Dual-Tone Calibration of Six-Port Junction and Its Application to the Six-Port Direct Digital Millimetric Receiver

Ji Li, Renato G. Bosisio, *Fellow, IEEE,* and Ke Wu, *Senior Member, IEEE*

***Abstract*— This paper describes a novel dual-tone calibration technique of six-port using two carriers with closely spaced frequencies. The samples for the calibration are extracted from outputs of four power detectors using a self-adaptive algorithm. The calibration procedure is fully automatic and can be implemented in a system capable of capturing the output waveforms of a six-port. Although the proposed method is applicable to general six-port calibrations, it finds itself particularly suitable for the calibration of a new six-port direct digital mm-wave receiver. It is shown that such a calibration of the digital receiver can be fulfilled on site simply by receiving usual incoming signals. Dual-tone calibrations made at 26.5 GHz, 33 GHz, and 40 GHz demonstrate the same order of accuracy as the conventional six-port calibrations. However, the new technique is much simpler and faster as it requires less effort in its implementation.**

I. Introduction

IT HAS BEEN RECOGNIZED that the calibration procedure proposed by Engen [1] and further developed by Hodgetts and Griffin [2] is the most robust method in the calibration of a six-port circuit. In this procedure, a six- to four-port reduction is made first to acquire five real calibration coefficients, and then a so-called "error box" calibration is carried out to resolve the three remaining complex calibration coefficients.

When applied to the calibration of a six-port reflectometer, the six- to four-port reduction procedure can be considered as to determine parameters of a directional coupler by locating the origins of three circles in a complex plane (the so-called W-plane). The common interception point of these three circles represents the virtual reading of a reflection coefficient measured by the equivalent four-port. Fortunately, this W-plane calibration can be accomplished simply by measuring a number of arbitrary but different terminations. The exact knowledge of the reflection coefficients of these terminations is not essential. Once the W-plane calibration is completed, an "error box" calibration, similar to the standard procedure in a conventional vector network analyzer, is performed. The error box calibration eliminates residual errors arising from directivity, source match, and frequency response of the equivalent four-port, and it fixes the position of reference plane as well.

Various methods have been used to generate the at least nine different terminations required for a W-plane calibration. The most straightforward way is to use a sliding short [1]. This method is very easy and reliable but difficult to automate. To automate the calibration, an "active load" method has been proposed in which a vector microwave modulator is used to feed a portion of the reference signal back into the DUT port. In this way, a variable load can be simulated [3]. An automatic termination generator (ATG) has also been proposed and applied to a commercial six-port reflectometer [4]. Such techniques require additional microwave hardware and auxiliary control circuits specifically made for such calibration purposes.

Recently, a direct mm-wave receiver (DMR) was proposed in which a six-port phase/amplitude/frequency discriminator is used [5], [6]. The DMR is capable of directly demodulating various digital modulated signals such as PSK and QAM. The new receiver provides an interesting alternative in various communication systems such as point-to-point communication systems and satellite ground terminals. However, it would be entirely impractical for a DMR to perform a calibration procedure if it were required to connect external physical standards to its input port. It is therefore necessary to develop a calibration method free from any external connection.

To accomplish this goal, a dual-tone six-port calibration method is proposed in this paper. The new dual-tone calibration is achieved by feeding a RF signal other than the reference source into the DUT port of the six-port. This signal can be either a unmodulated single carrier with a frequency adjacent to the six-port local source frequency, or a digital modulated signal with a frequency equal to or close to the local source frequency. The resulting output waveforms of the diode detectors of the six-port are actually beat-signals of the two RF signals (i.e., the six-port reference source and the second external RF signal). The voltage readings corresponding to a group of widely distributed terminations are acquired by properly sampling these waveforms. It is well known that the actual values of these terminations are not required for the W-plane calibration. Moreover, in the case where the absolute phase reference plane is not important as in a six-port DMR, the same waveform capture algorithm can also be applied to the error box calibration without the need to use precise standards such as short, open and match loads as for a six-port reflectometer. Obviously, a real time data acquisition system

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The authors are with the Poly-Grames Research Center, École Polytechnique, C. P. 6079, Succ. Center Ville, Montréal, Canada H3C 3A7.

