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DESIGN AND IMPLEMENTATION OF A HIGH DYNAMIC RANGE C BAND DOWN-CONVERTER

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Abstract—A technique that expands dynamic range (DR) of frequency down-converters in the C band frequency is presented. Primary characteristics of down-converter are evaluated to confirm that it can be used in microwave receivers. The C band down-converter is carried out by the combination of RF mixers, band pass interdigital filter, and X band combline filter which are designed entirely for this project. Attainment of the perfect receiver is the final purpose of this paper, and a method that causes 72 dB dynamic range, high tangential signal sensitivity and fine gain flatness is used for achieving the mentioned purpose. These efforts improve the dynamic range about 19 dB and gain flatness about 3.07 dB.

1. INTRODUCTION

Frequency down-converters are inseparable parts of the microwave receivers, and nowadays increased demands for various kinds of receivers result in the development of all sections and subsections in new receivers. Indeed, frequency down-converters are the main part of the receivers such as super heterodyne receivers and direction finding receivers etc. In addition, most of the receivers have sections that play a role as a down-convertor. For example, the super heterodyne receiver has been used in a wide variety of applications. This kind of receiver is desirable because it can offer high sensitivity and excellent frequency selectivity, which are vast important regarding in the competition among RF designers. All these characteristics indicate the importance of precise down-convertor [1].

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Dynamic range (DR) is one of the most important performance indexes, and the larger DR brings more flexibility to receiver. Expanding a receiver's DR has attracted great attention, and current research focuses on analog-to-digital converter (ADC), e.g., improving its performance [2] or using parallel sampling method [3], etc. Recently, reference [4] proposed a structure of two low noise amplifiers (LNA), which paid attention to another bottleneck of receiver DR — RF (Radio Frequency) front-end.

Down-converters should be examined by any nonlinear method that can support two tones as input signals, one of which is very large (the local oscillator) and the other very small (the input signal). Consequently, time-domain direct integration or harmonic balance methods are suitable algorithms for nonlinear and multi-tone synthesizes. Designing a C band frequency down-converter is the main purpose of this paper, which transfers input frequency to intermediate frequency where we can analyze its output with some processors. Finally, the important characteristics of implemented module are tested to confirm that this module can be used in receivers. These features are TSS, DR, $P_{1\,\mathrm{dB}}$ and output flatness.

2. DESIGN PROCEDURE

The frequency down-converters transfer the input RF signal to an intermediate frequency in which amplifiers and narrow-bandwidth filters are easily available. Figure 1 shows the circuit diagram of the frequency down-converters. The proposed down-converter has three stages that multiply input frequency by local oscillators. Any frequency in C band is transferred to 9.5 GHz with PLL synthesizer (LO1). The first LO is used for coarse tuning of signals. This PLL synthesizer works from 13.5 GHz to 17.5 GHz with step frequency of 20 MHz and an output level which can drive mixer and acceptable

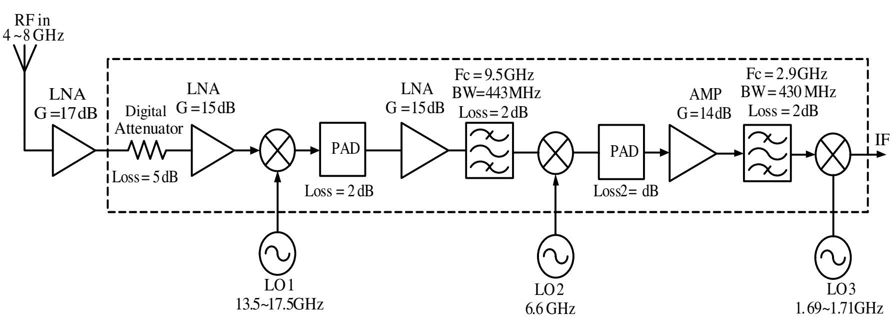

Figure 1. Circuit diagram of frequency down-converter.

phase noise. After that, RF signal is transferred to 2.9 GHz with local oscillator (LO2), then multiplied by 1.7 GHz with the last synthesizer (LO3). This procedure transfers C band frequencies to 1.2 GHz and contains three stages: one up converting stage and two down converting stages. The performance of the down-converter is significantly affected by the frequency conversion stages.

When strong signals are received, the input third-order intercept point (IIP_3) is the major factor that affects DR upper bound. With input power increasing, the undesirable third order product grows 3 times as fast as the fundamental wave. When the input signal power gets close to IIP_3, IM3 is comparable to fundamental wave, which will result in serious demodulation performance degradation and significant bit error rate [5].

