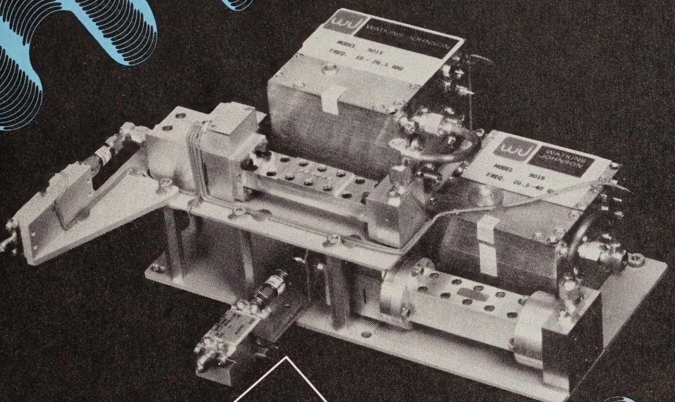
VOL. 16 NO. 6 NOVEMBER DECEMBER 1989

Millimeter-Wave Block Converters



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Tech-notes

This article describes millimeter-wave block converters for use at the input of microwave receivers. The addition of the block converter will extend the operating frequency of the receiver into the millimeter-wave frequency range. A millimeter-wave block converter accepts a range of frequencies (e.g., 26.5 GHz to 40 GHz) and converts it to a lower frequency range (below 18 GHz) so that existing receivers can be used for millimeter-wave surveillance. The design of this type of converter will be described, and the important parameters for block conversion will be discussed.

Signals of interest are continually being found at higher frequencies, with most activity occurring in the millimeter-wave band, up to 40 GHz. This indicates a need to extend the frequency range of microwave receiver stations to 40 GHz and above. With existing microwave receiver systems deployed to cover frequencies up to 18 GHz, a desirable way to build millimeter-wave receiver stations is to use these existing receivers with an upgrade that extends coverage into the millimeter-wave region. This can be achieved by adding, to the basic microwave receiver, a downconverter that changes signals within the 18 to 40 GHz range to new frequencies below 18 GHz so that the signals can be processed in the original microwave receiver.

Before discussing millimeter-wave downconverters in more detail, it should be noted that this technology, applied to existing radar warning receivers or other ESM systems, becomes a very cost-effective scheme. The existing receiver with its rf preselection, IF processing, digitizers, digital processing and other well-proven functions can be repeated or multiplexed to do the same tasks at millimeter-wave frequencies.

Although the "frequencies of interest" mentioned above are generally taken to be of military importance, we can expand the definition of them to include frequencies generally used in the rf community. This would include the use of block-conversion techniques for adding to the frequency coverage of instruments such as spectrum analyzers, vector network analyzers and other laboratory type equipment. The addition of a block converter to the input of a spectrum analyzer, for instance, would make it useful over the frequency range of the converter without generating as many annoying false signals compared with the usual harmonic mixer.

The specific implementation for a given system will depend upon several factors, some of which will be considered below.

Tuned versus Fixed Local Oscillator Downconverters

Signals within the 18 GHz to 40 GHz range can be converted to frequencies below 18 GHz using a frequency converter which has either a fixed or a tunable local oscillator (LO). As the microwave receiver already has highly accurate, synthesized LOs for processing of signals up to 18 GHz, it is desirable to use this capability in the design of the millimeter-wave converter.

A tunable LO with synthesizer characteristics capable of operation over this frequency range would be an expensive item and would duplicate some of the capabilities of the microwave receiver. A tunable LO would also require preselection to avoid spurious and image responses, and filtering of LO harmonics from the mixer would be very difficult. Therefore, the concept of a tunable LO for the millimeter-wave converter becomes unattractive, from both a technical and a cost point of view.

Fixed-frequency LOs are less costly to build, and require less filtering of their outputs. Signal processing in the receiver does not have to deal with a tunable LO, making the identification of converted signals easier. For cost and technical simplicity, the fixed LO frequency is preferred.

A block converter translates one range (block) of frequencies to a new block of frequencies, using a fixed LO. The actual bandwidth of the incoming block remains unchanged as a result of this conversion process. Accordingly, signal modulations or characteristic signatures remain unaffected. However, the frequency block may or may not be inverted (i.e., the high and low end of the original band may or may not be reversed), depending on the location of the LO frequency. If the LO is below the rf band, then there is no inversion of the signal band (i.e., IF = RF - LO); for the case when the LO frequency is above the rf band there is inversion (i.e., IF = LO - RF). The consequences of the changes in frequency, and possible inversions, can be removed by processing in the receiver. The block of frequencies between 18-GHz and 40-GHz is 22 GHz wide and cannot, therefore, be converted with a single LO frequency, unless the LO frequency is within the 22-GHz to 36-GHz range. With such a scheme, there would be loss of processing for rf signals around the local oscillator frequency due to the limited low-frequency capabilities of the microwave receiver. Ambiguities due to band folding and spurious signal generation together with the above limitations make this single block-converter approach technically impractical.

