JITTER MEASUREMENT TO 10 PPM LEVEL FOR PULSED RF POWER AMPLIFIERS 3 – 12 GHz

C. Gough†, S. Dordevic, M. Paraliev, Paul Scherrer Institut, 5232 Villigen PSI, Switzerland

Abstract

Linacs for FEL application require a low jitter RF path from RF source through pulsed amplifiers, klystron / modulators and cavities. For the SwissFEL project at the Paul Scherrer Institute, pulsed solid state power amplifiers of the 500 W / 3 µs class for driving the klystrons were required. For these amplifiers, a stable interferometer system was developed to measure the residual RF jitter levels to <10 ppm (parts per million) rms and <10 µrad (0.573 millidegree RF) rms. This paper describes the measurement system and gives some measurement results for 3 GHz, 5.7 GHz and 12 GHz amplifiers.

INTRODUCTION

Precision amplitude and phase noise measurement is well advanced for CW RF systems, typically giving a result in the frequency domain [1-2]. For the highest resolution, interferometer systems are used to give a phase noise floor of ~-180 dBrad^2/Hz in the frequency domain [3]. CW measurement systems have such high sensitivity that they can also be used for pulsed RF systems by choosing a reasonably high repetition rate (that is a high duty cycle), and accepting that the result has a sensitivity reduced by the factor of the duty cycle. The pulsed measurement is more troublesome then CW because of high dynamic range transients.

Three general methods were considered to measure the phase and amplitude jitter of an RF pulse amplifier:

- Method 1: Mix the output signal down to an intermediate frequency and use a high resolution ADCs (Analog to Digital Convertors) (typically 16-bit 250 Ms/s) to digitize this sinewave. The digital samples are quadrature demodulated to resolve amplitude and phase.
- Method 2: Mix the output signal down to an intermediate frequency and use a precision DC offset with diode limiters and amplifiers. This results in clipping most of the waveform away but leaving the peak of the sinewave to be amplified and measured. Amplitude information only is digitised.
- Method 3: Null the amplifier output signal precisely with an inverted copy of the amplifier input signal. This is an interferometer configuration and results in a difference signal which is amplified and mixed down to DC for digitizing.

Method 1 was suitable for general purpose full range measurements to ~30 ppm level only [4]. To reduce the noise contribution, pairs of ADCs were used to digitize the same analog signal and then the outputs were summed digitally.

Method 2 is very suitable when the signal is in the 10-100 V range at frequencies below ~50 MHz but has not been tested with down-conversion from higher frequencies and with lower signal levels. However, the method can achieve <1 ppm rms amplitude measurement cycle-by-cycle at 20 MHz [4].

Method 3 was chosen: (a) to have sensitivity <10 ppm; (b) to have equal sensitivity to amplitude and phase noise for calibration reasons; (c) to avoid large dynamic and thermal transients in the measurement system due to high signal levels. This last consideration comes because the absolute amplitude and phase slope during the RF pulse should also be measured accurately. The system developed in this paper is very similar to that described in [5], but benefiting from the general improvement in electronics over the intervening 50 years.

DEFINITION OF JITTER

In this application, jitter is dominated by random events in the time domain, and the analysis is best done in time domain to identify the causes. During the RF pulse, the interferometer output signal is mixed down to DC then sampled by an oscilloscope to give a single-shot waveform A^n:

$$\boldsymbol{A}_k^n = [a_1^n \dots a_k^n] \tag{1}$$

Typically, N = 100 waveforms, each with K > 300 samples at ~100 Ms/s, are stored. The mean waveform of **A** is:

$$\overline{\boldsymbol{A}_k} = \frac{1}{N}\sum_1^N \boldsymbol{A}_k^n \tag{2}$$

The jitter is defined as the average of the rms deviations from the mean waveform:

$$\sigma = \frac{1}{K}\sum_1^K \sqrt{\frac{1}{N}\sum_1^N (\boldsymbol{A}_k^n - \overline{\boldsymbol{A}_k})^2} \tag{3}$$