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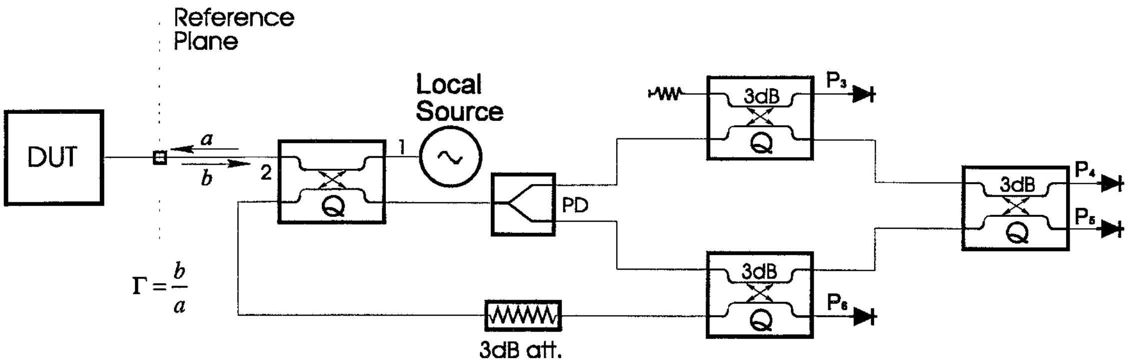


Fig. 1. A typical six-port reflectometer.

capable of capturing instant voltage waveforms, such as real-time A/D converter or multi-channel digitizing oscilloscope, is essential to accomplish the proposed dual-tone calibration.

The dual-tone calibration is fully automatic in its nature: no repetitive connection of standards or tuning of sliding short is needed. The frequency difference between the two RF signals can be varied over a wide range as long as it remains below the limit of frequency response of the diode detectors and the data acquisition system. Along with a dual-tone linearization of the diode detectors [7], [8], the whole internal six-port calibration can be completed in a single step. The standard error box calibration, as used in a conventional vector network analyzer, is performed for instrumentation purposes only. The requirement of an additional RF signal generator for the dual-tone method could be somewhat difficult in some cases. On the other hand, this is definitely ideal for the six-port DMR calibration where the received signal acts as the second tone, thereby eliminates additional physical connections.

II. Dual-Tone W-Plane Calibration of the Six-Port

A. W-Plane Calibration

The so-called W-plane calibration involves the solution of the following [2]

$$\begin{aligned} &pQ_1^2 + qA^2Q_2^2 + rB^2Q_3^2 + (r-p-q)A^2Q_1Q_2 \\ &\quad + (q-p-r)B^2Q_2Q_3 + (p-q-r)A^2B^2Q_2Q_3 \\ &\quad + p(p-q-r)Q_1 + q(q-p-r)A^2Q_2 \\ &\quad + r(r-p-q)B^2Q_3 + pqr = 0 \end{aligned} \tag{1}$$

for p, q, r, and A^2, B^2, where $Q_i = P_{i+3}/P_3$. These coefficients allow transformation of a six-port to a perfect four-port reflectometer in a notional "W" complex plane. The W-plane reflection coefficient is

$$\begin{aligned} W = &\frac{Q_1 - A^2Q_2 + r}{2\sqrt{r}} + j\,[r(p+q-r) + (p-q+r) \\ &\cdot Q_1 - (p-q-r)A^2Q_2 - 2rB^2Q_3]/ \\ &[\pm 2\sqrt{r(2pq + 2qr + 2pr - p^2 - q^2 - r^2)}]. \end{aligned} \tag{2}$$

Equation (1) may be solved by measuring at least nine (usually 13 in practice) arbitrary different terminations [1]. The dual-tone method is a novel technique that generates these terminations without resorting to actual physical connections.

B. Using a Second Carrier

Fig. 1 shows a typical configuration of six-port reflectometer. The conventional four- to six-port reduction procedure requires at least nine connections of different loads to the DUT port. Instead of connecting a real load to the DUT port, if a RF signal ("reflected wave" b) coupled from the reference source is fed back into the DUT port with the same amplitude and phase as the "incident wave" a, the equivalent reflection coefficient ($\Gamma = b/a$ will be unity as if the DUT port is open-circuited. It is theoretically possible to generate any equivalent termination by varying the amplitude/phase of the signal b. Such active load synthesis technique has been effectively used in the modeling of microwave power transistors [9], [10]. Let us consider the case where the "incident wave" a and the "reflected wave" b are different in frequency, as in the following:

$$a = |a| \cdot e^{j(2\pi f_1 t + \varphi_1)} \tag{3}$$

$$b = |b| \cdot e^{j(2\pi f_2 t + \varphi_2)}. \tag{4}$$

Suppose the frequency difference $f_2 - f_1$ is very small so that the S parameters of the six-port to be calibrated can be regarded as being the same at f_1 and f_2, the equivalent reflection coefficient then becomes

$$\Gamma = \frac{b}{a} = \left|\frac{b}{a}\right| \cdot e^{j(2\pi\Delta f t + \Delta\varphi)} \tag{5}$$

where $\Delta f = f_2 - f_1$, and $\Delta\varphi = \varphi_2 - \varphi_1$.

Clearly, the equivalent reflection coefficient Γ becomes a time-dependent vector whose amplitude is invariant and the phase is shifting around at a constant angular speed $2\pi\Delta f$. In the case of $|a| = |b|$, it is as if a perfect sliding short moving at a constant speed is connected.

