Direct Polar Display of Subnanosecond Millimeter-Wave Switching at 300 Mbit/s

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Abstract—**The switching behavior of an 80.99-GHz p-i-n diode modulator for the 40–110-GHz waveguide transmission system is studied. Detailed knowledge of the type of transient is required to establish rise-time and timing-error specifications compatible with the system performance objective. A test set displaying the transients at 300 Mbit/s in polar coordinates with a convenient time readout is described.**

The technique is suitable for the study of multiphase modulators in ultra-high-speed digital radio or satellite transmission systems.

I. Modulator Concept

P-I-N DIODE modulators have been built for 12 selected frequencies between 40 and 110 GHz, for use in a field evaluation test of the waveguide transmission system [1]. The lowest frequency was 40.235 GHz and the highest was 108.715 GHz. A modulator at 80.99 GHz is selected to demonstrate the general switching performance. While the system baud rate is 274 Mbit/s, all measurements reported were made using a 300 Mbit/s pseudorandom word.

The modulator consists of a p-i-n diode path-length switch [2] and two circulators [3] integrated into one housing. The first circulator is terminated and provides high input isolation. The second circulator guides both the incident CW signal to the p-i-n diode switch and the reflected phase-modulated signal to the output. A high-speed hybrid integrated driver is also inserted into the same housing, resulting in a compact modulator (see Fig. 1).

In response to the input data, the driver applies either a forward bias or a reverse bias to the p-i-n diode. The choice of the forward-bias current is a compromise between low loss and high switching speed. A larger forward-bias current means a lower p-i-n diode resistance [4] and therefore a lower forward-state loss. On the other hand, a larger forward-bias current results also in a higher stored charge and therefore a slower switching speed [5]. Good results were obtained with 5-mA forward bias. The reverse-bias voltage determines the power-handling capability since the peak RF voltage permitted at the diode is about equal to the reverse bias [2]. With 10-V reverse bias, a power-handling capability of about 250 mW has been observed. Ideally, the modulator exhibits a 180° differential phase shift and no amplitude change between the two bias states. The phase error and amplitude unbalance in a 500-MHz band are adjusted to be within $\pm 5°$ and ± 0.25 dB, respectively. The total modulator loss at 80.99 GHz was 2.5 dB, half of which was due to the two circulators and half to the p-i-n diode switch.

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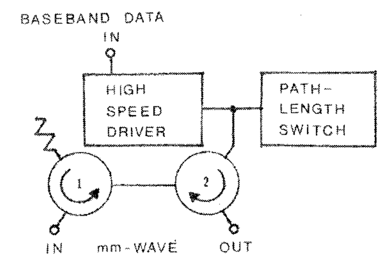
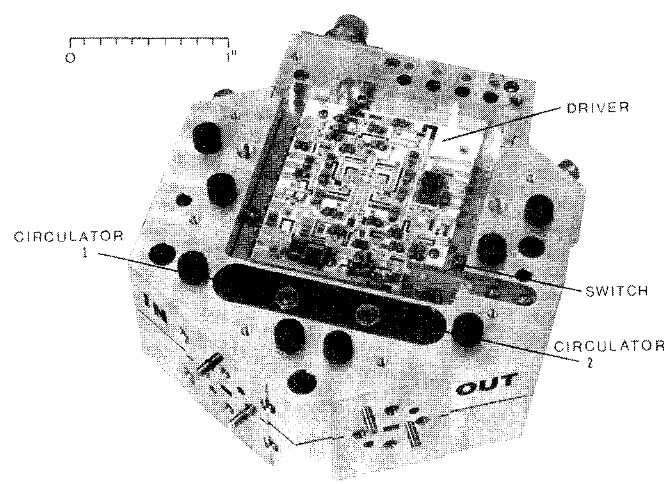


(a)



(b)

Fig. 1. Integrated WT4 modulator. (a) Block diagram. (b) Assembled unit.

