Design of a Ka Band Multi-Channel (25CH) Millimeter-Wave Downconverter for a Signal Intelligence System

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Abstract

In this study, we propose an approach for the design and satisfy the requirements of the fabrication of a reliable and stable high-frequency downconverter for the millimeter-wave (Ka band) and detail the contents of the approach. We design and fabricate a stable downconverter with a low noise figure, flat gain characteristics, and multi-channel characteristics suitable for millimeter-wave bands. The method uses the chip-and-wire process for the assembly and operation of a bare MMIC device into the RF path. To compensate for the mismatch among the many components used in the module, W/G transition, an image rejection mixer, a switch, and an amplifier suitable for millimeter-wave frequency characteristics are designed and applied to the downconverter. To reject the spurious signals generated from the complex local oscillation signals, the downconverter is designed to not affect the RF path. In the Ka-band downconverter, the gain is measured from 41.89 dB to 42.83 dB at 33–35 GHz with flatness of about 0.94 dB. The measured value of the noise figure at CH1 is 4.936 dB with a maximum value in the 0.75–1.25 GHz intermediate frequency. The third intermodulation measurement result is 61.83 dBc under a -50 dBm input power and above gain, and the switching to select a channel takes about 622 μs.

Key Words: Downconverter, Front-End, Ka Band, Low Noise Figure, Millimeter-Wave, SIGINT System.

I. INTRODUCTION

An aircraft signal intelligence (SIGINT) system is used strategically and tactically to gather signals from radars and communications systems. SIGINT systems are commonly divided into communications intelligence and electronic intelligence systems. As SIGINT systems require the ability to locate the source of enemy signals [1], direction finding (DF) to support this capability is an important requirement, and SIGINT systems generally operate in a wide frequency range of 20 MHz–40 GHz [2].

For electronic warfare (EW) requirements covering the Ka-band spectrum, a channelized receiver is highly desirable. The Future Combat Systems (FCS) radar and communications allocations exist in this band of the spectrum [3].

Broadband receivers with wide millimeter-wave bands are essentially required for electronic support systems. Millimeter-wave systems with at least 30 GHz millimeter-wave bandwidths with broad dynamic ranges, improved sensitivity, and front-end complexity are commonly required in the fields of communication, radar, and wireless communication. To satisfy such high requirements, general millimeter-wave systems include local oscillator (LO) and downconverter functions in their front-end receiver connected to an antenna. Moreover, the mixer, amplifier, and the many necessary components generally require multichannels in which the frequencies have been converted into sub-octave bands to satisfy the system requirements because of their bandwidth and dynamic range limits. Channelized receivers divided into multiple channels can be optimized to satisfy the
requirements of ESM receivers. The problems with this archi-
tecture include the fact that the multiplied local oscillation sig-
nals are fed back into the antenna path because of the low isola-
tion between the RF and LO ports of the mixer and the re-
duced dynamic range due to the cascade structure that down-
converts the frequencies [4–7].

Although numerous studies have been conducted to design
millimeter-wave ultra-wideband receivers suitable for electronic
support systems, many constraints remain on the design and
manufacturing of such broadband downconverters. To make the
receivers have low noise figures, flat gain characteristics in the
band, wide dynamic ranges, and band-specific frequencies that
can be selected with low-loss transmission lines, a local oscillator
circuit necessary for frequency conversion should be incorpo-
rated in the module, and a closed structure is needed to avoid
signal interference from other devices (IFF transponder, Dop-
pler radar, communication system, etc.) in the system environ-
ment where the receivers are installed [8, 9].

In this paper, a design approach is proposed not only to im-
prove the signal acquisition accuracy by maximizing the number
of receiving input ports (25 input channels) but also to convert
the signal frequencies of 33–35 GHz in the millimeter-wave
Ka-band into the intermediate frequency (IF) 1 GHz (instanta-
aneous frequency: 500 MHz) before the signals are delivered to
the signal processing unit. The content of the design approach
is as follows.

First, as the 33–35 GHz band received by an antenna th-
rough the free space medium has a waveguide structure, a trans-
ition structure should be designed to convert the channels into
micro-strip transmission lines. To implement 25 channels in the
form of grids in a limited space, the channels are designed using
the Rogers RT/duroid 5880 5-mil PCB material for small in-
sertion losses and better impedance matching.