The voltage at a detector port of the six-port is a vector summation of portions of a and b waves presented at the port. Let the "reflected wave" b be a single CW signal and the diode detectors fall in perfect square-law, the output voltage waveforms of the diode detectors are

$$V_{out.i} = \tfrac{1}{2}\left(|a_i|^2 + |b_i|^2\right) + |a_i| \cdot |b_i| \cos\left(2\pi\Delta f t + \Delta\varphi_i\right) \tag{6}$$

where ai, bi are waves corresponding to a, b at port a, respectively. Fig. 2 plots output voltage waveforms of a typical six-port with a CW "reflected wave" b.

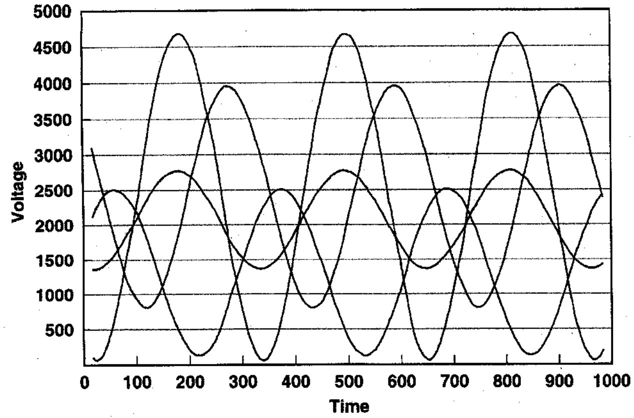


Fig. 2. Typical output waveforms of a six-port with a CW input.

It is noticed that a period of the waveform corresponds to a whole circle of Γ rotating in the complex plane. Therefore we capture waveforms of the four channels synchronously and choose the channel with maximum voltage swing as the reference channel. Selecting samples at an equal amplitude space in the whole voltage swing in this channel, these samples will present a group of equally spaced terminations on a circle in the Γ plane. By varying the amplitude of either the input CW signal or the local source, samples of Γ's well distributed over the whole Smith Chart can be obtained.

C. Dual-Tone Sampling Algorithm

The flow chart outlining the algorithm for selecting the samples is shown in Fig. 3. There is no particular restriction to frequency difference of the two signals. However, the maximum frequency of a capturable waveform depends on the sampling rate of the data acquisition systems and track/hold (T/H) circuit response time. It should be emphasized that the samples must be taken synchronously, i.e., the four power readings for each sample must be taken at the same instant. It is also possible to use T/H circuits to freeze the waveforms and then perform A/D conversion sequentially. The rule of thumb is that at least 20–30 samples be taken in a cycle when a CW signal is used. The power level of the "reflected wave" b has to be strong enough to suppress noise interference. The power level step should be around 3 dB but need not be precise.

The algorithm is designed to deal with a wide variety of signals such as unmodulated CW, PSK, QAM, and AM signals, etc. The signals involving wide band frequency modulation may be difficult to capture when the bandwidth of data acquisition system is narrow. Fig. 4 illustrates examples of sampling using the conventional 13-standard method and the proposed dual-tone method with different types of input signals. It can be seen that the samples are well distributed over the entire Smith chart regardless of the signal type. The slight shift of the origin of the dual-tone samples is due to the residual reflection at the DUT port.

It can be logically anticipated that for unmodulated CW signals, there should be a small frequency difference between the two carriers, whereas this is not crucial for a high-level digital modulated signal (phase states no less than four). On the other hand, a carrier-synchronized QAM (e.g., 16QAM, 64QAM) signal provides a perfect sample distribution.

