Products: Spectrum Analyzer FSP

Measurement of
Adjacent Channel Leakage Power on
3GPP W-CDMA Signals with the FSP

Application Note

This application note explains the concept of Adjacent Channel Leakage Ratio (ACLR) measurement on 3GPP W-CDMA signals with the spectrum analyzer FSP. Optimum operation for both the integrated bandwidth method and the new time domain method are shown via explanation of the signal behavior and the internal structure of the FSP. With the time domain method the spectrum analyzer family FSP provides a new approach featuring a very short measurement time compared to the integrated bandwidth method.

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1 Overview

One of the most important measurements on RF signals for digital communication systems is the leakage power in the adjacent channels. Leakage power influences the system capacity as it interferes with the transmission in adjacent channels. Therefore it must be rigorously controlled to guarantee communication for all subscribers in a network. On the other hand, time between battery charges for mobile phones is required to be as long as possible. This leads to low power consumption requirements for the mobile's transmitter. The high linearity especially of the transmitter output amplifier necessary for low adjacent channel leakage power and the low power consumption required for long battery lifetime are contradictory. This leads to very extensive testing, in production up to 100 % of the components must be tested. Therefore, for high throughput the ACLR tests must be very fast and accurate.
The concept of the FSP spectrum analyzer allows a new approach for measuring ACLR in time domain, giving an inherently low measurement time compared to the integrated bandwidth method.

This application note explains both the integrated bandwidth method and the time domain method. Dynamic range and speed considerations for ACLR measurement on 3GPP W-CDMA are presented for both methods.

2 Requirements

The 3GPP standard TS25.101 (section 6.6.2.2) sets the following requirements on ACLR measurement for mobile phones [1]:
Adjacent Channel Leakage power Ratio (ACLR) is the ratio of the transmitted power to the power measured after a receiver filter in the adjacent channel(s). Both the transmitted power and the received power are measured with a filter that has a Root-Raised Cosine (RRC) filter response with roll-off α =0.22 and a bandwidth equal to the chip rate.

Table 1 Requirement for ACLR according to 3GPP TS25.101

UE channel	ACLR limit
± 5 MHz	-32 dB or –50 dBm, which ever is higher
± 10 MHz	-42 dB or –50 dBm, which ever is higher

These values apply to the complete mobile phone. The requirements for the different modules and components of a mobile phone will be yet more demanding in order to meet the overall performance requirement.

3 The Integrated Bandwidth Method

For measuring channel power most modern spectrum analyzers provide software routines for calculating power within given channels. These routines calculate the power by integrating the power represented by the displayed trace pixels within the frequency range of the channel.

With N trace pixels within the channel the integration is performed according to the following formula:

$$CP = 10 \cdot \log\left[\frac{CHBW}{RBW \cdot k_n} \cdot \frac{1}{N} \cdot \sum_1^N 10^{\frac{P_i - a_{i/RRC)}}{10}}\right] \qquad (1)$$

Whereas

CP =   Channel Power in dBm

CHBW =   Channel bandwidth in kHz

RBW =   Resolution bandwidth used for measurement in kHz

k_n =   Correction factor for noise bandwidth of the used RBW

N =   Number of pixels within the channel

P_i =   Level represented by pixel i of the trace in dBm

$a_{i(RRC)}$ =   attenuation of the 3GPP RRC filter at pixel i in dB.

The 3GPP standard requires a Root-Raised Cosine (RRC) filter for measuring the power within each channel. Therefore each pixel used for integration is weighted with the attenuation of the RRC filter at the respective offset from the channel center frequency.

In order not to affect the RRC filter shape the spectrum analyzer resolution bandwidth needs to be narrow compared to the channel filter bandwidth. Therefore, the FSP uses the 30-kHz resolution filter.

For power measurements the FSP provides the sample detector or the RMS detector. The RMS detector is preferable as it measures power more repeatably compared to the sample detector. The RMS detector measures the power of the spectrum represented by a pixel by applying the power formula to the IF envelope voltage. For higher repeatability the measuring time per pixel can be controlled by the sweep time. With higher sweep time the time for power integration for each pixel increases; the trace becomes smoother with longer sweep times and with it the test results become more stable and repeatable. With the sample detector only one sample per pixel coming from the AD Converter is displayed independent of the sweep time. The trace is rather noisy and as a consequence the test results are not as repeatable as with the RMS detector.