The total IIP_3 of scheme for attenuator (AA) cascading with LNA, i.e., (IIP_3) total, can be derived as:

$$\frac{1}{(IIP_3)_{\text{total}}}=\frac{1}{(IIP_3)_{AA}}+\frac{G_{AA}}{(IIP_3)_{LNA}}=\frac{1}{(IIP_3)_{AA}}+\frac{1}{(IIP_3)_{LNA}ATT} \quad (1)$$

where, G_{AA} is the power gain of AA, ATT the attenuation of AA, and $ATT > 1$, meaning that $GAA < 1$.

It is clear that if ATT satisfies:

$$ATT > \frac{(IIP_3)_{AA}}{(IIP_3)_{AA}-(IIP_3)_{LNA}} \quad (2)$$

$(IIP_3)_{\text{total}}$ would be larger than $(IIP_3)_{LNA}$; in other words, the total IIP_3 gets rid of the restriction of LNA. If $(IIP_3)_{AA} >> (IIP_3)_{LNA}$, $(IIP_3)_{\text{total}}$ would be larger than $(IIP_3)_{LNA}$ whether (2) was satisfied or not. $(IIP_3)_{\text{total}}$ grows higher with the increasing of ATT, and the ultimate value is $(IIP_3)_{AA}$. This scheme overcomes such a restriction, and improves the ability to receive strong signals, in other words, extends DR upper bound. Furthermore, the enhanced value is only related to AA while having nothing to do with LNA.

Using a digital attenuator at the input of module before LNA is one of the biggest advantages of this design which can improve the upper saturation limit, and, as a result it expands the dynamic range. The digital attenuator provides attenuation among 0 dB to 31 dB. Also an equalizer is used in the first section of receiver to get a good flatness.

With the aid of harmonic balance algorithm, intermodulation frequencies that are produced by mixers in IF band frequency are analyzed. Figure 2 illustrates output power when frequency of input port is swept among 4 GHz to 8 GHz with constant LO frequency. Critical area for operation is the region where power difference of intermodulation harmonics is about 30 dB.

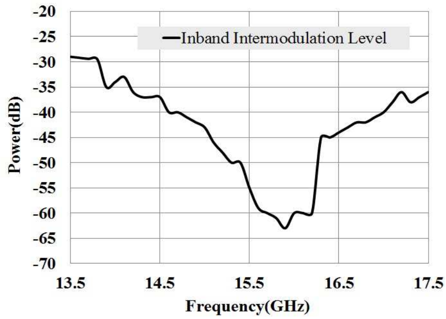

Figure 2. Output power vs. first local oscillator when RF input is swept between 4 GHz to 8 GHz.

Table 1. Comparison of measured $P_{1\,\mathrm{dB}}$.

Frequency (GHz)	4	5	6	7	8
$P_{1\,\mathrm{dB}}$ (dBm) @ 0 dB Attenuation	−17	−17	−16	−15	−13
$P_{1\,\mathrm{dB}}$ (dBm) @ 31 dB Attenuation	3	3	3	2	2

3. EXPERIMENTAL RESULTS

This module is a combination of amplifiers, mixers, attenuators, and filters, integrated for miniaturizing and attaining best response. Therefore, each of these parts will affect our results. For example, the efficiency of an amplifier is limited by the saturation of the output power because of nonlinear voltage and current limiting phenomenon. The distortion can be so high that it degrades the quality of the signal beyond acceptable levels. Therefore, distortion must be defined and evaluated, and usually it is one of the design specifications of power amplifiers. At first, saturation point is computed by gain saturation test with 0 dB attenuation, and secondly, it will be repeated for maximum attenuation. Therefore, we have two dynamic ranges, one of which is an instantaneous dynamic range and the other one an extended dynamic range. Saturation tests are done in C band frequencies by 1 GHz step frequency. Figure 3 illustrates the measurement result of saturation test. Comparing the results of these experiments shows that using the 31 dB attenuation can improve $P_{1\,\mathrm{dB}}$ about 20 dB. In Table 1, a summary of these tests are presented.

Sensitivity, an important parameter for a receiver, is affected by all used components. Sensitivity in a receiver is normally as the minimum input signal ($S_{\min}$) required for producing a specified output signal having a specified signal-to-noise ratio and is defined as minimum