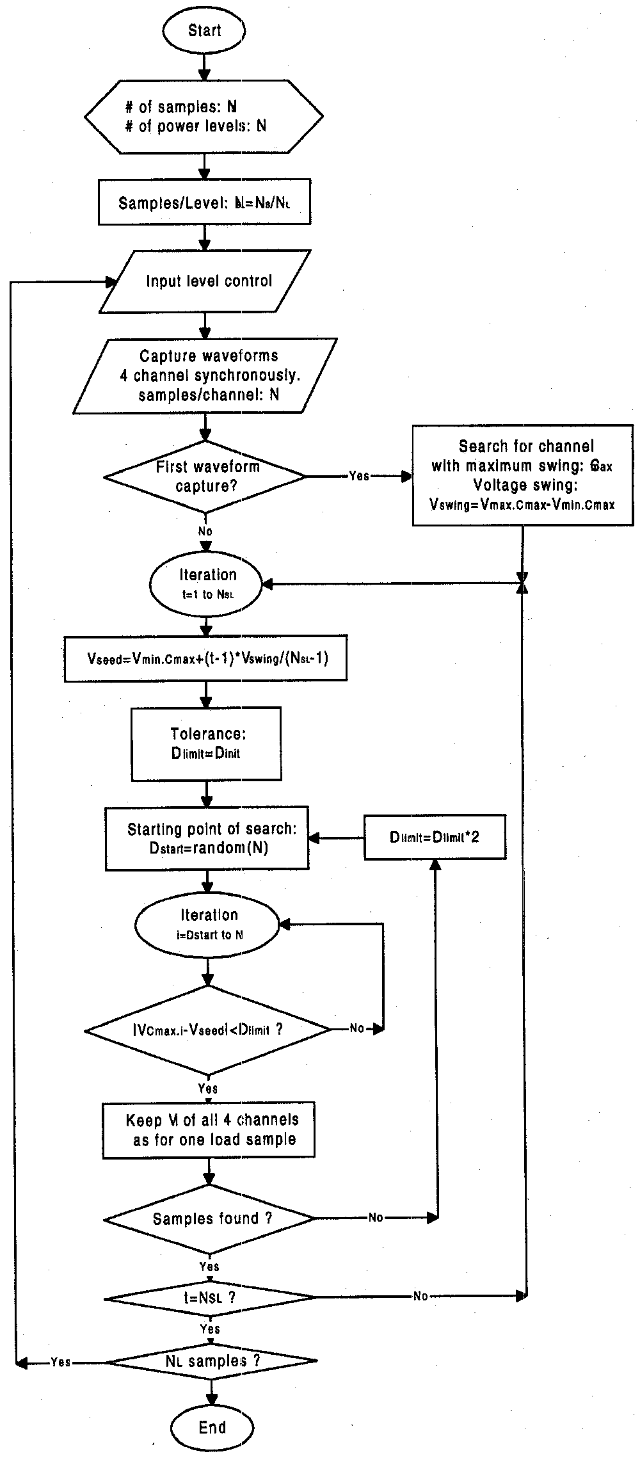


Fig. 3. Flow chart of the algorithm taking samples of W-plane calibration.

D. Waveform Smoothing

The two carriers used in the dual-tone calibration are generally incoherent signals, i.e., not being generated from the same oscillator. Consequently, the phase noise of the two carriers will introduce random instant phase jitters in the