For these reasons the FSP selects the RMS detector automatically when the adjacent ahannel power measurement is switched on.
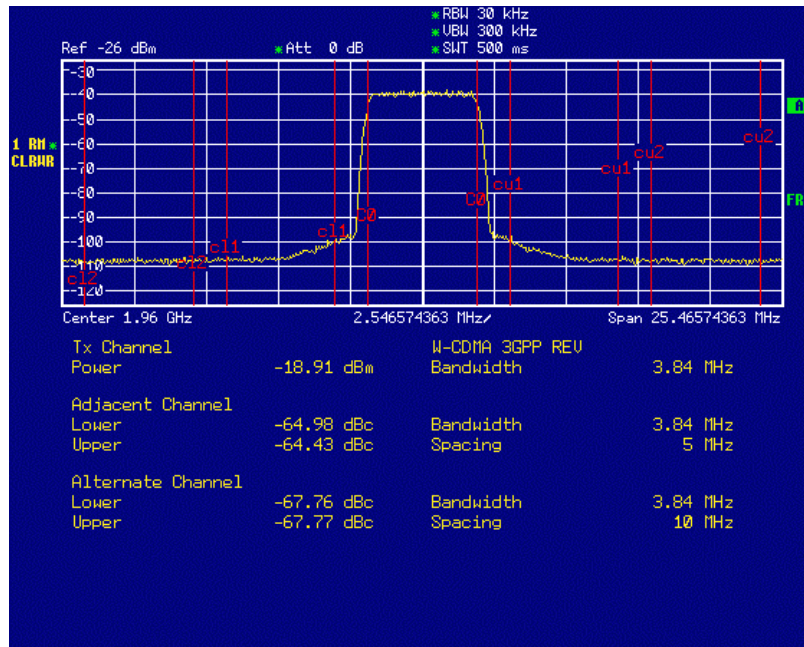
Fig 1   3GPP ACLR measurement using the IBW method

The FSP screen shot in figure 1 shows an ACLR measurement on a 3GPP up-link signal loaded with 3 code channels. The upper part shows the spectrum of the signal. In the lower half the numbered values of the channel power and the ACLR of the 5- and 10-MHz offset channels is displayed. Due to the use of the RMS detector the smoothness of the trace and the stability of the test results are dependent on the sweep time set.

Calculating the standard deviation of 100 consecutive ACLR measurements on a 3GPP up-link signal using different sweep times generates the graph in figure 2.
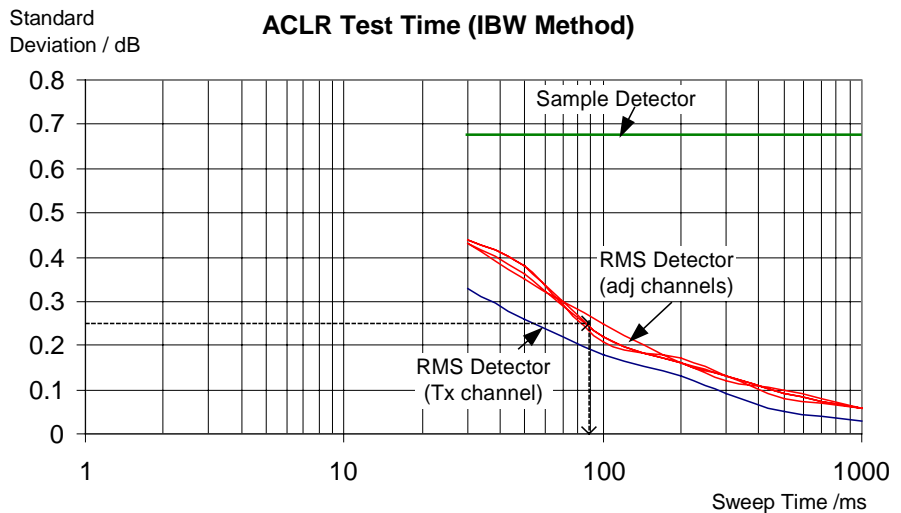

Fig 2 Standard deviation of the ACLR measurement on a 3GPP up-link signal dependent on the sweep time.

The x-axis shows the sweep time. On the y-axis the corresponding standard deviation of the test results is shown. For comparison it also shows the standard deviation with the sample detector used for the

measurement. It is not dependent on the sweep time. Reductions in deviation of sample-detected measurements requires that multiple traces be taken. The multiple traces are then averaged, resulting in smaller deviations. Measurement time is dependent upon available processor speed. The sample-detected method generally is much slower than the RMS method for similar measurement uncertainty. In addition there is a risk of an additional error due to the averaging process. As the trace averaging is performed in logarithmic scaling, the log average value must be corrected for absolute power. The correction factor is well known for white noise (+2.51 dB). Dependent on the amplitude statistics in the TX channel and in the adjacent channels it may vary between the channels. This leads to different results even with the relative power measurement in case of the ACLR measurement.

For 95 % confidence level for the measured power being within 0.5 dB of the true result the standard deviation is 0.25 dB (assuming a Gaussian distribution of the test results). The sweep time needed for 0.25 dB standard deviation using RMS detection is about 90 ms. The overhead for internal calculation of the power in the 5 channels and the transfer of the test results via GPIB to an external controller is 22 ms. This results in a total test time of 122 ms.

The narrow resolution bandwidth necessary with the Integration Bandwidth (IBW) method has the primary impact on sweep time. The sweep time of a spectrum analyzer is dependent on the span and the RBW according to (2).

$$SWT = \frac{span}{RBW^2} \text{ , whereas SWT = sweep time} \qquad (2)$$

This results in an enormous increase of the sweep time, if the 30-kHz RBW is used with the IBW method as opposed to the channel bandwidth maximum usable for the measurement. However, due to the limited selectivity provided by general spectrum analyzer filters, a resolution bandwidth equal to the channel bandwidth is not possible, even if it were available.