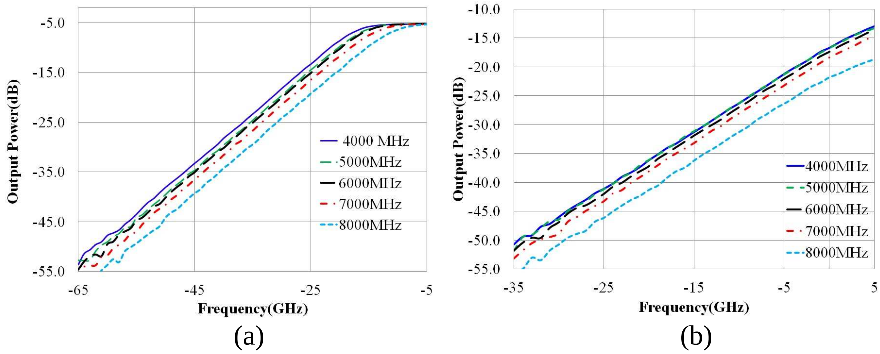

Figure 3. Output power vs. available input power (a) for 0 dB attenuation (b) for 31 dB attenuation.

signal-to-noise ratio times the mean noise power (see Equation (3) [9]).

$$S_{\min}(\text{dBm}) = -174(\text{dBm/Hz}) + NF(\text{dB}) + 10\log B(\text{Hz}) + (SNR)_{\min}(\text{dB}) \tag{3}$$

where, $S_{\min}$ is the receiver sensitivity; NF is the noise figure of the receivers; B is the effective bandwidth of the receivers; $(SNR)_{\min}$ is the required minimum signal-to-noise ratio (SNR) to process a signal. The value of noise figure is 6.34 dB, acceptable in this application.

PMDS is the minimum detectable signal, 3 dB above the noise floor in this equation, and tangential sensitivity (TSS) is the point where the top of the noise level with no signal applied is level with the bottom of the noise level on a pulse. It can be determined in the laboratory by varying the amplitude of the input pulse until the stated criterion is reached. The signal power is nominally 8 ± 1 dB above the noise level at TSS point. In addition, TSS depends on the RF bandwidth, video bandwidth, noise figure, and detector characteristics [9].

$$\text{DR} = P_{1\text{ dB}} - \text{PMDS} \tag{4}$$

Finally the output of down-converter is connected to DLVA which has more significant sensitivity and noise figure than the implemented module. Pulse modulation with 1 µs pulse width and 100 µs pulse period is imposed to input port of module, then detected signal is displayed on signal oscilloscope with IF bandwidth equal to 1 MHz. TSS point will be obtained from this method. Figure 4 shows the output pulse of DLVA in TSS point for input frequency equal to 6 GHz. In this situation, the minimum input power is -83 dBm when output of DLVA is in TSS point. According to the principle that TSS point is about 8 dB above the noise floor and PMDS about 3 dB above noise floor, dynamic range can be defined as the difference between TSS point

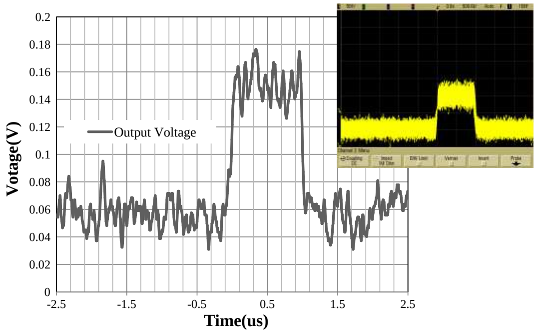

Figure 4. Output pulse of DLVA in TSS point.

and $P_{1\,\mathrm{dB}}$ in input power, based on Equation (5). As a result, DR of this module is 91 dB with attenuator (extended) and 72 dB without attenuator (instantaneous) at 6 GHz, thus we can improve DR about 19 dB with this proposed method.

$$\mathrm{DR} = P_{1\ \mathrm{dB}} - (\mathrm{TSS} - 5\ \mathrm{dB}) \tag{5}$$

The IF filters, following the mixers, are band pass filters used to pass the desired IF signal and reject all other frequencies generated in the mixers. Also all filters affect output response specially output flatness, so evaluations of filter responses are necessary. After up converting stage, the IF frequency is filtered by band pass filters. Center frequency of this traditional combline filter is 9.5 GHz with 443 MHz bandwidth (3 dB). The mentioned filter is implemented in round road structure which has good rejection in out of band and low insertion loss in band pass region and reduces frequencies due to inter modulations of mixer. Figure 5 shows the measured results of combline filter and photograph of the implemented combline filter. It can be seen that the minimum insertion loss in the passband is 1.91 dB at 9.5 GHz.

Filtered signal is converted to 2.9 GHz with multiplying by 6.6 GHz local oscillator and then filtering and amplifying will be done again. A 2.9 GHz filter, a microstrip interdigital filter, is implemented on TMM4 with 20 mil thickness. This filter has 430 MHz bandwidth with the minimum insertion loss 2.24 dB in the passband at 2.9 GHz. Figure 6 shows the simulated and measured results of 2.9 GHz filter and photograph of the implemented interdigital filter. The results of test are similar to the results of theoretical simulation except some differences generated by low accuracy in construction process. Output frequency of the last filter is transferred to 1.2 GHz with PLL synthesizer (LO3). Last synthesizer that works at 1.69 GHz to 1.71 GHz with 100 KHz step frequency is used for fine tuning of signals. Except the interdigital filter, all other parts of this module is implemented on TMM10I substrate with 15 mil thickness.

