected tap (assumed real) of the resulting concatenated response as P, and denoting the respective received energy per pilot chip and data chip as $\sqrt{E_{cp}}$ and $\sqrt{E_{cd}}$, the statistic at the outputs of the pilot despreader has the form:

$$p_1(k) = P\sqrt{E_{cp}} + n_p(k) \tag{2.26}$$

and the statistic at the m-th HS-PDSCH despreader output has the form:

$$d(k,m) = P\sqrt{E_{cd}}\,s(k,m) + n_d(k,m) \tag{2.27}$$

Accordingly,

$$E[p_1] = P\sqrt{E_{cp}} \tag{2.28}$$

$$E\left[\left|p_1\right|^2\right] = P^2 E_{cp} + s_{p_1}^2 \tag{2.29}$$

$$E\left[\left|d_m\right|^2\right] = P^2 E_{cd} + s_{d_m}^2 \tag{2.30}$$

Estimating the pilot process variance and data symbol energy as before, permits the HS-PDSCH SNR (which is given exactly by $SNR_{HS-PDSCH} = P^2 E_{cd} / s_{d_m}^2$) to be estimated. Noting further than the pilot SNR is $SNR_{CPICH} = P^2 E_{cp} / s_{p_1}^2$, then the ratio of CPICH SNR to HS-PDSCH can be expressed as required:

$$\frac{SNR_{HS-PDSCH}}{SNR_{CPICH}} = \frac{E_{cd}}{E_{cp}} \frac{s_{p_1}^2}{s_{d_m}^2} = \frac{E_{cd}}{E_{cp}} \frac{C_d}{C_p} \tag{2.31}$$

8. Appendix B

The proposed text for TR 25.858 is as follows.

8.2.2.2 Measurement feedback procedure

8.2.2.2.1 UE procedure

This section provides an example for UE internal procedure used in adaptive modulation and coding operation. (*Note: following procedures need to be checked by WG4*)

UE measures a DL channel quality indicator from P-CPICH (alternatively S-CPICH in case of beamforming with S-CPICH is used). ~~Taking into account the default power offset, P_{hs}, of the HS-DSCH code channel relative to P-CPICH (or S-CPICH), UE checks from its internal mapping table which TFRC from the TFRC reference list, it would be able to receive in this channel condition.~~ The DL channel quality indicator shall be defined as the signal-noise ratio (in dB) of the P-CPICH (alternatively S-CPICH) SNR_{cpich} after multipath combining.

For the purpose of reporting, the CPICH-derived SNR_{cpich} (dB) shall be uniformly quantized to P bit precision over the range A_{min} (dB) to A_{max} (dB) (A_{min} and A_{min} integers) according to:

1a) SNR_{cpich} –> $SNR_{cpich} - A_{min}$
1b) SNR_{cpich} –> $\mathrm{floor}(SNR_{cpich}+0.5)$
1c) SNR_{cpich} –> $\max(0, SNR_{cpich})$
1d) SNR_{cpich} –> $\min(2^{P}-1, SNR_{cpich})$
1e) Report SNR_{cpich}

where the floor function is equivalent to integer rounding towards minus infinity. At the Node-B, the value of SNR_{cpich} shall be recovered according to:

2a) SNR_{cpich} –> $SNR_{cpich} + A_{min}$

The values of P and A_{min} shall be 5 and –3dB respectively, thereby allowing quantization of SNR_{cpich} over the reported range -3dB to 28dB with 1dB resolution.

~~The UE uses the $BLER_{threshold}$ criteria to determine which TFRC it is able to receive in given channel conditions. If the channel quality, measured from P-CPICH (or S-CPICH in case beamforming with S-CPICH is used) indicates that TFRC x would be received with BLER< $BLER_{threshold}$ with HS-DSCH code channel power offset<y dB (steps *TBD*) in addition to the default power level, but TFRC x+1 would be received with BLER>$BLER_{threshold}$, then UE will report TFRC x and the required power offset y back to the network.~~

~~The UE should report a TFRC from the reference list together with a power offset at the transmission timing defined by higher layer signalling.~~

8.2.2.2.2 Parameters provided to the UE measurement feedback operation

Following connection specific parameters are informed to the UE by higher layer signalling:

~~1) P_{hs}, default power offset between HS-DSCH code channel and P-CPICH (or S-CPICH in case beamforming with S-CPICH is used).~~

~~2) $BLER_{threshold}$, BLER value that UE uses for selecting the TFRC. Possible values of the $BLER_{threshold}$ are FFS.~~

1~~3~~) Measurement feedback cycle *k*. *k* has a possible value of [1,5,10,20,40,80] corresponding to the feedback cycle of [2,10,20,40,80,160] msec. In addition, with the indication *k*=0, measurement feedback can be shut off completely. (*Note: It is to be determined whether k can have two values, in which case that k is selected depending on HS-DSCH activity*.)