4   The Time Domain Method

Due to the digital filtering concept of the FSP series spectrum analyzers it is now feasible to implement the filters necessary for direct channel power measurement.

Fig 3 explains with a simplified block diagram the implementation of the digital resolution filters up to 100 kHz bandwidth and the channel filters for ACP measurement.
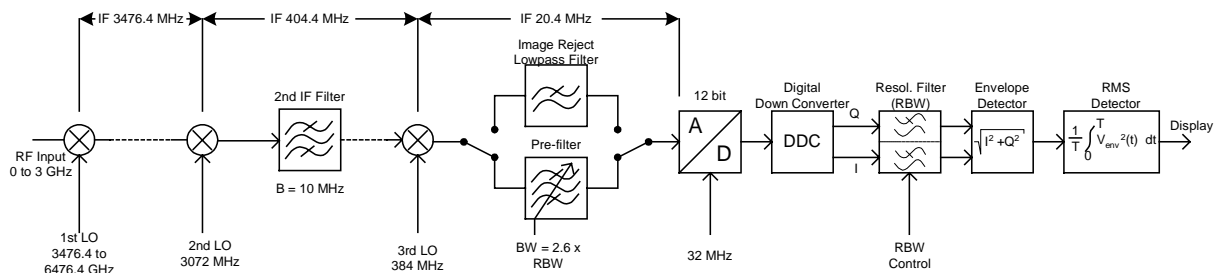

Fig 3   Simplified block diagram of the FSP signal path

The RF input signal is converted to the 20.4 MHz final IF using triple conversion. The 10 MHz IF filter is implemented in the 2nd IF path (404.4 MHz). In addition to the 10-MHz resolution filtering it rejects the image frequency of the 3rd IF (f_{in} -42.8 MHz).

For narrow resolution bandwidths a tunable pre-filter is available in the 20.4-MHz IF. The bandwidth of the pre-filter is about 2.6 times the selected resolution bandwidth. It prevents the AD Converter from being loaded with out of band signals. In the case of modulated signals like 3GPP W-CDMA it limits the spectrum applied to the AD Converter.

The 20.4-MHz IF is digitized using a 12-bit AD Converter (AD9042 from Analog Devices) with a sampling rate of 32 MHz (band-pass sampling).

The maximum attainable bandwidth w/o image response is about 1.5 MHz due to the tunable analog pre-filter. For wider bandwidth the 20.4-MHz IF is shifted to 21.4 MHz and a fixed-bandwidth image rejection filter is used. Bandwidths up to 6.4 MHz can be achieved that way.

The digitized IF is down-converted into IQ domain using a digital down converter (DDC). The inphase (I) and quadrature (Q) signals are lowpass filtered to achieve the desired resolution bandwidth. These digital lowpass filters are programmable and can be configured for nearly arbitrary filter characteristics. The FSP series uses digitally implemented Gaussian type resolution BW filters from 10 Hz to 100 kHz for fastest sweep speed.

For channel power measurement specific channel filters are available. For channel power measurement on 3GPP W-CDMA signals, for example, a Root-Raised Cosine filter is available with the 3 GPP characteristic figures (f_{chip} = 3.84 MHz, α = 0.22). Figure 4 shows the amplitude response of the filter.

```
Ref  0 dBm
  0
-10
-20
-30
-40
-50
-60
-70
-80
-5 MHz              Center              +5 MHz
```
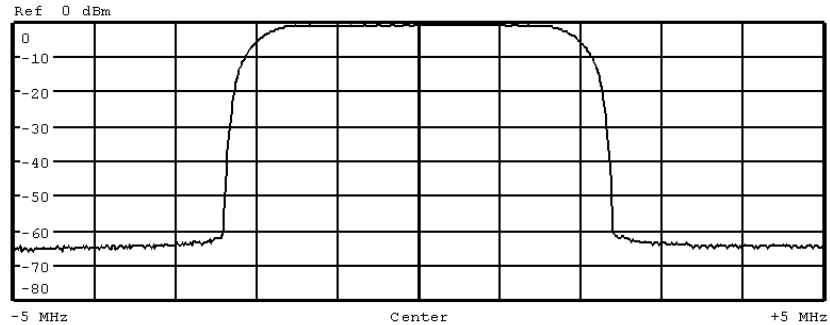
Fig 4   Amplitude response of the 3GPP W-CDMA RRC channel filter

At the output of the resolution filters the IF envelope voltage and its RMS value are calculated. The RMS integration time (T) is dependent on the sweep time and can be set in a wide range. With the FSP it is coupled to the selected sweep time in the frequency domain as well as in the time domain.