II. Transient Type and Timing Requirements

The point of interest here is the nature of the transient from one bias state to the other. It is best visualized in the complex plane of the transmission coefficient S_{21}. There the two bias states are represented by two points of equal magnitude and 180° difference in phase. For pure reactive switching, the transient lies on the circle going through the two points with the origin as the center. For pure resistive switching the transient is the straight line connecting the two points and going through the origin. No power is lost in pure reactive switching. However, the power is reduced during the resistive switching, since the power goes

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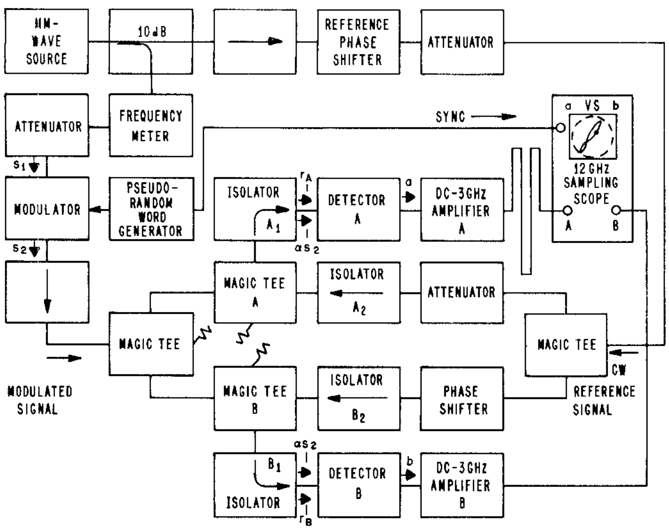

Fig 2 Block diagram of fast polar-display test set.

to zero when the origin is passed. Resistive switching is preferred in the system, because it leads to less stringent and therefore easier realizable timing requirements than reactive switching [6]. In practice, one finds a mixed reactive–resistive trajectory. In all cases studied by us the transient was more resistive than reactive.

To satisfy the system performance objective [1], the following requests on the rise time T_R and the timing error ε_T have been established through extensive computer simulation studies (under the assumption of a close to resistive switching transient)

$$T_R < T/4 \tag{1a}$$

$$|\varepsilon_T| < T/16 \tag{1b}$$

where T is the baud period with 3.6 ns.

III. Principle of Direct Transient Display

In our earlier work, we used a bridging method to mix the unmodulated with the modulated carrier at a square-law detector. By making four measurements for which the relative phase of the two carriers was changed in steps of 90°, we were able to construct the switching transient point by point. The technique was further enhanced with the aid of a minicomputer-controlled sampling scope for time averaging. The reduction in noise makes this approach particularly suitable for modulator studies under high noise conditions. A compensation method using only the modulated signal and an adjustable calibrated impedance [7] also permits a point-by-point construction of the transient. In the following we will discuss a method giving a direct polar display of the transient with higher accuracy and with an easy time readout. The polar-display test set is shown in detail in Fig. 2. The detectors A and B each receive a part of the modulated signal and also a part of the unmodulated carrier as a reference signal.

The CW signal s_1 entering the modulator is

$$s_1 = \mathrm{Re}\,[S_1] = \mathrm{Re}\,[S_0 e^{j\omega_0 t}] \tag{2a}$$

where S_0 is the amplitude and ω_0 the angular carrier frequency. The signal s_2 after the modulator has a time-dependent amplitude $S_0\,m(t)$ and a time-dependent phase $\phi(t)$

$$s_2 = \mathrm{Re}\,[S_2] = \mathrm{Re}\,[S_0 m(t) e^{j[\omega_0 t + \phi(t)]}]. \tag{2b}$$

The complex transmission factor S_{21} of the modulator is the ratio of the complex output signal S_2 to input signal S_1, as defined in (2a) and (2b)

$$S_{21} = \frac{S_2}{S_1} = m(t) e^{j\phi(t)}. \tag{2c}$$

Each detector receives the same fraction α of s_2 and a reference signal. The CW reference signal r_A for detector A can be adjusted to

$$r_A = R_0 \cos\left(\omega_0 t + \frac{\pi}{2}\right) \tag{3a}$$

and r_B for detector B to

$$r_B = R_0 \cos(\omega_0 t) \tag{3b}$$

where R_0 is the reference amplitude.

