Filters for High-Speed Diode Modulators and Demodulators

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Abstract—**The relationship between switching time, transient suppression, and filter complexity is derived for filter-type diode switches, along with the theoretical minimum switching time. Design criteria are developed for high-speed switches and detectors in waveguide and TEM transmission lines.**

Introduction

DIODE SWITCHES are used not only for providing high isolation and amplitude modulation but also as basic building blocks in making phase modulators [1], [2] and frequency translators [3], [4]. In many cases the modulation rate is limited by filters associated with the modulator rather than by the diode itself.

For example, in a radar application a diode in the receiver line is switched into the reflecting state to prevent stray transmitted power from saturating the receiver. When the diode is switched too fast, switching transients can cause receiver saturation that the diode is supposed to prevent. In another example, a line modulator is placed before a detector and a sensitive narrowband amplifier for instrumentation purposes. Transients from the modulator can cause an indication of power in the line even when the RF source has been disconnected from the line. In the case of a phase modulator used in a scanning array antenna, the limit on the modulation rate limits the scan rate. In a frequency translator, the amount of frequency shift is limited by the modulation rate.

The problem of limited modulation rate is that of separating modulation power and RF power [5] which appear across the same two-terminal diode junction. When the modulation frequency is low and the RF is high, there is no problem separating the powers with simple filters. As the modulation rate increases, the modulation power has higher frequency components and the modulated RF power has lower (as well as higher) frequency components. For increasing modulation rates, modulation power spectral components and modulated RF power spectral components approach each other, requiring filters to have steeper skirts to separate the powers. When the powers are not separated well enough, modulation power appears in the RF transmission line, since modulation transients, and troublesome RF power can be coupled to other parts of the system via the modulation circuit. To demodulate a carrier with wide video bandwidth information on it, the demodulator must satisfy filter requirements similar to those of the modulator.

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Another technique exists for separating modulation power and RF power but it will not be developed in this paper. By using two or four matched diodes and one or two balanced transmission lines the powers can be separated even when their spectral components overlap. For example, Butson and Glover [6] put diodes on the top and bottom walls of a waveguide (a balanced transmission line) and took their IF output from the common point of the diodes on a line parallel to the broad waveguide wall and emerging from the center of the narrow waveguide wall, this line being the center conductor of the coaxial (unbalanced) transmission-line output. Voltage and current vectors having equal and opposing components in their undesired transmission lines keep the two lines isolated. An impedance asymmetry of 1 percent will provide 40 dB isolation between the two transmission lines. As another example, the sampling gates of "sampling" oscilloscopes typically have four matched diodes in the two balanced transmission lines. Modulation current through the diodes develops negligible current in the RF transmission line.

Minimum Switching Time

The general characteristics of filters for diode modulators [7] may be seen from Fig. 1. Unmodulated RF power enters through port 1. Baseband modulation is supplied through port 2. The modulated RF power emerges through port 3. The voltage waveforms and the spectra of the power at each port are shown adjacent to the port numbers. The filter on port 1 needs to pass power at f_0 only; however, it should not deteriorate the waveform of the modulation power; that is, it must be a high-impedance circuit from dc up to some high frequency, as seen from port 2. The filter on port 2 must pass power at the lower frequencies and stop RF power. In addition, this filter must present a high impedance to RF power as seen from port 1 and the diode. The filter on port 3 must pass power at all frequencies above and below f_0 corresponding to the bandwidth of the modulation power. This filter must also present a high or low impedance to the diode at the modulation frequencies depending on whether the diode is in shunt or in series. For pulse modulation as shown in Fig. 1 it is also recommended that the filters be constant-time-delay filters to prevent "ringing."