For the time domain method the FSP steps through the different channels specified by the ACP setup and measures the power in each channel according to the sweep time set up. Measurement time for each channel is defined as the sweep time divided by the number of channels. For each channel the FSP generates a single number for the power within the channel.
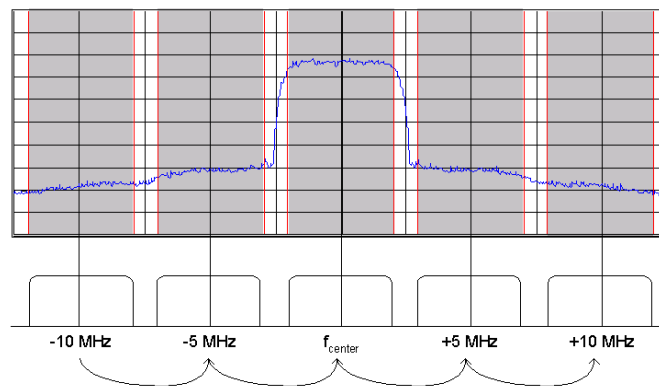

Fig 5   Principle of the Time Domain Method

Fig 5 illustrates the procedure of the time domain method. Stepping through the different channels is equivalent to setting a channel filter to the respective channel.

The total time needed for a complete ACLR measurement is determined by the sweep time plus the time the synthesizer needs for switching its frequency through the different channels. The FSP uses a VCO local oscillator design providing a fast settling time compared to the YIG oscillator design used in most spectrum analyzers. The time needed to set up a new frequency is about 6 ms.

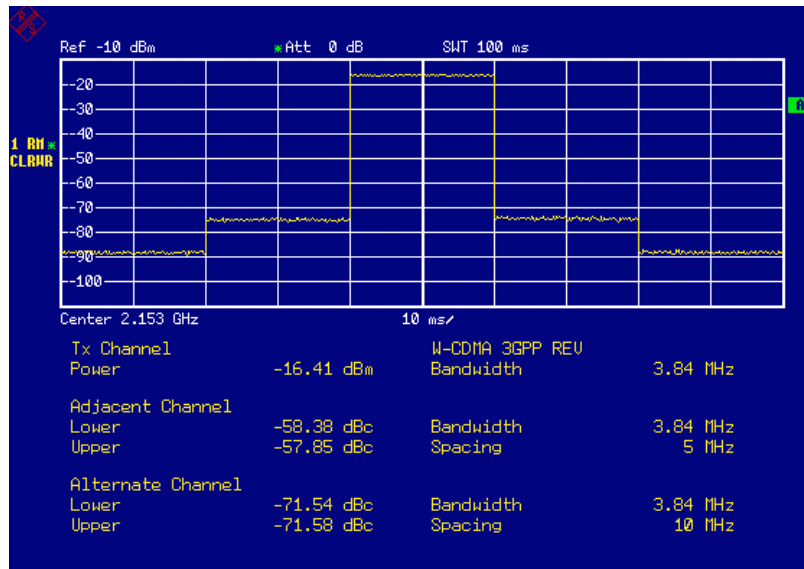Figure 6 shows a screen shot with an ACLR measurement on a 3GPP W-CDMA up-link signal.

```
Ref -10 dBm          *Att  0 dB      SWT 100 ms

 -20
 -30                                                            A
 -40
 -50
1 RM
CLRWR -60
 -70
 -80
 -90
-100

Center 2.153 GHz                     10 ms/

Tx Channel                    W-CDMA 3GPP REV
Power              -16.41 dBm  Bandwidth           3.84 MHz

Adjacent Channel
Lower              -58.38 dBc  Bandwidth           3.84 MHz
Upper              -57.85 dBc  Spacing                5 MHz

Alternate Channel
Lower              -71.54 dBc  Bandwidth           3.84 MHz
Upper              -71.58 dBc  Spacing               10 MHz
```

Fig 6   3GPP ACLR measurement using the Time Domain Method

The trace in the diagram in the upper half of the screen shows the power of the different channels in time domain. In the lower half of the screen the measured power levels are listed together with the channel configuration.

Due to the wider bandwidth used for the channel power measurement, the time required for a measurement is short compared to the IBW method. Similar to figure 2, the graph in figure 7 shows the standard deviation of 100 consecutive measurements.