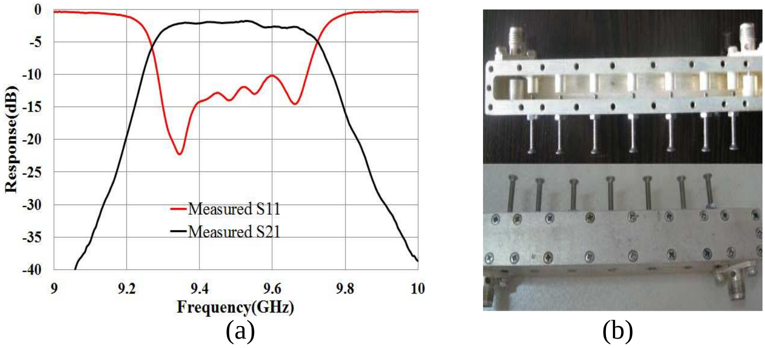

Figure 5. (a) Measured results of 9.5 GHz band pass filter. (b) Photograph of the implemented combline filter.

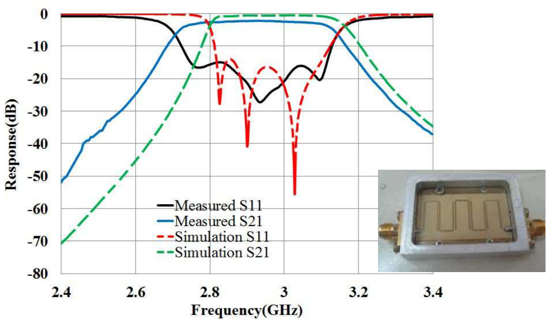

Figure 6. Measured and simulation results of 2.9 GHz filter.

The last but not least important characteristic of down-converter is output flatness which means that the output level of this module should not alter from a defined limit. According to this fact all used elements in this study should have specific flatness, and the flatness between ± 3 dB is acceptable. The optimum flatness of output should follow a linear behavior which states that any nonlinear behavior in the C band is not acceptable. In this situation, the flatness characteristic is improved by using an equalizer in the input port. Figure 7 shows the measured results of the output flatness in the C band with and without an equalizer. The total amount of ripple in output in the pass band is 4.68 dB. The ripple improves about 3.07 dB by using 6 dB equalizer. The maximum ripple with equalizer is 1.61 dB. Input level for this measurement is -40 dBm. Figure 8 shows constructed module that has 6.35 cm width and 11.5 cm length.

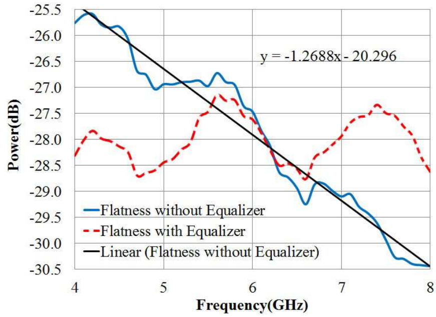

Figure 7. Output power vs. input frequency with and without equalizer.

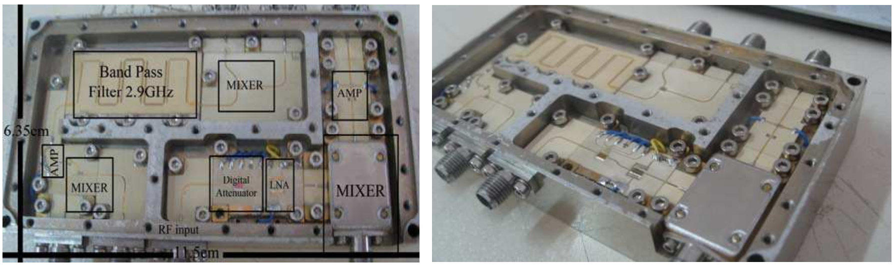

Figure 8. Evaluation board for C band down-converter.

4. CONCLUSION

The implemented down-converter in this paper provides a high dynamic range for wide band radar receivers. Selecting proper frequencies for LO and IF by using harmonic balance algorithms yield to convenient results. A series of tests are performed to determine the down-converter characteristics of receivers. As a result, the obtained values of gain, saturation, dynamic range, TSS and flatness of suggested structure are more than adequate for this purpose. This structure could be used in some radar systems. These results are in relatively close agreement with estimated performance expectations.

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