Second, the signals are downconverted into IF signals with a
center frequency of 1 GHz and a frequency band of 500 MHz
through a low-noise amplifier (LNA) and a frequency converter
(packaging type) including an image suppression function. At
this time, the local oscillator signals necessary for frequency con-
version are phase locked with the internal voltage-controlled
oscillator (VCO) with an 8 GHz band using the external refer-
ence frequency of 1 GHz supplied by the system. As the voltage
turning (VT) voltage of the VCO is high, the structure of the
loop filter connected to the phase-locked loop (PLL) output
terminal was designed as an active type instead of a passive type.

Third, as the RF lines must be designed considering the am-
plitude matching among multiple channels (25 channels), the
RF lines of the individual channels are designed not only to be
electrically the same but also to have the same component
mounting positions and wire bonding length.

Fourth, to secure the degree of isolation between the input
and output channels, the mechanism is designed to prevent sig-
als from flowing into other channels using valley form equip-
ment.

II. MILLIMETER-WAVE DOWNCONVERTER STRUCTURE

In this paper, a Ka-band multi-channel (25) millimeter-wave
downconverter module mounted on the front-end terminal of
an antenna is designed. This module has a complicated struc-
ture with 25 RF input (33–35 GHz) channels and 16 IF output
(0.75–1.25 GHz) channels. The module is divided into module
#1 and module #2 when it is mounted because of space re-
strictions for the entire system and to secure excellent noise fig-
ures. Module #1 is mounted near the antenna. A conceptual
diagram of the entire system is shown in Fig. 1.

Fig. 2 is a detailed block diagram of the multi-channel down-

![Fig. 1. Configuration of the entire millimeter-wave downconverter.](image-url)
converter necessary for the SIGINT system. The multi-channel
downconverter amplifies signals with a center frequency of 34
GHz (bandwidth 2 GHz) received from the antenna using an
LNA and converts the amplified signals into IF signals with a
center frequency of 1 GHz (RF bandwidth 2 GHz) through a
frequency converter (MIXER), which includes an image sup-
pression function. At this time, the local oscillator signals,
which are necessary for frequency conversion having a frequency
sweeping function with a 1 GHz bandwidth, are phase locked
with the internal VCO with an 8 GHz band using the external
reference frequency of 1 GHz supplied by the system, converted
to have a 16 GHz band through a frequency doubler, and input
into the LO port of the MIXER.

The frequency converter (MIXER) is a component that con-
tains a doubler inside, and the 16 GHz band is converted into a
32 GHz bandwidth inside the MIXER. In this case, the fre-
quency converter has a frequency variable function, so that the
local oscillator signals have a bandwidth of about 1.5 GHz (after
multiplying).

III. DESIGN AND SIMULATION

1. W/G to M/S Line Transition Simulation

In most millimeter systems, the waveguide and the micro-
strip are the two commonly used transmission lines. The wave-
guide is usually employed to connect the antenna and the mil-
limeter receiver or transmitter because of its low insertion loss
[10]. To transmit signals in the super high 34 GHz frequency
band from the waveguide (W/G) form to the micro-strip line
structure, the electric field mode must be converted through a
transition design.

As shown in Fig. 3(a), a transmission line with a single-layer
structure was placed at a point of the height of the Lambda/4
(about 1.7 mm) of the center frequency from the bottom surface
of the instrument of the W/G structure. For fine impedance
matching, signals were simulated in the form of patches at the
end of the transmission line, where the electric field is propagat-
ed. As shown in Fig. 3(b), the design results indicated the inser-
tion loss was not larger than 0.1 dB and that the reflection coef-
icient was not larger than -23 dB in the 32–36 GHz band.

To relieve the problems by comparing the design value and
the actual measured value before installing on the module, a jig
was fabricated to make back-to-back measurement possible
using an Rogers RT/duroid 5880 PCB with permittivity of 2.2,
as shown in Fig. 4(a). As a result of the measurement in Fig.
4(b), insertion losses not exceeding -1 dB and reflection coeffi-
cients not exceeding -13 dB were obtained. The reason why the
measurement graph was wider by 1 GHz upward and down-
ward from the band (33–35 GHz) was that the frequency drift
was considered because of the errors that could occur in the ac-
tual fabrication.

2. PLL Design for the LO Section

To design the PLL local oscillation that is phase-synch-
ronized with the external reference frequency, the external REF applied to the REF port of the PLLIC was injected after converting it into 200 MHz using a five-divider, and the LO frequency in the 16 GHz band was created in the VCO output using a doubler.