Standard
Deviation /dB

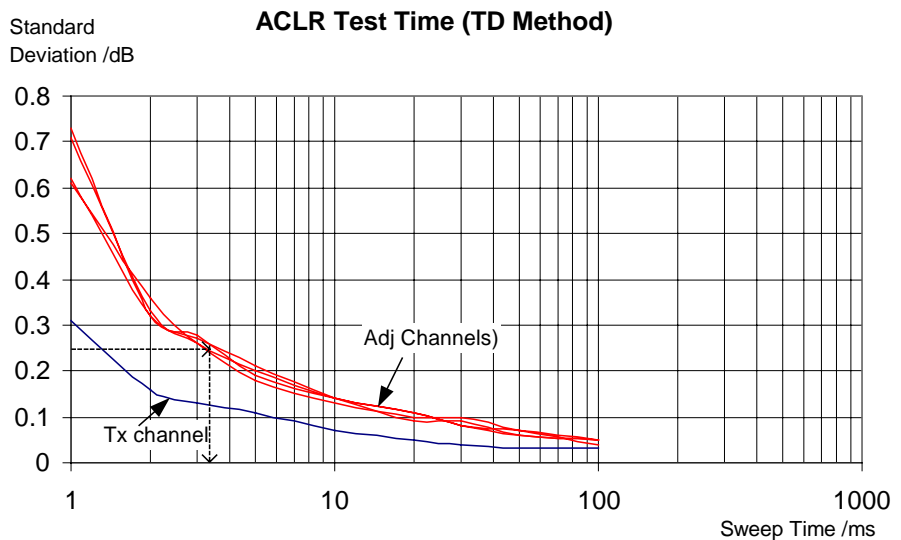**ACLR Test Time (TD Method)**

Fig 7 Standard deviation of the ACLR measurement on a 3GPP up-link signal dependent on the sweep time using the TD method.

According to figure 7 the net test time for a standard deviation of 0.25 dB, (a repeatability of 95 % confidence level) is about 3.2 ms. Taking into account the time the sythesizer needs for switching between the different channels and the transfer of the test results to an external controller via GPIB the total test time for 5 channels is 23 ms. Compared to the total test time needed with the IBW method (112 ms) this is an improvement by a factor of 4.8.

5   Dynamic Range

Additionally to the test time, the inherent dynamic available for ACLR measurement is an important factor. The inherent power of the spectrum analyzer must be well below the limits for the device under test. This assures that the test result is minimally affected by the ACLR floor of the spectrum analyzer.

The inherent dynamic range of the FSP when measuring power in the adjacent channels is determined by three factors:

- The internal noise floor.
- The phase noise of the FSP.
- The spectral re-growth due to 3rd order intermodulation of the analog signal path in the 5-MHz offset channel.

Note: *With the TD method in addition the intermodulation of the AD Converter must also be considered.*

From these factors the optimum level settings can be determined taking the internal structure of the FSP into account.

Two parameters must be considered when setting up a spectrum analyzer to achieve dynamic range:

- The RF attenuation in order not to generate spectral re-growth in the internal mixer stages.
- The reference level in order not to overdrive the IF path following the internal bandpass filters and the AD Converter.

With the FSP in basic configuration the RF attenuator can be set in 10-dB steps. When the option B25 is installed, an additional mechanical 5-dB step is available. In addition the option B25 provides an electronic attenuator from 0 to 30 dB in 5-dB steps. The use of the electronic attenuator is especially advisable for production testing due to the frequent switching of the RF attenuator. Compared to the mechanical switches in the standard attenuator the electronic attenuator provides nearly unlimited lifetime, whereas mechanical switches are limited to about 5 million switchings.

The reference level can be set independently from the input attenuator for a chosen range. The level of -10 dBm maximum allowed at the input mixer determines the maximum allowable reference level. For 0-dB RF attenuation the maximum reference level is -10 dBm, for example. When RF attenuation is switched on, the allowed reference level is increased by the value of the RF attenuation up to a maximum of 30 dBm.

The reference level sets the level at the AD Converter input. It can handle signals up to 3 dB above reference level without clipping. For greater signal levels it begins to clip the signal and generates distortion.

For 3 GPP W-CDMA signals the reference level (and dynamic range) is optimum when the AD Converter is driven to full scale. Due to the crest factor of the W-CDMA signal the mean power of the signal will be well below full scale.

The FSP screen shot in figure 8 shows the Complementary Cumulative Distribution Function (CCDF) of the 3 GPP up-link signal used for the measurements in this application note. For comparison the CCDF is measured with 10 MHz resolution bandwidth and 100 kHz resolution

bandwidth. When measuring the W-CDMA signal with a narrow resolution bandwidth the CCDF approaches the CCDF of white noise, also shown in figure 7.
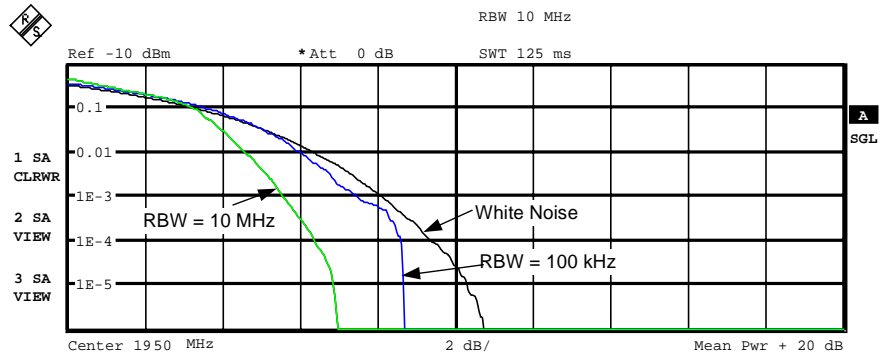

Fig 8   Complementary Cumulative Distribution Function (CCDF) of the 3GPP up-link signal loaded with three code channels.

The peak power of the wide band signal measured with 10 MHz resolution bandwidth is about 7 dB greater than the mean power (crest factor = 7 dB).