The minimum switching time may be derived with the aid of Fig. 2. A low-pass filter with a finite 3 dB bandwidth, BW, will cut off some of the infinite spectrum of a zero rise-time pulse giving it a 10 to 90 percent switching time of

$$\tau_F = 0.44/\mathrm{BW}. \tag{1}$$

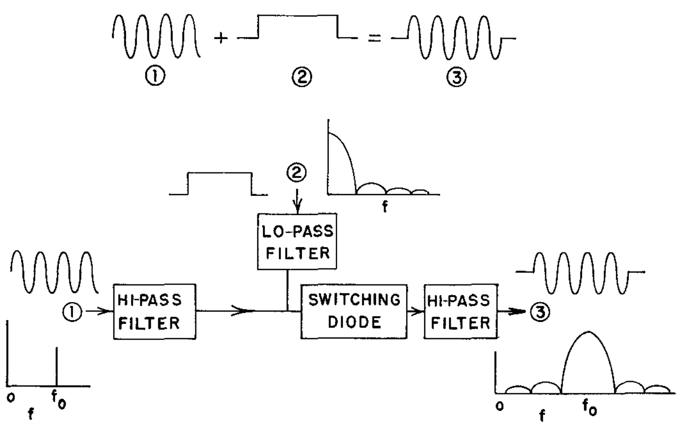

Fig. 1. Filter circuits for diode switches.

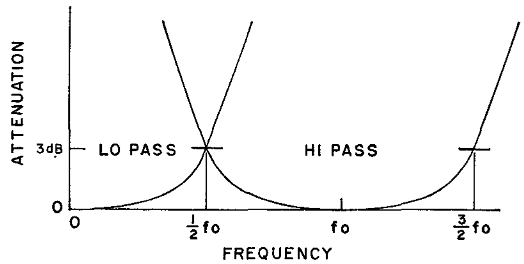

Fig. 2. Filter specifications for minimum rise time.

The high-pass filter must pass the spectral components of the modulated carrier or further deteriorate switching speed. Assuming that the difference between the carrier frequency and the cutoff frequency of the high-pass filter is equal to the bandwidth of the low-pass filter, then, because rise times add as the sum of their squares, the switching times of the modulated carrier are given by

$$\tau = \sqrt{2}\tau_F = \frac{0.44\sqrt{2}}{\mathrm{BW}} = \frac{0.62}{\mathrm{BW}}\,. \tag{2}$$

Since the spectra have to be separated, the best arrangement of filters from a switching speed point of view is to have their 3 dB frequencies coincide at $f_0/2$ as shown in Fig. 2. Then the minimum switching time becomes

$$\tau_{\min} = \frac{0.62}{f_0/2} = \frac{1.24}{f_0}\,. \tag{3}$$

This equation indicates that the lowest switching time that could be generated using filters on a 1 GHz carrier is 1.24 ns, and on a 10 GHz carrier is 0.124 ns.

Waveguide Modulators and Demodulators

Experimental work was done in X-band waveguide at 9.3 GHz. A typical X-band diode switch is shown in Fig. 3. The waveguide performs the function of high-pass filters with a cutoff frequency of 6.5 GHz. Because this is 2.8 GHz below the carrier frequency, the waveguide will have a rise time of about 0.16 ns. The 10 pF capacitor performs the function of the low-pass filter. For a 50 ohm pulse generator driving a 50 ohm diode switch the filter response is as shown in Fig. 4. The 3 dB point is at 0.6 GHz which for a capacitor provides a rise time of 0.6 ns.

The rise time of the low-pass filter can be decreased to 0.1 ns by building the π section filter also shown in Fig. 4. This filter has an additional advantage over the capacitor in that its attenuation of RF power becomes very high at the design frequency. The physical embodiment of this filter is shown in Fig. 5. Because of the capacitance existing between the two ends of the center conductors of the 31 ohm sections, the length of the 100 ohm section had to be made shorter than $\lambda_0/4$ to present an inductance for parallel resonance with this capacitance at the design frequency.

To evaluate this technique a pulse modulator and a demodulator were made up of such filters in conjunction with 1N263 point-contact germanium diodes. The demodulated waveforms are shown in Fig. 6. The observed OFF–ON time is 0.30 ± 0.05 ns and ON–OFF time is 0.18 ± 0.02 ns. A 0.25 ns rise-time pulse generator was used in conjunction with a 0.1 ns rise-time oscilloscope. The theoretical rise time of the combination pulse generator, modulator input filter, waveguide (as a high-pass filter), demodulator output filter, and oscilloscope is 0.34 ns.