At least two blocks of rf frequencies are needed to convert the full 18-GHz to 40-GHz frequency range to new frequencies below 18 GHz. For convenience, the band is split at 26.5 GHz to coincide with the change of waveguide from K-band (WR42) to Ka band (WR28). This allows the use of full bandwidth K-band (18 GHz to 26.5 GHz) and Ka-band (26.5 GHz to 40 GHz) components in the design of the two converter assemblies. Typical specifications for a pair of block converters to meet the requirements of a millimeter-wave receiver system are given in Table 1, and a block diagram of one of these converters is shown in Figure 1.

The components used to meet the performance specifications given in Table 1 can be waveguide, MIC, stripline, finline or any transmission-line technology capable of operating at these high frequencies. Waveguide components that operate up to 40 GHz have been in existence for some time and their performance is usually the best in terms of conversion loss (due to the low-loss waveguide transmission lines). MIC components can operate at these frequencies, are usually a lot smaller, and can be manufactured in large volume at lower cost. MIC mixers tend not to perform as well as waveguide mixers, but can be integrated into smaller packages. However, the design engineering costs of a MIC circuit are high compared with the mature waveguide technology. While waveguide is much bulkier than MIC, these block converters, and other devices, can be manufactured from essentially standard components and assembled by bolting together waveguide flanges. If the receiver system can afford the size of the waveguide approach, there is no reason to expend the development effort to build a MIC converter that would be harder to maintain and would not have any performance advantages.

Local-Oscillator Sources

The source used to generate the LO signal for the block converters de-

K-Band Block Converter:

RF Input Range	18 to 26.5 GHz
LO Frequency	34.5 GHz (derived by tripling 11.5 GHz from a phase-locked source)
Phase-Lock Reference	5 MHz
IF Frequency Range	8.0 to 16.5 GHz
SSB Noise figure	19 dB, maximum
Conversion Gain (RF to IF)	16 dB, minimum
Conversion Gain Flatness	±2 dB @ const. temp.
Input 1-dB Compression Point	-6 dBm, minimum
Spurious Outputs (±mRF±nLO)	-25 dBc @ -10 dBm I/P
Level of 11.52 GHz at IF Output	-70 dBm, maximum
LO (34.5 GHz) Leakage at RF Port	-50 dBm, maximum

Ka-Band Block Converter:

RF Input Range	26.5 to 40 GHz
LO Frequency	44.5 GHz (derived by tripling 14.83 GHz from a phase-locked source)
Phase-Lock Reference	5 MHz
IF Frequency Range	4.5 to 18.0 GHz
SSB Noise figure	19 dB, maximum
Conversion Gain (RF to IF)	16 dB, minimum
Conversion Gain Flatness	±2 dB @ const. temp.
Input 1-dB Compression Point	-6 dBm, minimum
Spurious Outputs from (±mRF±nLO)	-25 dBc @ -10 dBm I/P
Level of 14.83 GHz at IF Output	-70 dBm, maximum
LO (44.5 GHz) Leakage at RF Port	-50 dBm maximum

Table 1. Typical specifications for K and Ka-band block converters.