AMPLIFIER MEASUREMENT SYSTEM DESCRIPTION

In Fig. 1, the choice of particular components was the result of tedious trial and error. Many components suffer from low level jitter problems. However, the mechanical attenuators of Hewlett Packard design were exceptionally stable. Mechanical delay lines from Narda and Colby also proved usable, even with relatively large reflection coefficients. The mechanical phase shifters have the advantage that they do not add noise; so far, attenuators and phase shifters based on analog biasing of diodes have not been succesful. Minicircuits ZVA183 amplifiers were used, with the supply voltage about 11.0 V instead of 12 V, presumably putting the noisy internal regulator out of operation. All mains connections were isolated using

† email address christopher.gough@psi.ch

Copyright © 2017 CC-BY-3.0 and by the respective authors

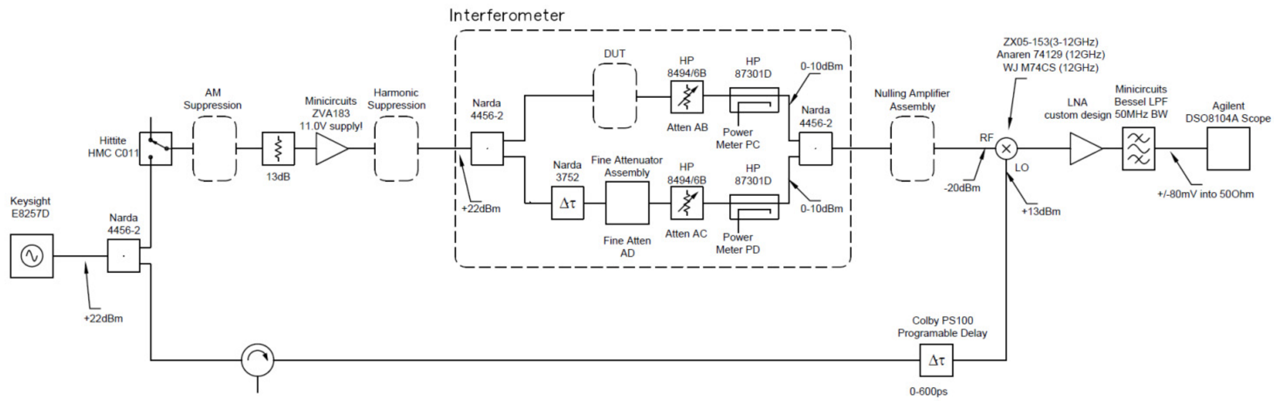

Figure 1: Simplified hardware schematic for the Amplifier Measurement System (AMS).

low capacitance isolation transformers from Fug GmbH. All low voltage supplies were provided using low capacitance "medical grade" DC-DC convertors from Traco Electronic AG. Computer control was via an isolated USB - GPIB convertor.

In Fig. 1, a single mixer is used with a CW signal on the Local Oscillator (LO) port. This LO signal is swept through 360 degree RF in small steps by programming a delay line. The RF signal for the amplifier under test is split from the LO signal and is pulsed on and off for a few microseconds by a GaAs switch.

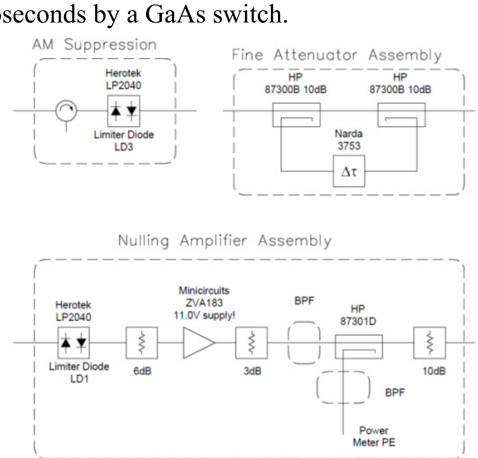

Figure 2: Details of Fig. 1.

In Fig. 2, the nulling amplifier and mixer are protected from burnout by a Herotek limiter diode and fixed attenuators. The fine attenuator AD and the phase shifter are then finely adjusted to find the null. This procedure relies on three power meters (PC, PD and PE) to guide the user.