Several factors may cause the data to be different from the simplified theory. First, the impedance of the modulator and demodulator diodes is not 50 ohms as assumed for calculating the filter responses and their rise times. When the modulating diode is ON, it is in the conduction state and presents a very low impedance to the modulation source. Similarly when the demodulator diode is providing rectified current (also during ON time), it has a low impedance. The high impedance of the diodes associated with OFF is also different from the characteristic impedance of the filter as portrayed in Fig. 4. Thus the pulse is reflected by the diode as an open circuit and travels back to the 50 ohm pulse generator. To prevent these mismatches from deteriorating the data, the pulse generator was padded by either attenuators or a time delay (line length between pulse generator and diode modulator).

Bias to the diode modulator, opposite in polarity to the pulse, was supplied through the other broad waveguide wall in a circuit similar to Fig. 3 (turned upside-down). The longer the pulse, the larger the required capacitance. This capacitance is part of a diode mount used for measuring diodes. The waveguide walls are thicker than normal and a block is cut from the bottom of the waveguide and replaced with a thin dielectric insulator where the cut was made. The broad waveguide wall is cut $\lambda_g/4$ in front of and behind the diode. The narrow waveguide wall is cut through to the inside face of the bottom broad waveguide wall. The waveguide wall thickness is a quarter wavelength in the insulation dielectric. This capacitor provides minimum interference with the RF power and can be added to externally for longer pulses. This capacitor should not deteriorate the pulse waveform.

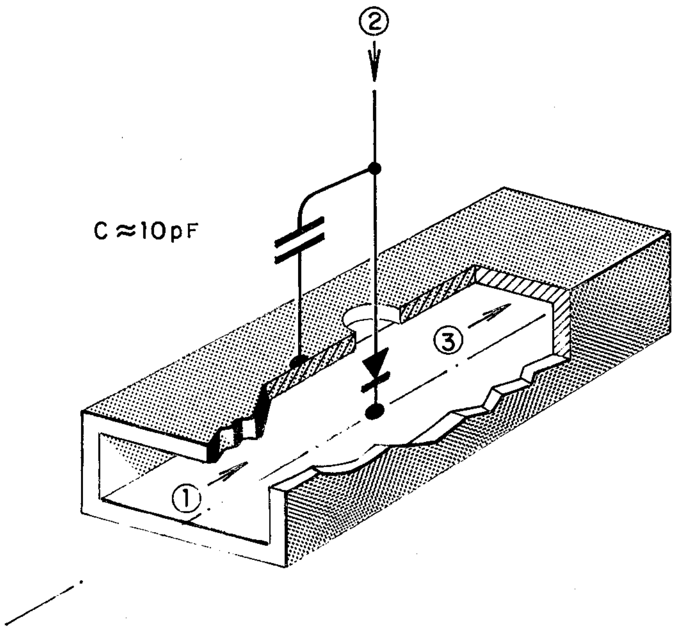

Fig. 3. Conventional waveguide diode switch.

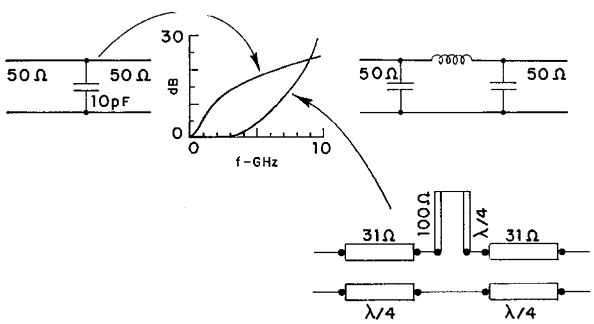

Fig. 4. Frequency response of low-pass filters.