2~~4~~) Measurement feedback offset *l*. The exact definition of *l* is to be determined.

The use for the measurement feedback cycle *k* and feedback offset *l* is illustrated in Figure 1. (*Note: The timing relations in the figure do not reflect the differences in DPCH frame offset.)*

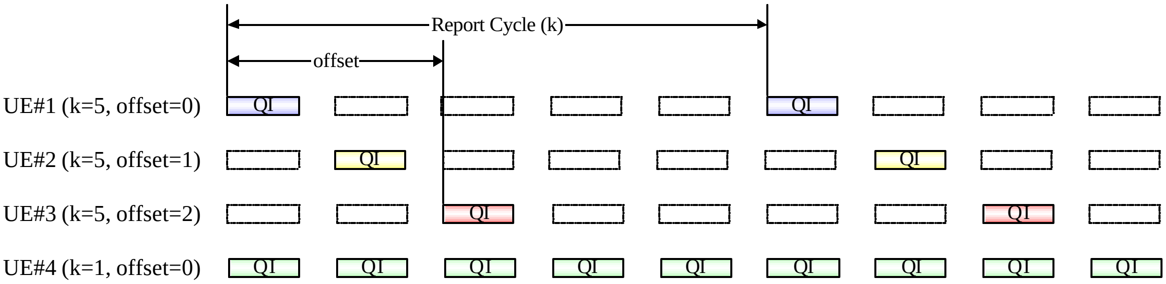

Figure 1 An illustration of feedback measurement transmission timing

~~8.2.2.2.3Reference TFRC list~~

~~Only one reference TFRC list is specified, that all UEs should use for creating the measurement report. Each TFRC contains modulation, transport block set size, and number of code channels~~

~~Example of TFRC reference list with 6 TFRCs is shown in **Table 3**~~

~~Table 3. An example of a TFRC reference list~~

~~TFRCs~~	~~Modulation~~	~~Transport block set size~~	~~# of code channels~~
~~TFRC1~~	~~QPSK~~	~~1200~~	~~5~~
~~TFRC2~~	~~QPSK~~	~~2400~~	~~5~~
~~TFRC3~~	~~QPSK~~	~~3600~~	~~5~~
~~TFRC4~~	~~16QAM~~	~~4800~~	~~5~~
~~TFRC5~~	~~16QAM~~	~~6000~~	~~5~~
~~TFRC6~~	~~16QAM~~	~~7200~~	~~5~~

~~8.2.2.2.4Measurement feedback definition~~

~~Measurement report can take up to [32] values. It contains one TFRC at a time from the reference list and a power offset value.~~

~~An example of possible UE measurement report list for 6 TFRCs and intermediate steps of 1 dB for power offset are shown in Table 4.~~

~~Table 4 An example of a measurement report list~~

~~TFRC~~	**~~Power offset~~**	~~Measurement report value~~

~~TFRC1~~	~~12 dB~~	~~0~~
	~~11 dB~~	~~1~~
	~~10 dB~~	~~2~~
	~~9 dB~~	~~3~~
	~~8 dB~~	~~4~~
	~~7 dB~~	~~5~~
	~~6 dB~~	~~6~~
	~~5 dB~~	~~7~~
	~~4 dB~~	~~8~~
	~~3 dB~~	~~9~~
	~~2 dB~~	~~10~~
	~~1 dB~~	~~11~~
	~~0 dB~~	~~12~~
~~TFRC2~~	~~2 dB~~	~~13~~
	~~1 dB~~	~~14~~
	~~0 dB~~	~~15~~
~~TFRC3~~	~~2 dB~~	~~16~~
	~~1 dB~~	~~17~~
	~~0 dB~~	~~18~~
~~TFRC4~~	~~2 dB~~	~~19~~
	~~1 dB~~	~~20~~
	~~0 dB~~	~~21~~
~~TFRC5~~	~~2 dB~~	~~22~~
	~~1 dB~~	~~23~~
	~~0 dB~~	~~24~~
~~TFRC6~~	~~2 dB~~	~~25~~
	~~1 dB~~	~~26~~
	~~0 dB~~	~~27~~
~~NA~~	~~NA~~	~~28~~
~~NA~~	~~NA~~	~~29~~
~~NA~~	~~NA~~	~~30~~
~~NA~~	~~NA~~	~~31~~