Due to the bandwidth limitation in the analog signal path of the FSP the signal level applied to the AD Converter is low with the IBW method compared to the TD method.

With the TD method only the 10 MHz resolution filter at the 2nd IF (see fig 3) is available for analog pre-filtering. It does not attenuate the W-CDMA TX signal when measuring in the 5-MHz offset channel. The AD Converter has to cope with the complete signal. With a crest factor of 7 dB the reference level must be set 7 dB higher relative to the mean power.

With the IBW method the FSP is set to 30 kHz resolution bandwidth. With 30 kHz resolution bandwidth an analog band pass filter is used in front of the AD Converter providing a bandwidth of 78 kHz. The power of the W-CDMA signal is reduced by

$$10 \cdot \lg\left(\frac{3840\,\text{kHz}}{78\text{kHz}}\right) = 16.9\text{dB} \tag{3}$$

Therefore the mean power in front of the AD Converter is 16.9 dB lower compared to the TD method.

According to the central limit theorem the amplitude distribution of a digital modulated signal approaches the distribution of white noise in case of narrow band filtering (see figure 8); the crest factor of the filtered signal changes. For the following considerations the crest factor will be assumed to be 10 dB. Taking into account the different crest factors (10 dB vs. 7 dB) the reference level with the IBW method can be set 14 dB lower compared to the TD method. These lower reference levels used in the IBW method result in an increased gain used at the input of the AD converter. The increased gain in the IF results in a lower noise figure.

The following graph shows the FSP noise figure at a -30 dBm reference level versus the set reference level.
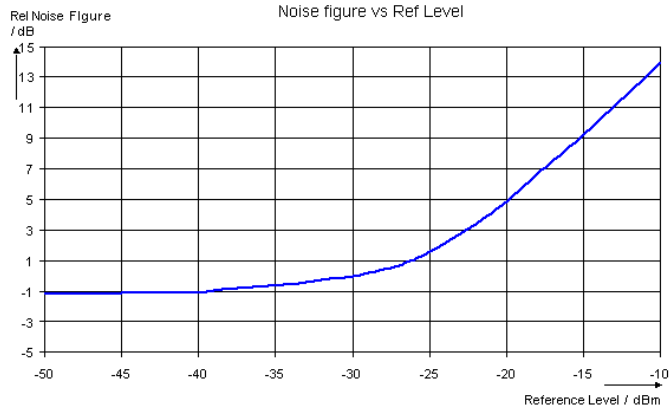
Fig 9   Relative noise figure of the FSP dependent on the reference level set to a -30 dBm reference level and 0 dB RF attenuation, valid for digital implemented RBW

Taking into account the different reference levels possible for the IBW method and the TD method the following graphs show the dynamic range achievable dependent on the mixer level of the FSP. (Mixer level = mean power at the input mixer = mean power at the RF input - RF attenuation)

With the IBW method shown in figure 8 the optimum mixer level is determined mainly by the inherent noise floor and the spectral re-growth intercept. At about -20 to -22 dBm level at the input mixer the FSP provides a dynamic range of about 65 dBc for ACLR in the 5-MHz offset channel.
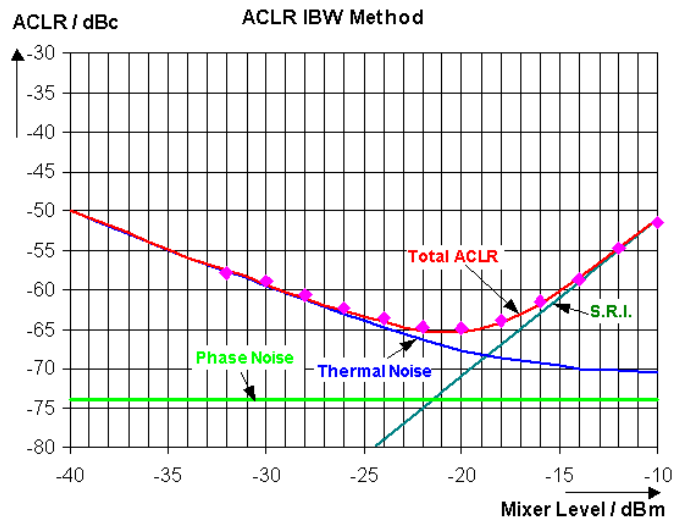

Fig. 10 IBW method: FSP inherent dynamic range for 3 GPP ACLR in the 5-MHz offset channels dependent on the level at the input mixer. The dots show measured values using a FSP7

With the TD method the AD Converter must cope with the TX signal when measuring in the 5-MHz offset channel. Therefore the intermodulation products from the AD Converter contribute to the leakage power in the offset channels. Contrary to intermodulation(IM) products generated in the analog signal path, the level of the IM products generated by the AD Converter are nearly independent of the signal level. The ACLR due to intermodulation of the AD Converter decreases linearly with the decrease of

the mixer level. In addition, the inherent thermal noise power is greater compared to the IBW method, due to the higher reference level that must be used to avoid overloading the AD Converter.

Both the inherent thermal noise power and the intermodulation of the AD Converter contribute to the total power in the 5-MHz offset channel.