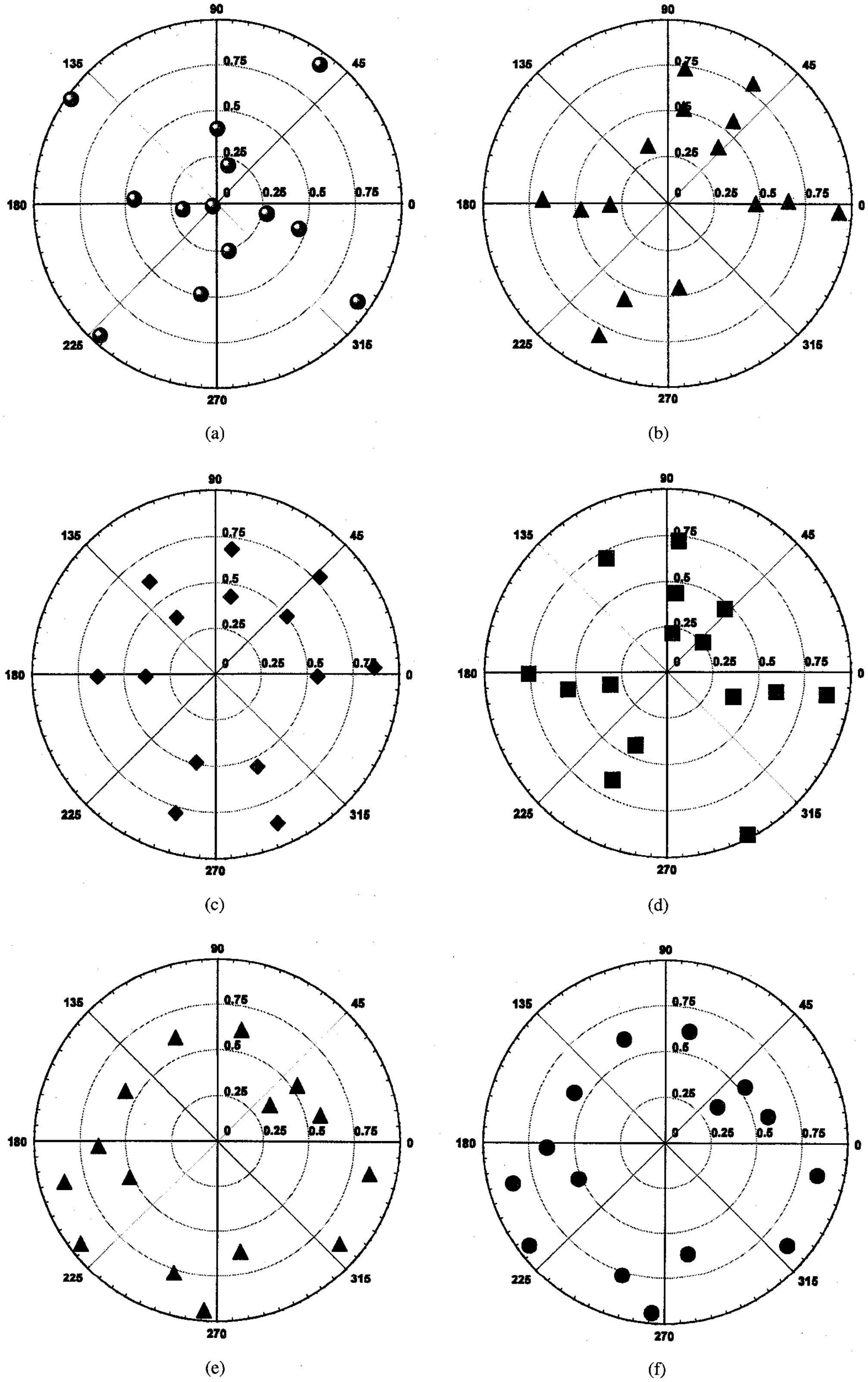


Fig. 4. Sample distributions using 13-standard method and dual-tone method. (a) Conventional 13–std method; (b) CW input 3 levels; (c) CW input 2 levels; (d) QPSK frequency unlocked 3 level; (e) 16QAM frequency unlocked 1 level; (f) 64QAM frequency unlocked 1 level.

equivalent Γ. When a CW "*b*" wave is used for the calibration, the output waveforms of detectors become noisy sinusoidal waves. It is preferable sometimes to smoothen the captured waveforms before sampling, although this usually does not affect the calibration result provided that the response time of the data acquisition circuit be fast enough to follow the jitters. A simplified Savitzky–Golay method called "moving window average" [11] is found to be very effective in handling such

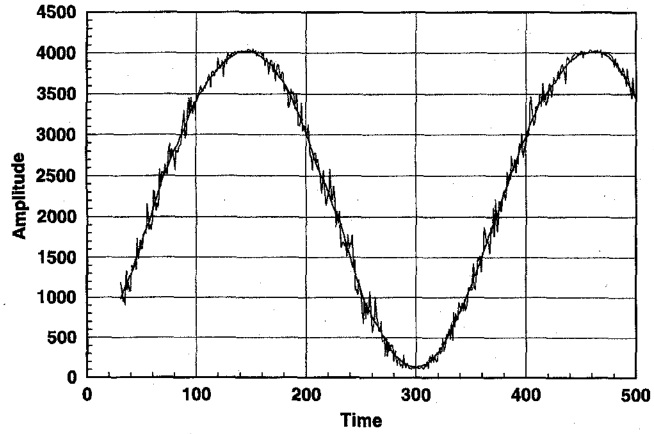


Fig. 5. Waveforms before and after smoothing.

problems. The characteristic equation of this method is

$$v_{smooth.i} = \sum_{n=-n_L}^{n_R} c_x v_{orig.i+n} \tag{7}$$

where $v_{smooth.i}$ is the voltage after smoothing, v_{orig} is the original data and $c_x = 1/(n_L + n_R + 1)$, n_L and n_R are widths of the windows on the left and right sides, respectively. Fig. 5 illustrates a real captured waveform before and after smoothing. It clearly indicates the effectiveness of the method.

The dual-tone method captures real time waveforms at the six-port power detector ports. This makes it more vulnerable to noise than a typical reflectometer calibration using standard power meters or multi-sample/averaging A/D converters. An improved algorithm proposed by Potter and Hjipiers [12] could be helpful to obtain more stable solution for (1).

III. Dual-Tone Error-Box Calibration of the Six-Port DMR

A. Error Box Calibration

A two-port error box is inserted between the notional four-port and the DUT such that the virtual reflection readings from the notional perfect reflectometer obtained by W-plane calibration are transferred to the real reflection coefficients Γ. This is done through a bilinear transformation

$$\Gamma = \frac{e - W}{cW - d} \tag{8}$$

where c, d, and e are complex constants related to the S parameters of the error box

$$\begin{aligned} c &= -s_{22} \\ d &= s_{12}s_{21} - s_{11}s_{22} \\ e &= s_{11}. \end{aligned} \tag{9}$$

A standard calibration using three known standards (e.g., short, open, and matched loads) for a conventional vector reflectometer can be applied here. In addition, a fourth load that needs only to be roughly known has to be used to determine the sign in (2) [2].