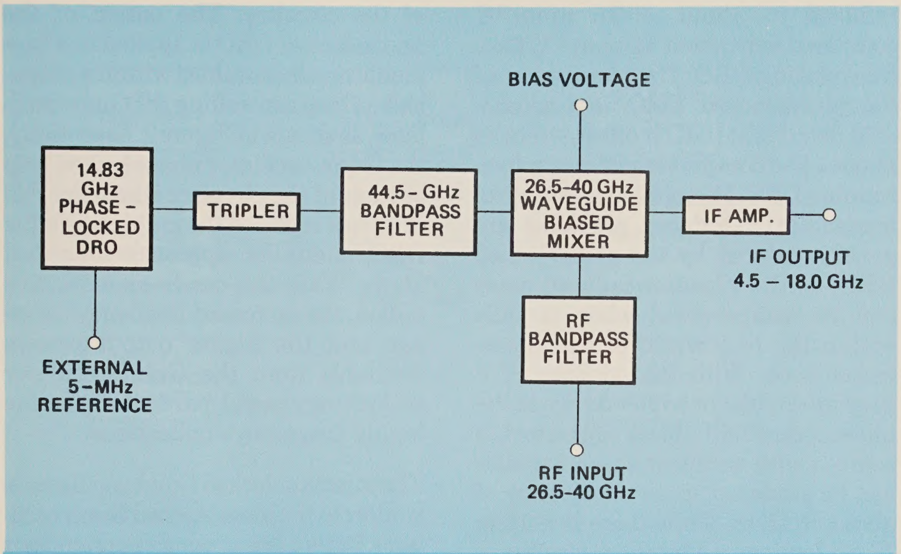

Figure 1. Typical Ka-band block converter.

scribed above should be very stable for receiver applications. There are several ways to generate LO signals suitable for millimeter-wave block converters, which are usually based on dielectric resonant oscillators and Gunn oscillators. Other forms of oscillators can be used but the above two types are the most popular and will be described below.

Dielectric Resonant Oscillators (DROs)

DROs are based upon a very stable, high permitivity dielectric puck to control the frequency of operation [1]. Fundamental DROs are normally limited to frequencies less than 30 GHz, because of limitations in the dielectric material and the oscillator circuitry used to support oscillations at the higher frequencies. A solution to the limited high-frequency operation of DROs is to follow the DRO with a multiplier, which was the technique used for the block converter shown in Figure 1. Although the frequency stability of DROs is good, drift problems and high close-in phase noise can be a problem with them in receiver applications. If the deficiencies of DROs are unacceptable, then a varactor diode can be coupled into the oscillator circuit to provide a limited amount of voltage-tuning. This voltage-tuned DRO can be used in a phase-lock circuit to lock the frequency of the DRO to a crystal reference. Crystal references around 100 MHz operating on the 5th overtone provide a good low phase-noise reference for the loop. With the circuitry shown in Figure 1, this 100-MHz crystal reference is itself phase locked to an external 5-MHz signal that the receiver system uses as a system-wide reference. Using this scheme, only the system 5-MHz reference needs to be accurately controlled, as all other reference signals are derived from this 5-MHz signal. The double phase-lock circuitry only allows certain frequencies, depending on the division factors in the two loops.

Using this type of DRO circuitry, the thermal sensitivity of DROs can be reduced to about ±0.25 ppm/°C compared with about ±5 ppm/°C for a free-running DRO. The phase noise of the phase-locked DRO is typically -110 dBc/Hz at 10-KHz offset, which is about a 15-dB improvement over a free-running DRO. Microphonics and other mechanically induced problems are greatly reduced by the phase-locked DRO, as the vibration-induced noise can be compensated when it falls within the bandwidth of the phase-locked loop. With the addition of a passive doubler or tripler device to the phase-locked DRO, block converter LO sources with excellent characteristics can be produced, up to a frequency of about 50 GHz. While there is a 10 to 16 dB loss caused by the multiplier, a biased mixer can be used which reduces the LO drive power requirement to about 0 dBm.

Gunn Oscillators

The Gunn oscillator is a comparatively efficient millimeter-wave oscillator that can operate at frequencies in excess of 100 GHz [2]. Its use in millimeter-wave block converters can overcome the limited output frequency and low output power of the DRO/multiplier LO scheme described above. Thermally compensated Gunn oscillators can rival the thermal stability of free-running DROs and can provide more than 10 mW, which can be used to directly drive a mixer without external bias. This has the advantage of increasing the dynamic range of the millimeter-wave receiver. As with DROs, it is common to improve the performance of the Gunn oscillator in order to use it in a block converter for millimeter-wave receiver applications by either phase or injection locking the Gunn oscillator.

The phase-locked Gunn oscillator uses a phase-locked DRO as a reference signal in the harmonic mixing portion of the circuitry. The output of the phase-locked DRO is applied to a harmonic mixer contained within a phase-locked loop controlling the Gunn oscillator, as shown in Figure 2. Essentially, the Gunn oscillator phase-locked loop portion of the circuitry shown in Figure 2 has replaced the multiplier in the DRO/multiplier approach described above. While this can be an expensive option, the increased frequency coverage and the higher output powers available from the Gunn oscillator make it very useful, particularly for the higher frequency applications.