As shown in Fig. 5, to synchronize the phase with the phase-locked external reference frequency, the output frequency of the VCO was N-divided in the PLLIC, the phase was compared with the reference frequency in the phase detector (PD), the error voltage corresponding to the phase difference between the two signals was outputted, and the VT terminal of the VCO was controlled through the secondary passive loop filter, which has the properties of a low-pass filter [11]. In this case, as the maximum frequency input into the RF port of the PLLIC is generally not higher than several GHz, although it may differ by PLLIC, the VCO output frequency should be made to be inputted after two or four divisions using a prescaler. Likewise, as the maximum frequency allowed to the RF port of the PLLIC is not higher than several hundred MHz even when the external reference frequency provided is high, the external reference frequency should be made to be inputted after five divisions using a prescaler [12].

The equation for the VCO frequency phase synchronized with the external reference frequency is expressed in Eq. (1).

\[ f_{VCO} = [(P \times B) + A] \times \frac{f_{REFIN}}{R}, \tag{1} \]

where

- \( f_{VCO} \): Output frequency of the VCO,
- \( P \): Dual modulus prescaler (8/9 or 16/17),
- \( B \): Division ratio of the 13-bit counter,
- \( A \): Division ratio of the 6-bit counter,
- \( f_{REFIN} \): External reference frequency (here, frequency after the prescaler),
- \( R \): Division ratio of the external reference frequency.

The final analyzed phase noise values in the 16 GHz band were expected to be -78.5 dBc/Hz at 1 kHz offset and -81.5 dBc/Hz at 10 kHz offset, and the value of the reference phase noise was a five-divided value from a 1 GHz external reference.

The overall phase noise of the analog PLL (Table 1) is as follows:
The phase noise of the final output was calculated by Eq. (2)

\[ Total\ PN = 10 \cdot \log \left( 10^{\frac{VCO_{PN}}{10}} + 10^{\frac{REF_{PN}}{10}} + 10^{\frac{PS_{PN}}{10}} + 10^{\frac{PD_{PN}}{10}} \right). \]  

The phase noise of the final output was calculated by Eq. (2) [13].

The local oscillation signals in the 16 GHz band generated as such were inputted into the frequency mixer for frequency conversion. As the mixer included a doubler, the actual local oscillation signals were in the 32 GHz band.

3. Analysis of Noise Source in Low Dropout Regulators

The difference between insignificant noises and significant noises is the degree to which the noises affect the operation of the circuit in question. For example, a switching power supply has a significant amount of output voltage ripple at 3 MHz. If the circuit power by it has a bandwidth of only a few hertz, such as a temperature sensor, this ripple may be of no consequence.

Conversely, if the same switching power supply powers an RF PLL, the result can be quite different [14].

A low dropout (LDO) that shows excellent power supply rejection ratios (PSRR) in PLL circuits using VCO was applied. Fig. 6 shows the PSRR characteristics of LDOs with improved PSRR characteristics. The phase noise characteristic value expected at 10 kHz offset points was -81.5 dBc/Hz. Selecting and applying LDOs with lower characteristic values than the previous is important.

4. Design for the Synchronized Main Path versus Monitoring Port

A switch (SP2ST) was designed so that when the signals received through 25 channels were converted into IF frequencies and transmitted to the signal processing area of system #1, a power divider was applied to each main path output to sequentially select the 25 paths. A switch was provided to a separate system #2 to design a monitoring port that could determine the signal intensity and the channels where the signals come in Figs. 7 and 8 illustrate the block diagram and simulation results.

In addition, a circuit was designed to supply the synchronized control signals that fit the monitoring port. As timing is important when selecting channels 1–25, the switching time should be considered when selecting the RF switch.

![Characteristic graphs by the frequency offset of LDOs with excellent PSRR characteristics.](image-url)
IV. FABRICATION AND MEASUREMENT

A millimeter-wave Ka downconverter, which has 25 RF input ports (33–35 GHz), 16 output ports (0.75–1.25 GHz), and one external reference frequency (1 GHz) input port, consisted of module #1 (millimeter-wave unit) and module #2 (control and switching unit). Module #2 was designed and fabricated in the form of four submodules. Fig. 9(a) illustrates the entire setup in which the two modules are connected to each together.