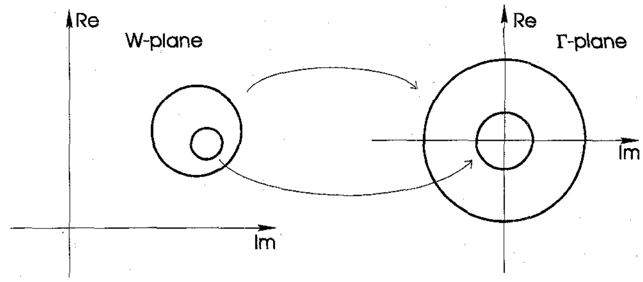


Fig. 6. W-plane to Γ-plane bilinear transformation: the circles are mapped into circles but not concentric.

B. Necessity of Error Box Calibration for the Six-Port DMR

It is in fact impossible to connect standards to a six-port DMR in the presence of preceding low noise amplifier and other circuits. This also inhibits the application of a typical error box calibration to a DMR. On the other hand, the absolute phase reference plane is not important for digital demodulation, therefore a simplified error box calibration can be adopted.

It would be nice to bypass the error box calibration regarding the fact that the bilinear transformation (8) maps a circle in W-plane into a circle in Γ-plane. Unfortunately, it is found in most cases that the equal-reflection circles in W-plane are not concentric as shown in Fig. 6. This results in distortion of the constellation diagram of digital modulation and it is therefore unavoidable to use transformation (9) for the six-port DMR.

C. Dual-Tone Error Box Calibration for DMR

In a six-port DMR, the phase reference plane is not important. This allows us to adopt a procedure that is similar to the dual-tone W-plane calibration. Rewrite (6) as

$$V_{out.i} = \tfrac{1}{2}\left(|a_i|^2 + |b_i|^2\right) + |a_i| \cdot |b_i| \cos\left(\varphi_{a_i} - \varphi_{b_i}\right). \tag{10}$$

The $V_{out.i}$ reaches the maximum and minimum values when

$$\begin{aligned} &\text{Max.:} \quad \varphi_{b_i} = \varphi_{a_i} - 2n\pi \\ &\text{Min.:} \quad \varphi_{b_i} = \varphi_{a_i} - 2n\pi - \pi \\ &\qquad\qquad n = \cdots -2, -1, 0, 1, 2 \cdots. \end{aligned} \tag{11}$$

The swing of the waveform is given by

$$\Delta V = V_{out.i.\max} - V_{out.i.\min} = 2|a_i| \cdot |b_i|. \tag{12}$$

Let $b' = kb$, thus $b'_i = kb_i$, and the swing becomes

$$\Delta V' = V'_{out.i.\max} - V'_{out.i.\min} = 2k\,|a_i| \cdot |b_i|. \tag{13}$$

Then

$$\frac{\Delta V'}{\Delta V} = k. \tag{14}$$

From the definition of $\Gamma = b/a$, then

$$\Gamma' = \frac{b'}{a} = \frac{kb}{a} = k\Gamma. \tag{15}$$

Obviously, the swing of waveform represents magnitude of the equivalent Γ term, and if a Γ corresponding to the maximum voltage at one port (the reference port) is taken as

TABLE I
CALIBRATION COEFFICIENTS USING CONVENTIONAL 13-STD METHOD AND DUAL-TONE METHOD

Freq.	26.5 GHz			33 GHz			40 GHz		
	13-std	cw 3-level	cw 2-level	13-std	cw 3-level	cw 2-level	13-std	cw 3-level	cw 2-level
p	2.09850	2.05330	2.00897	2.64877	2.66865	2.66332	1.54830	1.51983	1.48923
q	1.45223	1.45686	1.42006	1.71614	1.75197	1.74824	0.84404	0.81323	0.80494
r	1.40958	1.41367	1.35935	1.52518	1.50819	1.51315	1.09592	1.09138	1.07826
A^2	0.58268	0.58215	0.53294	0.29511	0.29469	0.29379	0.35069	0.35050	0.34374
B^2	1.08903	1.12284	1.07837	1.18485	1.19764	1.19214	0.71840	0.69764	0.68318

TABLE II
MEASURED REFLECTION COEFFICIENTS USING HP8510 VECTOR NETWORK ANALYZER AND SIX-PORT REFLECTOMETER CALIBRATED BY THE USE OF CONVENTIONAL 13-STD METHOD AND DUAL-TONE METHOD