The attainable dynamic range using the TD method is shown in figure 9. Neither the phase noise nor the spectral re-growth of the analog signal path greatly influence the dynamic range. The optimum mixer level is at the highest possible value without overloading the AD Converter.
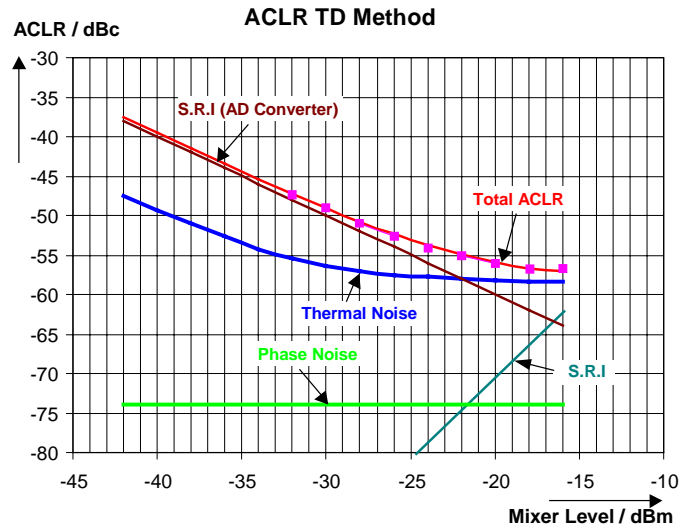
ACLR TD Method

Fig. 11 TD method: FSP inherent dynamic range for 3 GPP ACLR in the 5-MHz offset channels dependent on the level at the input mixer. The dots show measured values using a FSP7.

Note: *The AD Converter can be loaded by 3 dB above reference level without clipping the signal. Fig. 9 assumes signal peaks up to reference level only. The graph can be applied for signals with 6 to 9 dB crest factor.*

When measuring the power in the 10-MHz offset channel the TX signal is attenuated by the 2nd IF 10-MHz filter in the FSP analog signal path (see fig 3). Compared to the 5-MHz offset power measurement the signal level at the AD Converter is much lower, therefore. This is used by the FSP to gain dynamic range in the 10-MHz offset channels by decreasing the reference level by 20 dBso that the AD Converter wideband noise contribution to the overall noise floor of the FSP is reduced. The spectral re-growth in the 10-MHz offset channel is due to 5th order intermodulation products rather than 3rd order intermodulation in the 5-MHz offset channel. 5th order modulation products in general show lower level compared to 3rd order IM products.

The inherent dynamic range for ACLR measurements in the 10-MHz offset channel is shown in the following graph.
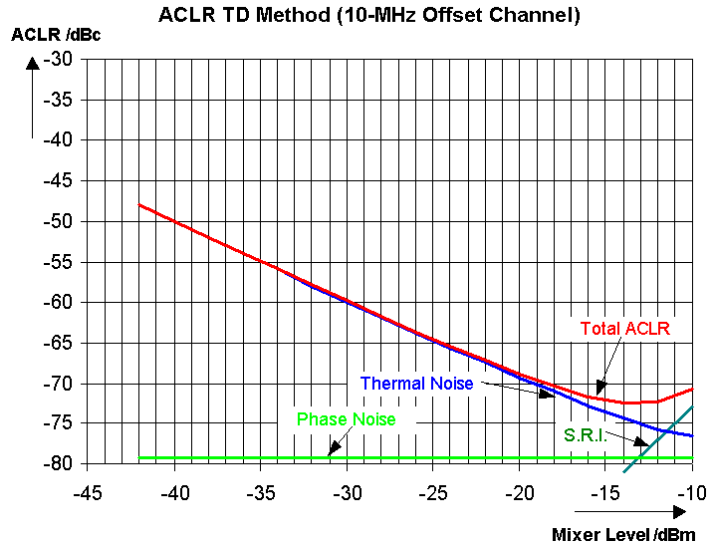
Fig 12   TD method: FSP inherent dynamic range for 3 GPP ACLR in the 5-MHz offset channels dependent on the level at the input mixer

When selecting one of the 3GPP standards offered for ACLR measurements, the FSP adjusts the reference level automatically when measuring the power in the 10-MHz offset channel.[1]

To free the user from calculating the optimum mixer level the FSP provides automatic routines to establish the optimum RF attenuation and reference level. With the signal from the device under test applied to the FSP the user simply activates the soft key ADJUST REF LEVEL in the ACPR menu of the FSP to optimize the dynamic range of the instrument.

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[1] Automatic reference level setting is implemented in the FSP firmware from version 1.20. For previous versions the reference level is constant for all offsets.

6 Measurement Accuracy

When testing ACLR measurement accuracy is a very important issue. Low errors due to the test equipment leave a higher margin at the device under test or can be used to reduce test time with the trade-off of a reduced repeatability (see figure 2 and figure 7).