The injection-locked Gunn oscillator is similar to the phase-locked Gunn oscillator, but is a less expensive option, as it requires less circuitry. Figure 3 shows a block converter that uses an injection-locked Gunn oscillator as the LO source. The schematic of the injection-lock circuitry is given in Figure 4, with the performance of the 42-GHz signal, as displayed on a spectrum analyzer, shown in Figure 5. The phase-locked DRO is used to drive a harmonic generator which has one of its harmonics at the required LO frequency. The Gunn oscillator is also mechanically tuned as close as possible to this required LO frequency. The low-level harmonic signal from the harmonic generator is applied to port 1 on the circulator. This signal is transferred to port 2 by the circulator and then enters the cavity of the Gunn oscillator. If the free-running frequency of the Gunn oscillator is within about 5 MHz of this harmonic signal, the Gunn oscillator frequency will be pulled to the injected harmonic frequency. The output of the Gunn oscillator is then transferred to port 3, emerging as a high power LO signal with the frequency characteristics and stability of the harmonic of the DRO and at a higher power level. This technique is very useful, but *does* suffer some problems not seen with the phase-

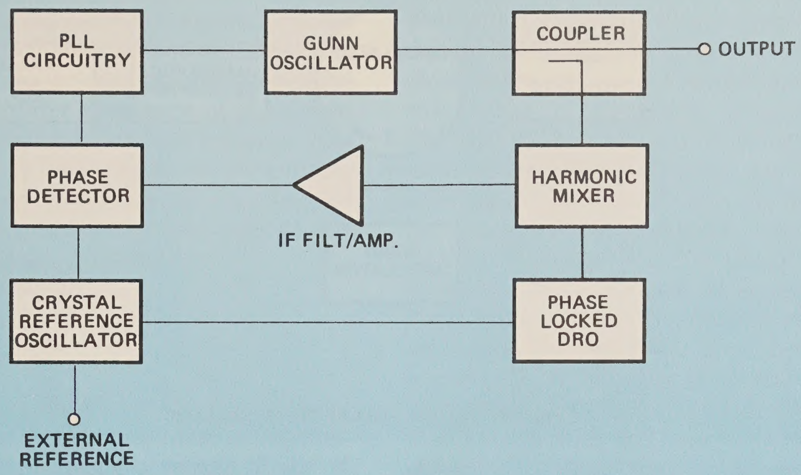

Figure 2. Millimeter-wave phase-locked Gunn oscillator.

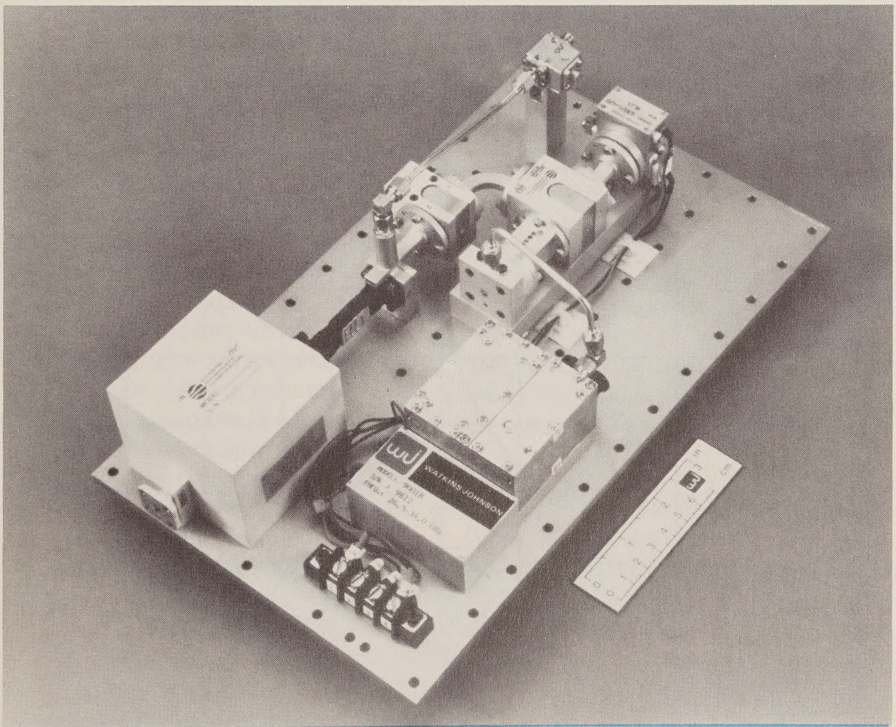

Figure 3. Block converter with injection-locked local oscillator.