	26.5 GHz				33 GHz				40 GHz			
	HP8510	13-std	cw 3-level	cw 2-level	HP8510	13-std	cw 3-level	cw 2-level	HP8510	13-std	cw 3-level	cw 2-level
#1	0.973∠ -177.6°	1.043∠ -179.5°	1.010∠ 180.0°	0.968∠ -179.1°	0.997∠ -178.0°	0.980∠ 179.5°	0.978∠ 178.7°	0.970∠ 179.3°	0.985∠ 178.5°	0.990∠ 178.2°	0.996∠ 176.8°	0.998∠ 175.7°
#2	0.999∠ 91.9°	0.996∠ 88.5°	0.998∠ 88.3°	0.998∠ 88.2°	0.969∠ -103.0°	0.979∠ -107.6°	0.980∠ -107.3°	0.980∠ -107.6°	0.975∠ 36.7°	0.969∠ 36.9°	0.972∠ 36.8°	0.972∠ 36.9°
#3	0.962∠ -78.3°	0.975∠ -77.4°	0.973∠ -77.3°	0.973∠ -77.3°	0.985∠ 80.4°	0.982∠ 79.8°	0.983∠ 80.2°	0.982∠ 79.8°	0.972∠ -133.8°	0 968∠ -134.2°	0.971∠ -133.8°	0.974∠ -133.7°
#4	0.049∠ -121.5°	0.032∠ -148°	0.029∠ -156°	0.026∠ -153°	0.038∠ -77.0°	0.019∠ -46°	0.020∠ -52°	0.019∠ -51°	0.025∠ -93.2°	0.025∠ -109°	0.026∠ -95°	0.026∠ -95°
#5	0.422∠ 104.0°	0.425∠ 102.5°	0.427∠ 101.7°	0.428∠ 101.5°	0.523∠ 31.8°	0.554∠ 32.6°	0.546∠ 32.1°	0.549∠ 32.3°	0.413∠ -66.9°	0.416∠ -69.8°	0.418∠ -69.1°	0.420∠ -69.2°
#6	0.218∠ 89.3°	0.212∠ 89.5°	0.215∠ 89.5°	0.218∠ 88.6°	0.277∠ 54.1°	0.289∠ 53.3°	0.287∠ 53.1°	0.289∠ 52.8°	0 255∠ -75.6°	0.266∠ -77.3°	0.268∠ -76.9°	0.270∠ -76.1°
#7	0.066∠ -38.9°	0.091∠ -39.7°	0.093∠ -37.0°	0.092∠ -35.9°	0.061∠ -175.0°	0.048∠ 178.0°	0.049 -178.0°	0.049∠ -178.0°	0.100∠ -16.5°	0.113∠ -23.3°	0.116∠ -23.0°	0.114∠ -23.0°
#8	1.016∠ 95.7°	1.000∠ 90.7°	1.001∠ 90.7°	0.999∠ 90.3°	0.982∠ -99.8°	0.983∠ -103.9°	0.985∠ -103.6°	0.985∠ -103.5°	0.988∠ 39.2°	0.972∠ 39.7°	0.976∠ 38.8°	0.980∠ 39.4°

the phase reference such that $\Gamma = |\Gamma| e^{j0}$, then the Γ at the minimum voltage at this port should be $\Gamma' = |\Gamma'| e^{j\pi}$.

Accordingly, we define

$$\begin{aligned} \Gamma_1 &= 1 \cdot e^{j0} && \text{at} \quad V_{out.ref.\max} \\ \Gamma_2 &= 1 \cdot e^{j\pi} && \text{at} \quad V_{out.ref.\min} \\ \Gamma_3 &= k \cdot e^{j0} && \text{at} \quad V'_{out.ref.\max} \\ \Gamma_4 &= k \cdot e^{j\pi} && \text{at} \quad V'_{out.ref.\min} \end{aligned} \tag{16}$$

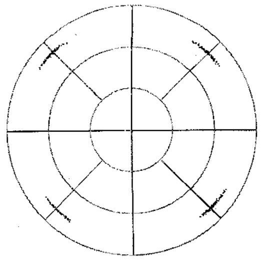
and use these values to solve unknowns c, d, and e in (8) such that the error box calibration of the six-port DMR is accomplished. The above approach cannot determine the sign in (2). Nevertheless, it is found that a wrong sign in (2) leads to a reversed phase mapping in the constellation diagram, which can be readily detected and corrected in the demodulation algorithm.

IV. RESULTS

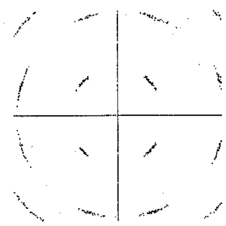
The proposed dual-tone calibration is performed on a six-port circuit at 26.5 GHz, 33 GHz, and 40 GHz. A PC486 is used as a controller of the measurement system. A data acquisition card with track/hold amplifiers captures waveforms in real time. CW signals with step power control are used for the dual-tone calibration. The conventional 13-standard calibration is also made using the same system.

Table I lists the W-plane calibration coefficients p, q, r, A^2, and B^2 obtained from both conventional 13-standard (13-std) method and dual-tone method at three different frequencies. It can be seen that the dual-tone method yields almost the same values of calibration coefficients as the conventional 13-std method. The insignificant difference between the two methods is basically of the same order as the fluctuation of values in two different 13-std calibrations. It is also found that a two-level power control is sufficient for a good calibration. An attempt to eliminate the power control was not successful.