As the ACLR is determined by the difference of two power levels only those spectrum analyzer parameters which contribute to relative display accuracy must be considered. These are the display non-linearity and the frequency response within the range of the channels considered. Due to the digital implementation of the resolution bandwidths and the detectors in the FSP, the display non-linearity is due only to the AD Converter. The FSP specification from reference level to -70 dB from reference level is <0.2 dB.

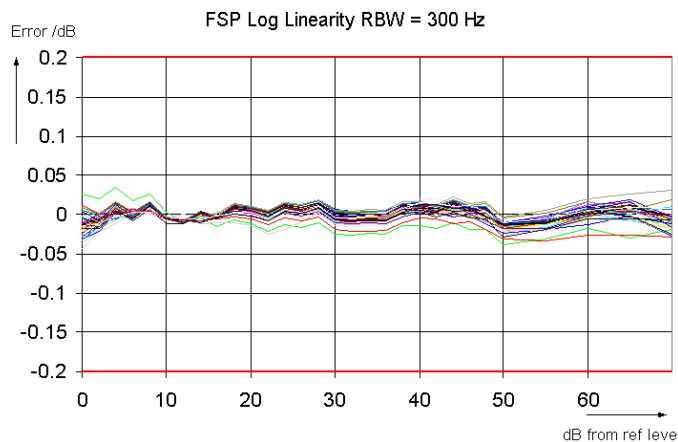The following graph shows the display non-linearity measured on 30 FSPs.

Fig 13   Measured display non-linearity measured with 30 FSPs

To avoid the influence of the FSP noise floor the 300 Hz resolution bandwidth is selected for this measurement. It shows that with all 30 FSPs the display nonlinearity is within $\pm$ 0.05 dB

The frequency response of the FSP is specified with <0.5 dB from 10 MHz to 3 GHz. However, for ACLR measurement only frequency ranges of $\pm$12.5 MHz around the FSP center frequency have to be regarded. In this narrow frequency band a frequency response of less than 0.1 dB ought to be expected.

From these two error contributions the total measurement uncertainty with 95 % confidence level is 0.116 dB.

In addition to the inherent level errors of the FSP the signal to noise ratio is an important factor in the expected error. The inherent adjacent channel power of the FSP adds to the power of the device under test. Both linear power levels are added and result in the displayed power level.

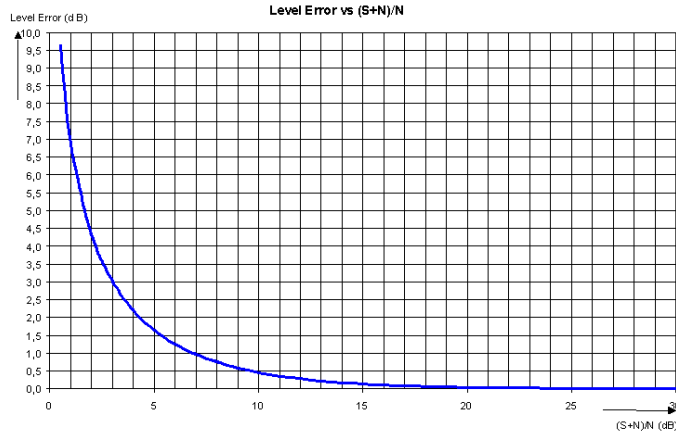The following graph shows the error due to the signal to noise ratio.

Fig 14 Level error dependent on (S+N)/N

With a (S+N)/N in the offset channels of 10 dB the error will be 0.5 dB. In the TX channel the error is negligible due to the high S/N. Therefore a minimum margin of 10 dB is desirable for low ACLR error. Regarding the limits set by the 3GPP TS25.101 specification, the FSP can easily meet this requirement for both, the IBW and the TD method.

Table 2   3GPPspecifications vs. optimum dynamic range of the FSP for typical signals, IBW vs. TD method.

UE channel	ACLR limit	FSP Dyn Rge IBW Method	FSP Dyn Rge TD Method
± 5 MHz	-32 dBc	- 65 dBc	- 57 dBc
± 10 MHz	-42 dBc	- 67 dBc	-71 dBc

7   Literature

[1] Technical Specification TS 25.101 V3.1.0, 3rd Generation Partnership Project (3GPP), Technical Specification Group (TSG) RAN WG4, UE Radio transmission and Reception (FDD)

[2] Application Note 1EF40_E, Measurement of Adjacent Channel Power on Wideband CDMA Signals

[3] Josef Wolf and Bob Buxton, "Measure Adjacent Channel Power With a Spectrum Analyzer, " *Microwaves & RF,* January 1997, pp. 55-60.

[4] Bob Buxton, Steve Stanton and Josef Wolf, "ACP Measurements on Amplifiers Designed for Digital Cellular and PCS Systems," Proceedings of the 6th Annual Wireless Symposium, Feb 1998, pp. 122 to 127

8   Ordering information

Type of instrument

		Ordering number
FSP3	9 kHz to 3 GHz	1093.4495.03
FSP7	9 kHz to 7 GHz	1093.4495.07
FSP13	9 kHz to 13,6 GHz	1093.4495.13
FSP30	9 kHz to 30 GHz	1093.4495.30

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