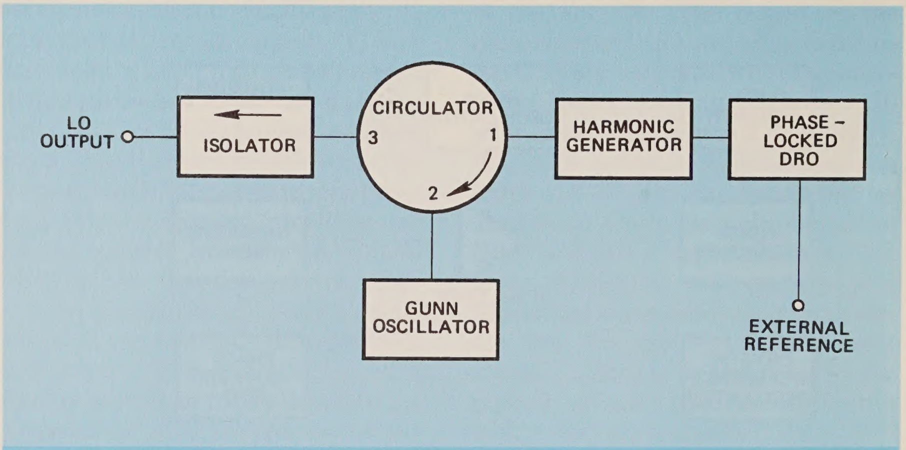

Figure 4. Injection-locked Gunn oscillator.

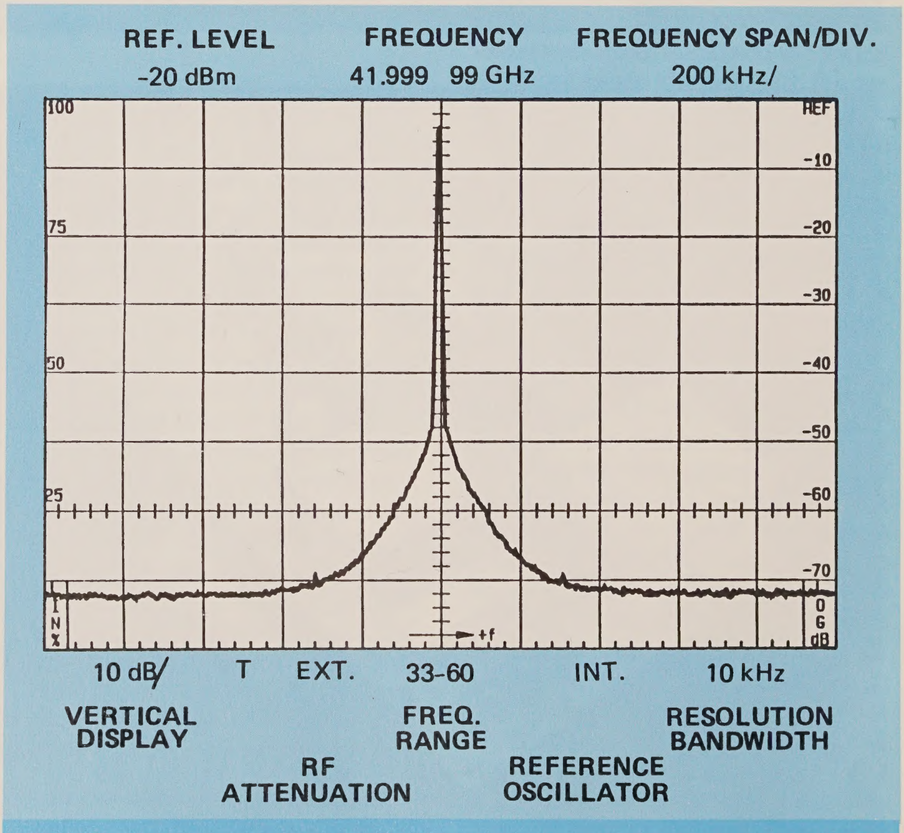

Figure 5. Typical injection-locked Gunn oscillator spectrum.

locked Gunn oscillator. The injection-lock range is limited to about ±5 MHz around the harmonic signal and the Gunn oscillator may not acquire lock, or could lose lock during operation if the free-running frequency of the Gunn oscillator drifts outside this range. This drift may be caused by thermal effects or aging, which can be corrected by thermal enclosures and periodic recalibration, respectively.