With the reference arm of the interferometer switched out (Attenuator AC at full attenuation), the full scale signal is measured on the oscilloscope. The Attenuator AB is then reduced by 40 dB (factor of 100 in voltage) and the reference arm is switched in and balanced, the signal is within the linear range of the oscilloscope but limited to ±1% dynamic range. The absolute attenuation and phase on either DUT or reference arm is not important, just the value 40 dB. With the nulled signal somewhere on screen on the oscilloscope, successive pulse waveforms (typically N = 100 taken at 100 Hz) are downloaded to a computer. All waveforms for all LO phase angles are downloaded, stored in ASCII format and analysed for jitter.

The mixer output waveforms are in the millivolt range and near to DC so these are vulnerable to LF noise. The mixer is mounted in a solid copper housing to bypass ground currents away from the mixer diodes. The LF LNA uses four parallel channels of low 1/f noise amplifiers with modest gain and ±4 V output swing that is soft clamped to ±1 V. This clamping minimises overdrive transients in the LNA and scope. The four parallel channels are connected to the four scope input channels and these channels are summed numerically by the scope to increase the effective dynamic range by a factor of ~two.

Without the interferometer operating, parts of this system are also used measurement of amplifier gain curve, RF power stability and pulse shape. Extensive software development was also needed to configure the instruments and download/analyse the oscilloscope waveforms.

The AMS was used to test amplifiers at 3, 5.7 and 12 GHz. In principle, there is no layout change with frequency change, due to the use of very broad band components. Unfortunately in practice, some component changes are needed for different frequencies: the high power directional couplers on the DUT amplifier output; the isolator before the Limiter Diode LD3 and Colby delay line; the narrow band filter after the interferometer; and the mixer for highest sensitivity, especially at 12 GHz.

A special feature of testing power amplifiers is that these are often operated approaching saturation, giving variable suppression of source AM on the DUT arm of the interferometer while leaving the source AM on the reference arm unchanged. To reduce this effect, source AM suppression diodes before the interferometer are needed. Approaching amplifier saturation, power meters

Copyright © 2017 CC-BY-3.0 and by the respective authors

ISBN 978-3-95450-182-3

PC and PD have varying harmonics applied and the power meter readings during interferometer balancing may be wrong by up to 1 dB. This is annoying but does not affect the jitter measurement which is centered on the fundamental.

RESULTS

In principle, the interferometer balancing could be fully automated and the procedure could be completed in a few seconds. In practice, there is a continuing struggle to have reproducible results and measurement during a 10 minute period is needed to give some certainty about the result. This means the slow sweep over all RF phases is not a disadvantage.

In Figs. 3-5, the LO phase is swept 360 degree RF in one direction, then swept back in the other direction. When the two curves overlap, this gives confidence that there was no systematic drift in jitter which may occur on the time scale of minutes. The jitter difference at each phase is a measure of short-term reproducibility. The presence of characteristic spikes on the jitter plot indicates different types of intermittent contact problems in RF components.

In Fig. 3, the two peaks in this plot of ~35 ppm correspond to AM measured on positive and negative extrema of the mixer DC characteristic. The two troughs of ~18 µrad correspond to the phase jitter on the positive and negative going zero crossing. The phase is expressed in units of urad, having the same sideband power as for amplitude noise measured in units of ppm. Experience shows that a power amplifier adds AM through gain changes much more than through random additive noise, so AM is always larger than PM.

When the jitter curve is flat, this means there is no difference between AM and PM indicating that the system is measuring random noise with no correlation between upper and lower frequency sidebands. This is an ideal result and is usually at the noise floor of the AMS, meaning the amplifier is not adding measurable jitter.

CONCLUSIONS

The AMS measures <10 ppm rms amplitude and <10 µrad phase at any RF frequency 3 – 12 GHz and any pulse length or reprate. The AMS could be upgraded for higher sensitivity but this would also give reduced dynamic range and increased sensitivity to timing jitter. Low jitter from 500 W-class amplifiers is critically dependent upon low flicker noise biasing for transistors, stable mechanical design of RF mounting and enclosures and control of SMPS noise. Poor design or poor measurement can easily give non-reproducible jitter values in the 200-1000 ppm range.