Under the assumption of square law detectors, the lowest generated frequency components of interest at the detector outputs are

$$a = K\left[\tfrac{1}{2}R_0^2 + \frac{\alpha^2}{2} S_0^2 m^2(t) + \alpha R_0 S_0 m(t) \sin\phi(t)\right] \tag{4a}$$

$$b = K\left[\tfrac{1}{2}R_0^2 + \frac{\alpha^2}{2} S_0^2 m^2(t) + \alpha R_0 S_0 m(t) \cos\phi(t)\right]. \tag{4b}$$

The constant K reflects the detector sensitivity. A comparison between (4a) and (4b) with (2c) shows that the terms $\alpha K R_0 S_0 m(t) \sin\phi(t)$ and $\alpha K R_0 S_0 m(t) \cos\phi(t)$ are proportional to the imaginary and real part of the transmission factor S_{21}, if specific modulator reference planes are selected.

By keeping

$$|\alpha S_0 m(t)| \ll R_0 \tag{4c}$$

the term $(\alpha^2/2) K S_0^2 m^2(t)$ can be dropped and the constant dc term $\frac{1}{2}KR_0^2$ can easily be separated from time-varying terms.

If balanced mixers were available, the dc offset would automatically disappear.

IV. Experimental Results

The higher the millimeter-wave frequency, the less sensitive the detectors become. At the test frequency of 80.99 GHz the two commercial low-noise (noise figure < 4 dB) broadband (dc-3 GHz) 20-dB preamplifiers were therefore essential for a reasonably sharp sampling scope display.

A display of A versus B on a sampling scope, with compensation for the constant dc components so that the origin is in the center, plots the transient magnitude as a function of phase, as shown in Fig. 3. The transients from forward to reverse and reverse to forward take different paths, but both turn out to be close to resistive.

Fig. 3(a)–(h) result from 45° stepping of the reference phase. In a perfect test set a change in reference phase should