Mixers

The mixer that is used to convert the rf block of frequencies must be capable of instantaneous coverage of the rf bandwidth and have an IF bandwidth of about 14 GHz. The type of mixer chosen for the circuit in Figure 1 is a single-balanced waveguide biased mixer. While there are many styles of mixer that could be used to convert millimeter-wave signals, the most common ones used for broadband block conversion are the single-balanced waveguide and MIC styles.

The waveguide mixers used for the block converters shown in Figure 1 are of the traditional cross-bar construction [3], using two pill-packaged Schottky-barrier diodes. DC bias is applied to the diodes, which are connected in series across the E plane of the waveguide, in order to reduce the LO power requirements to the levels generated by the tripler shown in Figure 1. This reduces the compression point of the mixer by about 10 dB and results in a reduced dynamic range. At millimeter-wave frequencies, this loss in dynamic range is not usually a problem as the signal levels expected at these frequencies are relatively low. The cost advantages of using a biased mixer often outweigh the disadvantage of the reduction in dynamic range. Should maximum dynamic range be required, then one of the high power LO generation schemes described above can be used with a mixer that does not require dc bias. The mixer used for the Ka-band block converter shown in Figure 1 covers the rf frequency range of 26.5 GHz to 40 GHz using WR28 waveguide with a UG599/U flange, and the LO frequency is at 43.89 GHz, which is fed through WR22 waveguide using a UG383/U flange. The LO signal is applied to the diodes at their common junction, where the IF signal is also extracted. The two diodes appear connected in parallel to the IF signal as viewed from the IF/LO port connection. The IF signal is diplexed from the coaxial LO feed connected to the common junction of the diodes by a lowpass filter. The LO waveguide acts as a high-pass filter, since the cut-off frequency of the waveguide is above the maximum IF frequency. The mixer can be operated with either the rf or the LO signal applied to the common junction of the diodes, but there is a preference, based on spurious generation, as only the even harmonics of one of these signals is significantly suppressed due to the balance of the mixer. DC bias is applied to one of the mixer diodes, which are in series, and the bias current flows through both diodes. A dc block must be connected to the IF port to avoid grounding this bias current. A view of the construction of a typical mixer for use in a block converter is presented in Figure 6.

When space is of concern, a MIC mixer can be used instead of a waveguide type. While double-balanced MIC mixers covering the frequency range of 18 to 40 GHz (e.g., the WJ Spacekom SMC1840) do exist, it is difficult to simultaneously have an IF band that can extend up to 18 GHz. For this reason, the MIC mixers normally used for block converters are of single-balanced design using a diplexing

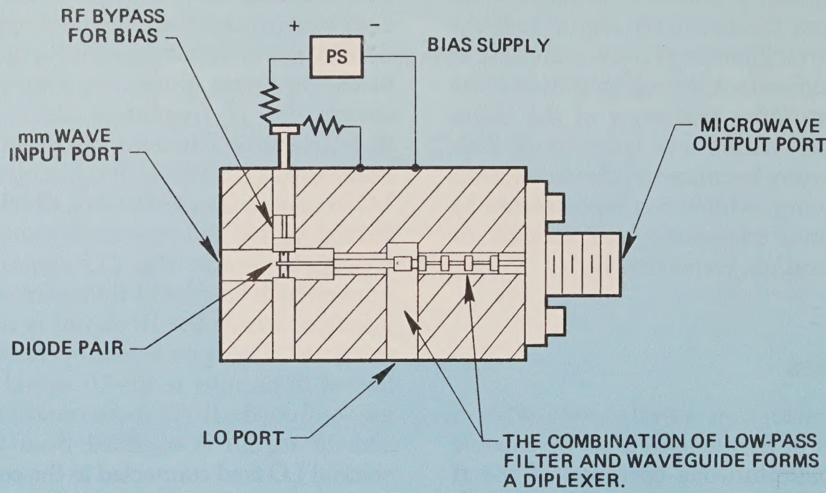

Figure 6. Cut-away view of a typical block-converter mixer.

technique to extract the IF signal in a similar manner to the waveguide mixer described above. These mixers are smaller than their waveguide counterparts and can be produced using thin-film techniques that give repeatable performance at a lower cost than the precision waveguide mixers. DC bias can be introduced into MIC mixers, but this becomes very difficult at millimeter-wave frequencies due to the very small dimensions of the mixer circuit. The parasitic elements of the bias circuitry can prevent the mixer from performing correctly. Therefore, MIC mixers for millimeter-wave applications are usually used with higher power LO sources.