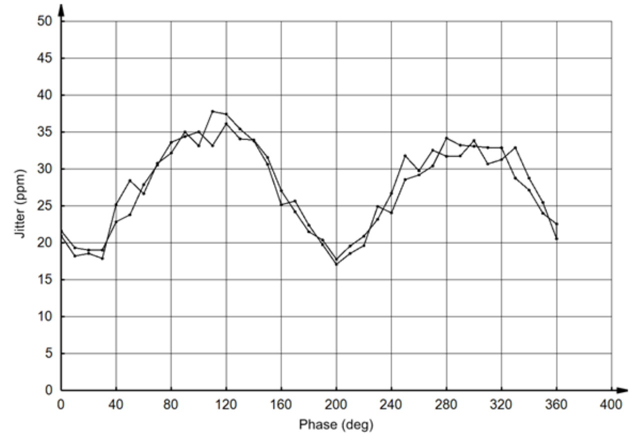

Figure 3: Jitter Measurement for S-band amplifier (PSI in-house development) at 304 W (AM = 35 ppm, PM = 18 µrad).

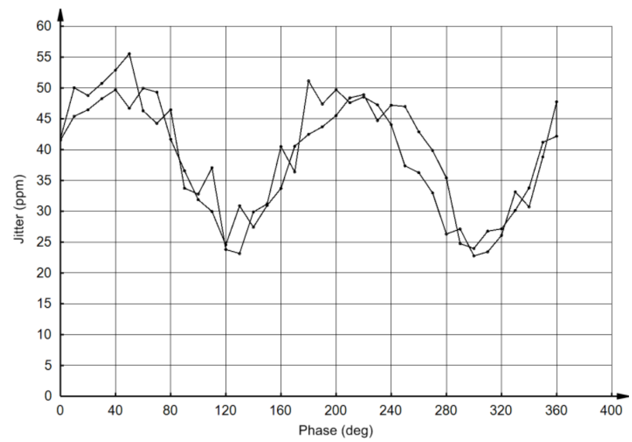

Figure 4: Jitter Measurement for commercial C-band amplifier at 450 W (AM = 50 ppm, PM = 27 µrad).

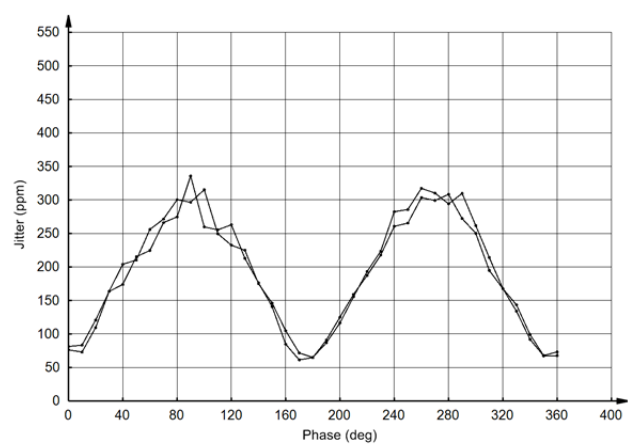

Figure 5: Jitter Measurement for commercial X-band SS amplifier at 330 W (AM = 320 ppm, PM = 60 µrad).

Copyright © 2017 CC-BY-3.0 and by the respective authors

REFERENCES

[1] Thomas R. Faulkner and Robert E. Temple, “Residual phase noise AM noise measurements and techniques”, HP Application Note A-116, 2001 (Part No. 03048-90011).

[2] *“Pulsed Carrier Phase Noise Measurements using Agilent E5500 Series Solutions*”, Agilent PN E5500-1, Jan. 2000.

[3] E. Rubiola and V. Giordano, “Advanced interferometric phase and amplitude noise measurements”, *Rev. Sci. Instr,.* vol. 73(6), June 2002, pp. 2445-2457.

[4] M. Paraliev and C. Gough, “Comparison of high resolution balanced and direct conversion measurement of SwissFEL resonant kicker amplitude”, *in Proc. 2015 IEEE Pulsed Power Conf.,* Austin, USA, May 31-June 4 2015.

[5] K. Sann, “The measurement of near-carrier noise in microwave amplifiers”, *IEEE Trans. on Microwave Theory and Techniques*, vol. MTT-16, no. 9, Sep. 1968, pp. 761-766.

Copyright © 2017 CC-BY-3.0 and by the respective authors

ISBN 978-3-95450-182-3