Subsequently, standard error box calibrations are carried out using the W-plane calibration coefficients given in Table I. Reflection coefficients of various loads are measured by the use of the calibrated reflectometer. The same loads are also measured with a HP8510 vector network analyzer for reference purpose. Table II presents the measured Γ of those loads. It can be seen that the overall maximum error is 0.6 dB/5° except for very low reflection loads nos. 4 and 7. It is noticed that a system error appears between six-port and network analyzer measurements which might be caused by different calibration standard sets used for HP8510 and six-port. In most cases, the difference between the 13-std and the dual-tone calibrations is in the order of 0.1 dB/1.0°. Note that the readings are taken from a real time continuous display of reflection coefficient, and only one sample is taken for each reading without averaging. Due to the presence of noise,



(a)



(b)

Fig. 7. Constellation diagrams measured by a six-port digital receiver calibrated using dual-tone W-plane and error box method. The angular spread is represents the phase/frequency jitters of the oscillators used in the experiment. (a) QPSK. (b) 16QAM.

the displayed reading itself has a variation of 0.1 dB/1.0° for most situations. Therefore, the accuracy of the new dual-tone calibration is actually of the same order as that of a 13-std calibration.

A dual-tone error box calibration is also accomplished. Fig. 7 shows constellation diagrams of QPSK and 16QAM modulations measured by a six-port receiver calibrated by the dual-tone method. The phase span of the vectors is caused by phase noise of the signal generators and slow acquisition time. No obvious distortion of the constellation is observed.

V. CONCLUSION

A novel dual-tone W-plane calibration for six-port is proposed and shown to be able to achieve, at a high speed, the same order of accuracy as the conventional 13-standard calibration as applied to six-port reflectometers. With the proposed dual-tone method, the calibration of a six-port direct millimetric digital receiver including diode linearization, W-plane calibration and error box calibration can be implemented as desired in an operating receiver.

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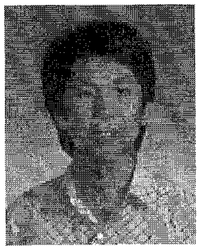
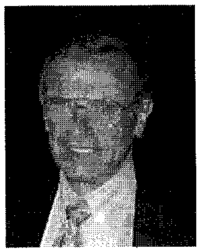


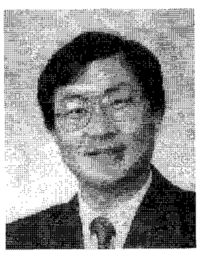
Ji Li received the B.S. from Xi'An Jiaotong University, Xi'An, China, and the M.S. from the Academy of Science of Posts & Telecommunications, MPT, China. He is approaching the conclusion of his Ph.D. study at the École Polytechnique de Montréal, Canada.

Telecommunication systems and microwave circuit design are among his research interests.



Renato G. Bosisio (M'79–SM–F'95) received the B.Sc. degree from McGill University, Montreal, Canada, and the M.S.E.E. degree from the University of Florida, Gainesville.

He has been engaged in microwave research and development work with various firms: Marconi and Varian in Canada, Sperry in the USA, and English Electric in England. He is Professor Emeritus and Head of Poly-GRAMES, a university microwave research center at École Polytechnique de Montréal, Montreal, Canada, where he teaches microwave theory and techniques. He is actively engaged in six-port technology, dielectric measurements, and computer-aided testing and design of active and passive microwave devices in M(H)MIC.



Ke Wu (M'87–SM'92) was born in Jiangsu, China. He received the B.Sc. degree (with distinction) in radio engineering from Nanjing Institute of Technology (now Southeast University), Nanjing, China, and the D.E.A. degree in electronics and Ph.D. degree (with distinction) in optics, optoelectronics, and microwave engineering from Institut National Polytechnique de Grenoble (INPG), France.

During the years 1983 to 1987, he conducted research in the Laboratoire d'Eléctromagnetisme, Microondes et Optoélectronics (LEMO), Grenoble, France. From 1988 to 1992, he was a Research Associate at the University of Victoria, Victoria, BC, Canada. In 1992, he joined the Départment de génie électrique et de génie informatique at the École Polytechnique de Montréal where he is now an Associate Professor. His main research interests include analysis and design of various microwave/millimeter-wave integrated circuits and antennas, high-speed interconnects and packaging effects, numerical methods, dielectric material characterizations, and superconducting devices. He is also interested in research and design of broadband optoelectronic components and lightwave transmission systems with emphasis on traveling-wave electro-optic modulators, couplers, and switches.

Dr. Wu received a U.S.R.I. Young Scientist Award in 1987, and together with two coauthors, the Oliver Lodge Premium from the IEE in 1988. He has published extensively and serves as editorial or review boards of various technical journals.