Fig. 9(b) is a photograph of the inside of module #1, which is an area where the RF frequency is mixed with the LO frequency in the mixer (Hittite’s HMC1065LP4E), which includes an image rejection function to output IF signals in the 1 GHz band. The PCB was designed to be installed using the epoxy on the entire bottom surface of the equipment, so that the frequency in the Ka-band could be transmitted through the transmission line without any loss or distortion.

Fig. 9. (a) Photograph of the complete setup (from 33–35 GHz to 1 GHz). (b) Photograph of module #1 (RF section). (c) Photograph of module #1 (DC section).
RT/duroid 5880 5-mil PCB was used up to the mixer stage where the Ka band exists, and Rogers RO4003 20-mil PCB was used at low frequencies in the L band converted into IF to make SMT possible. 

Fig. 10(a) shows the inside of module #2, Fig. 10(b) the PLL board including the PLLIC and VCO, (c) The IF board including the AMP and filter, (d) The switch module including the distribution board.

Fig. 11. Results of the measurement of gain using a Scalar Network Analyzer: (a) 33.0–33.5 GHz band, (b) 33.5–34.0 GHz band, (c) 34.0–34.5 GHz band, and (d) 34.5–35.0 GHz band.
board that generates the LO signal needed to convert the RF frequency into the IF frequency, Fig. 10(c) the IF board for amplifying and filtering the IF signal, and Fig. 10(d) the distribution board in which 25 paths are distributed and switched to 16 paths.

1. **GAIN Measurement Results**

Fig. 11(a)–(d) show the total gain of channel 1 from the RF input port to the IF output port. As the bandwidth of the IF frequency was about 500 MHz, the total gain was measured four times after changing the RF and LO frequencies to 500 MHz.

The total gains were as follows:
- Up to 42.83 dB in the 33.0–33.5 GHz band
- Up to 42.15 dB in the 33.5–34.0 GHz band
- Up to 41.89 dB in the 34.0–34.5 GHz band
- Up to 42.26 dB in the 34.5–35.0 GHz band

2. **NF Measurement Results**

As for the results of the measurement of noise figures at the IF frequency in the L band that was changed from the Ka band, which is a receiving frequency band, the value measured at the 0.75–1.25 GHz band was 4.936 dB with a maximum value. This result is mainly due to the noise figure characteristics from the W/G transition structure to the mixer in which the LNA is included. Therefore, the characteristics vary greatly depending on how much the insertion loss is minimized, as shown in Fig. 12.

3. **Third Intermodulation Measurement Results**

In channel #1, which is representative, the third-order IMD characteristics that show the linearity of the module measured at the center frequency of 1 GHz of the IF frequency was measured as 61.83 dBc (Fig. 13).

4. **Switching Speed Measurement Results**

Switching speed is important at the system level because the speed of the switch used to select the RF path determines how fast the signal can be received. It took 622 μs to switch to another channel and select the channel through which the signals came in Fig. 14.

V. CONCLUSION

In this paper, we designed and fabricated a Ka-band downconverter module with a 25-channel RF input port with a low noise figure, flat gain characteristics, and reliability by applying the chip-and-wire process for assembly into the RF path and operation of a bare-type MMIC device. To compensate for the mismatch among the many components used in the module, W/G transition, an image rejection mixer, a switch, and an amplifier suitable for millimeter-wave frequency characteristics were designed and applied to the downconverter.

The main RF line was a dielectric substrate (RT/duroid 5880), which had a relative dielectric constant of 2.2 and a dielectric thickness of 0.127 mm, and Al2O3 (ceramic, ATC Co.), which had a relative dielectric constant of 9.8 and a dielectric thickness of 0.254 mm. In the Ka-band downconverter module, the gain was 41.89–42.83 dB at 33.0–35 GHz, with flatness of about 0.94 dB. The measured value of the noise figure at CH1 was 4.936 dB, with a maximum value in the 0.75–1.25 GHz IF frequency. The third intermodulation measurement result was 61.83 dBc. The switch to selecting a channel took 622 μs.
The millimeter-wave (Ka band, 33–35 GHz) 25CH down-converter module proposed in this paper is considered applicable to the modules that are installed at the rear end of the antenna of the SIGINT system in the field of EW to collect signals precisely. However, further studies on how to satisfy and increase the dynamic range still have to be conducted.

This work was funded by the technology development business fund of the Ministry of SMEs and Startups in 2019.

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