Amplifiers

Amplifiers for use in block converters can be for IF, RF or LO amplification. The IF amplifiers have to be low-noise amplifiers and have broad bandwidth in order to cover the required range of IF frequencies up to 18 GHz. Balanced, distributed or matrix amplifiers have the necessary characteristics for the IF amplifier, depending upon the bandwidth required. At the rf frequencies, the most important consideration is noise figure. While advances in submicron GaAs FET amplifier technology have been very significant in recent years, broad-bandwidth amplifiers with a low enough noise figure to improve receiver sensitivity are not readily available. While some very impressive results have been reported for narrow bandwidth, the inclusion of an rf amplifier is not usually cost effective.

LO driver amplifiers can be useful if maximum receiver dynamic range is required. As discussed earlier, an improvement in dynamic range of about 10 dB can be obtained if the front-end biased mixer can be replaced with a non-biased kind. This requires a higher LO drive power, typically greater than 10 mW and, therefore, one of the higher output power LO schemes described above would have to be used, or an LO driver amplifier added. The small gate dimensions of FETs, with gate lengths less than a quarter of a micron and gate widths of the order of 100 microns, that operate in the millimeter-wave range of frequencies are very delicate devices and obtaining enough power to drive a standard mixer can push these devices to their performance limits. However, the narrow bandwidth required, due to the fixed frequency LO, does make this type of device feasible, as larger gate-width devices can be matched over the small frequency range required and sufficient output power extracted. Millimeter-wave amplification of any kind is an expensive option at the present time and is, therefore, not used in many applications because of cost constraints.

Spurious Signals

One of the most common problems with receiver electronics is the generation of unwanted (spurious) signals. Multiple tone intermodulation products are generated when multiple rf signals enter the receiver, and, in general, are described by [4]:

$$f_I = (\pm m_1 \cdot f_{R1} \pm m_2 \cdot f_{R2} \pm m_3 \cdot f_{R3} \ldots) \pm n \cdot f_L$$

Where,

$$m_1, m_2, m_3, \ldots = 1, 2, 3, \ldots$$

$$n = 1, 2, 3, \ldots$$

Two-tone, third-order intermodulation products are commonly called out in device specifications. These products are of the form $(\pm 2f_{R1} \pm f_{R2}) \pm f_L$ and $(\pm f_{R1} \pm 2f_{R2}) \pm f_L$, and are called third-order products because the sum of the coefficients of the two rf signals, f_{R1} and f_{R2}, equals 3. The order, N, is a useful parameter, as a change of 1 dB in the power level of both input rf signals will cause the absolute power level of the intermodulation product to change N dB. Hence, the relative suppression of the spurious signals compared to the IF signal power level will change by an amount equal to (N-1) dB for every 1 dB increase in the level of the two input signals. A measure of the device that is commonly used is the 3rd-order intercept point. This is defined as the rf power at which the levels of the required signal power and the undesired intermodulation products are equal. As the device usually is in compression before this power is reached, the intercept point, IP in dBm, can be determined only by extrapolation, using the following formula:

$$IP = \frac{R}{(N-1)} + S$$

Where,

R = Relative suppression of spur in dB
N = Order or intercept
S = Signal level in dBm

Using the above relationship, the intercept point can be determined by making measurements at convenient signal levels. The performance of the device at other signal levels can then be predicted.

Single-tone intermodulation products are a simplified version of the general multiple-tone intermodulation products equation shown above. The single-tone IF response, f_1, due to the combination of rf and LO signals, is given by:

$$f_I = \pm m \cdot f_R \pm n \cdot f_L$$

Where, $m = 1, 2, 3, \ldots$

$n = 1, 2, 3, \ldots$

This is the quantity usually used to describe the spurious performance of mixers. When m or n is equal to 0, then the resulting spurious signals will be the harmonics of the rf and LO frequencies, respectively. For the case of single-tone intermodulation, a common way to describe a spurious response is to give its level in dBc, relative to the required IF signal. The $m \cdot f_R$ product will increase relative to the required IF signal $|f_R \pm f_L|$ by (m–1) dB for every 1-dB increase in the f_R signal level. In the case where the LO is derived by multiplication, then sub-harmonics of the LO also need to be considered. For the case of a harmonically derived LO signal, the $\pm n \cdot f_L$ term should be replaced by $\pm n \cdot f_L/k$, where the integer k is the multiplication factor used to derive the LO signal. A bandpass filter in the LO line to the mixer is normally sufficient to reduce the multiplier harmonics to an acceptable level. Care has to be taken in specifying the rejection characteristics of the filter as higher harmonics can mix in the mixer to produce spurious signals within the IF bandwidth.