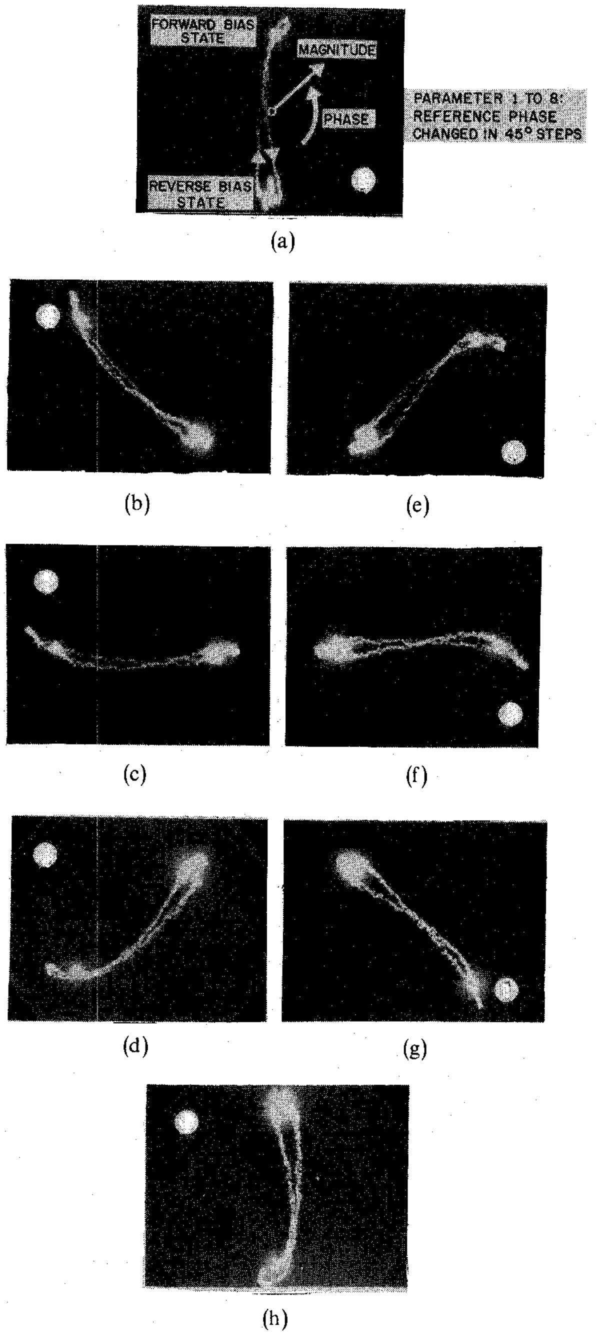

Fig. 3. Switching trajectories of modulator operating at 300 Mbit/s with reference phase as parameter.

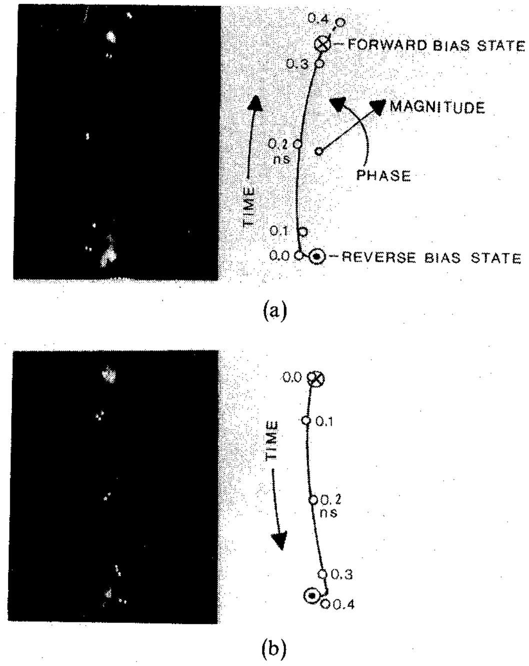

Fig. 4. Time dots on (a) reverse to forward and (b) forward to reverse switching transient.

only rotate the transient but not change its shape. The observed changes are insignificant in terms of system's performance. In order to arrive at a display with changes as small as shown, the millimeter-wave and baseband components have to be carefully selected. The corresponding components in branch A and branch B of Fig. 2 should be matched in performance and the reference phase shifter should be free of loss variations as a function of phase setting.

In addition to knowing the trajectory path of the transient, the time parameter corresponding to each point on the path must be known accurately. The bright-dot feature of the sampling scope used is very convenient in determining the time. The bright dot can be moved in a calibrated manner along the trace.

In Fig. 4, which corresponds to Fig. 3(a), the brightness is reduced to the point where only the bright time dots and the locations of the forward- and reverse-bias states are visible. The time separation between the dots is 0.1 ns. Fig. 4(a) shows the reverse to forward transient and Fig. 4(b) the forward to reverse transient. One sees from the time-dot separation that the switching speed is fastest in the center portion in each case. The rise times are approximately equal

$$T_R \approx 0.4 \text{ ns} < \frac{T}{4} = 0.9 \text{ ns}$$

well below the system requirement.

One observes further that the transients overshoot the new state before settling into it, see Fig. 3(a) or 4(a). The corresponding overshoot with oscillatory ripples is also visible on the amplitude- and phase-detected signals, Fig. 5(a) and (b), respectively. The overshoots are partly due to the tuning of the millimeter-wave detector and partly to the millimeter-wave circuit. The ripples were found to be directly related to the bandwidth of circulator 2. For the results shown in Figs. 3 and 5 the circulator bandwidth was 0.6 GHz, defined by the isolation exceeding 25 dB. With a circulator having a 4-GHz bandwidth, the ripples practically disappear. The explanation is that the fast phase modulation causes appreciable frequency components which, for the case of the narrow-band circulator, are partially outside the well-matched and well-isolated frequency range. This in turn leads to a multiple path propagation from the p-i-n diode to the output with interference effects on the output signal.