The mixer used in the Ka-band block converter described above is of single-balanced design. Double-balanced designs theoretically only generate one quarter of the possible intermodulation products that a single balance mixer would generate. Essentially, the difference between single-balanced and double-balanced mixers with regards to spurious generation is that the double-balance mixer rejects the even harmonics of the LO and rf signals while the single-balanced design rejects only one of these sets of harmonics. With the arrangement of LO described in the section on mixers above, the even harmonics of the LO signal are balanced in the single-balanced mixer discussed. By careful choice of the ports that the LO and rf signals use, then, the spurious performance of the single-balanced mixer can be made to be similar to that of the double-balanced design.

In general, spurious responses are difficult to predict and correct. The balancing of the circuit, matching of the diodes and variations in diode parameters, all affect the level of spurs. The termination of the harmonic components of the waveforms present in the mixer affects the level of these spurs and with these products extending into many tens of GHz, the termination that they see is unpredictable. The best way to determine the spurious levels to be found in a mixer is to measure a representative sample. Spurious levels can be optimized for a particular application by choosing the correct type of mixer and by applying the LO and rf signals to the correct ports.

Conclusion

Present day instruments and surveillance equipment can be upgraded to receive millimeter-wave signals. The approach presented here is one that uses the existing equipment to its fullest, keeping the cost of the upgrade below what it would be in the case of duplicating all of the capabilities that the present instruments posses. With careful attention to the design of the block converter, as well as to the converter-receiver interface requirements, it is possible to realize sophisticated designs at millimeter-wave frequencies.

Although the components used in millimeter-wave block converters are expensive, excellent performance can be achieved. Integrated MIC technology is useful in millimeter-wave block converters, but the development costs are very high and should only be expended should the size and performance criteria require it. The advent of cheaper millimeter-wave amplifiers will further aid in design improvements. A compromise should be achieved between cost and performance.

References

1. Kajfez, D. and P. Guillon, "Dielectric Resonators," Artech House, 1986, Chapter 10.
2. Greim, D. C., "W-Band Varactor Tuned Oscillators with 10-GHz Bandwidth," Microwave Journal, Cover Article, Vol. 31, No. 10, October 1988.
3. Maas, S., "Microwave Mixers," Artech House, 1986, pp. 259–261.
4. Henderson, Bert C., "MIXERS: Parts 1 and 2," Watkins-Johnson Tech-notes, Vol. 8, Nos. 2 & 3.

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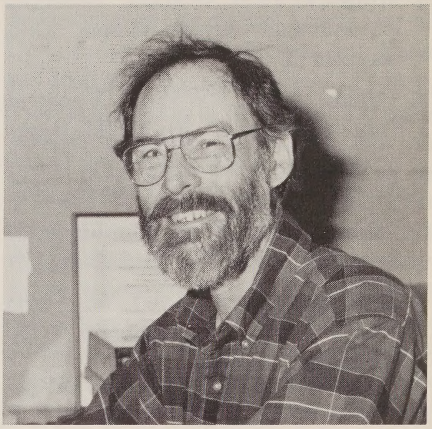
Dr. Hey-Shipton is Manager, Subsystems Product Engineering Department within the Subsystems Division of Watkins-Johnson Company, Palo Alto, California. He is responsible for the development of the complex subassemblies being built in Palo Alto (which cover the dc to 110 GHz frequency range) and the products from the Spacekom operation formerly located in Santa Barbara, California. At Spacekom, Dr. Hey-Shipton was responsible for the research, development, design, and management of devices and products at that facility for operation at millimeter-wave frequencies. These products include mixers and multipliers built using both waveguide and MIC techniques, Gunn diode oscillators, low-noise amplifiers, and the integration of these products. The operating frequency range for these devices and subassemblies is predominantly from 20 GHz to 110 GHz. He was also responsible for documentation, control, production testing, drafting, and program management areas of the operation. Dr. Hey-Shipton holds a B.Sc. from University of Manchester, England, and a Ph.D. from the University of Leeds, England.



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Mr. Denning is Senior Engineer, Spacekom Section Subsystems Division, Palo Alto, California. He has extensive design experience in the design of millimeter waveguide balanced mixers and is presently involved in the design of wideband downconverters for input frequencies that range from 18 GHz to 100 GHz. His most recent work has been on a subsystem to convert the 18 to 40 GHz band down to the 2 to 18 GHz band using phase-locked local oscillators and broadband low-noise amplifiers and mixers. Mr. Denning attended California Polytechnic University, Pomona, California.

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