No measurable system performance impairment could actually be traced to the described overshoots.

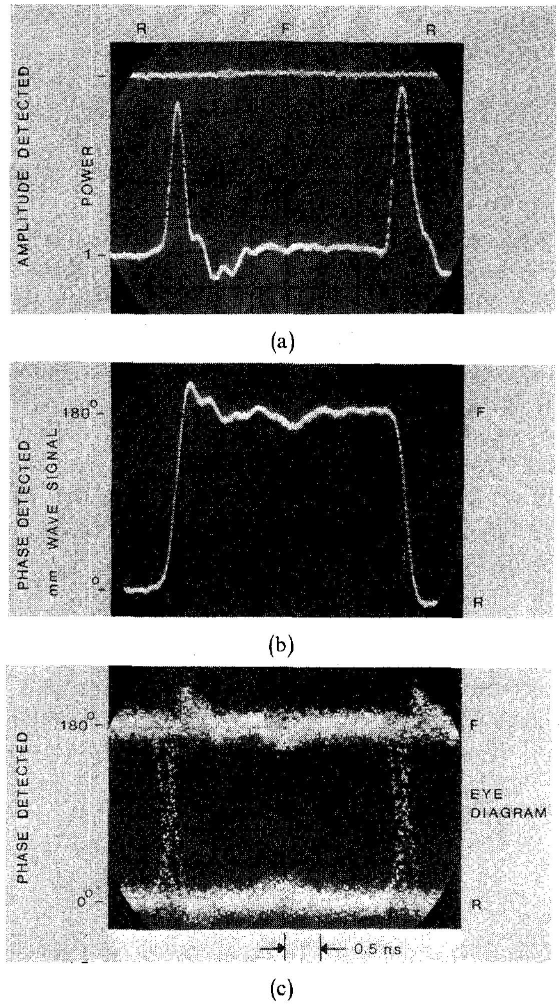

Fig. 5. Modulator output at 300 Mbit/s. (a) Amplitude-detected signal. (b) Phase-detected signal. (c) Eye diagram of (b).

Fig. 5(a) shows deep amplitude attenuation dips. From this fact alone one expects a strongly resistive transient as has been demonstrated by the detailed polar transient display.

The timing error ε_T can be evaluated from the eye diagram of the phase-detected signal, Fig. 5(c). The 0° state corresponds to reverse bias, the 180° state to forward bias. The reverse to forward transient is one line while the forward to reverse is split up into several lines as a consequence of the charge-storage effect in the p-i-n diode. The stored charge and, therefore, the time required to remove it, depends on the number of immediately preceding time slots in the forward-bias state. The pseudorandom word used tests many possible bias combinations. From Fig. 5(c) one reads a timing error, which is the half-width of the timing jitter

$$\varepsilon_T \approx \pm 0.1 \text{ ns}$$

and therefore

$$|\varepsilon_T| < \frac{T}{16} = 0.23 \text{ ns}$$

well within the system specifications.

The minimum achievable rise time and timing error depend on the driver performance, the millimeter-wave switch layout, and p-i-n diode time characteristics, which are storage and fall time [8]. The diode in the 80.99-GHz sample modulator has a storage time of 3.0 ns and a fall time of 0.3 ns.

V. Conclusions

A test set has been described which provides a polar display of subnanosecond transients of the complex transmission factor for a p-i-n diode modulator operating at 300 Mbit/s. An important feature is that the time parameter can be read out very conveniently along the polar trace with a time-calibrated bright dot. This technique is applicable to the study of PSK modulators used in ultra-high-speed digital transmission systems. While the results obtained here were for a two-phase modulator, the technique may be used for modulators with any number of phase states.

The modulator transients were found to approximate the purely resistive type for which the measured rise times and timing error are well compatible with the system performance objective.

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