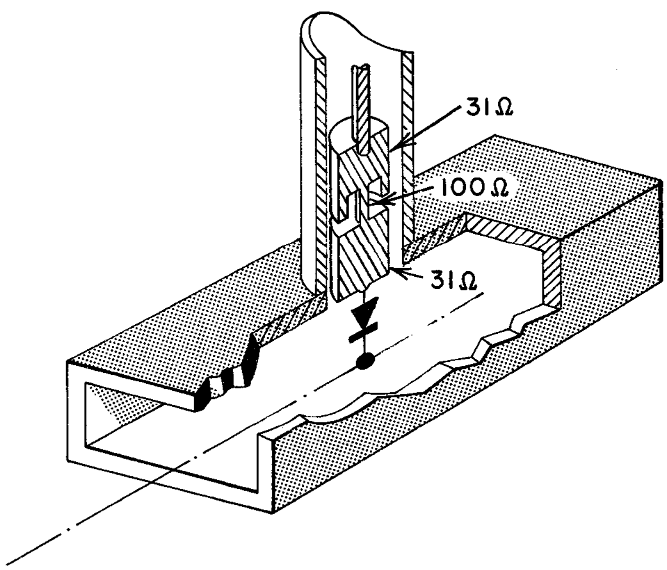

Fig. 5. Construction of 0.1 ns low-pass filter for X-band diode switching.

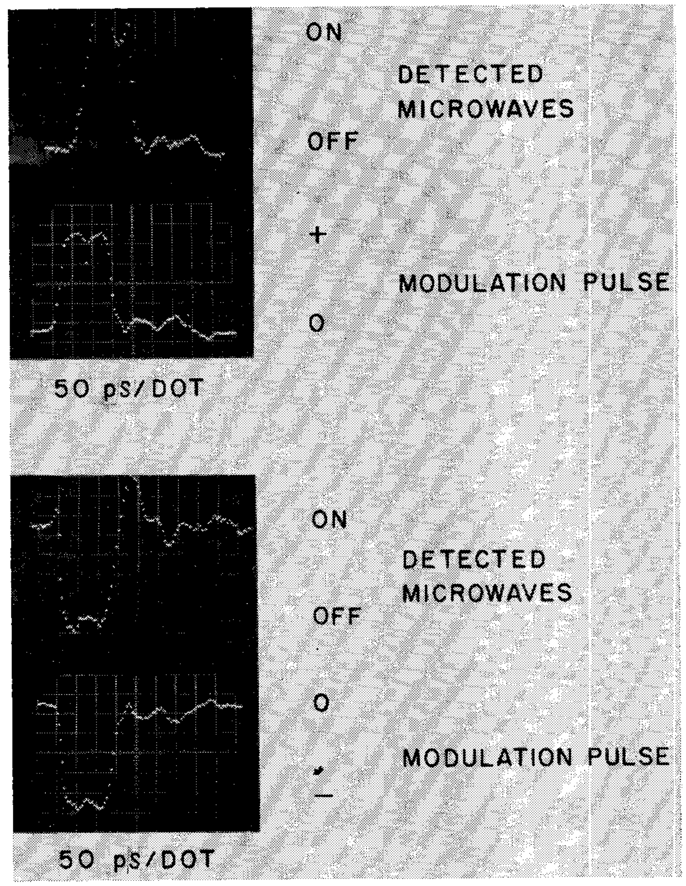

Fig. 6. Modulation pulse and detected output at X band.

A second factor that might alter the pulse waveform is the nonlinearity between voltage into the modulator and voltage out of the demodulator [8]. This nonlinearity can cause the experimental switching times to be less than the theoretical switching times.

A third factor possibly altering the demodulated pulse waveform is dispersion in the low-pass filters and in the waveguide. A more sophisticated system would have constant-time-delay low-pass filters and a waveguide delay equalization circuit.

Theoretically a zero rise-time pulse generator used with these filters would impart a rise time to the microwave power of about 0.19 ns compared with a minimum with optimum filters of 0.135 ns.

TEM Transmission-Line Modulators and Demodulators

Normally a diode modulator is designed to provide maximum RF bandwidth. The biasing lead and ground return are high-impedance quarter-wavelength lines connected to the center conductor with a lumped circuit capacitor providing an RF ground at the bias input [7], [9], [13]. These bias leads allow moderate modulation speeds but the theoretical limit may be more closely approached and the suppression of RF power out of the modulation port may be increased by using the filter circuits shown in Fig. 7.

The attenuation of the modulation input filter (port 2 to port 3 with the diode replaced by a Z_0 transmission-line

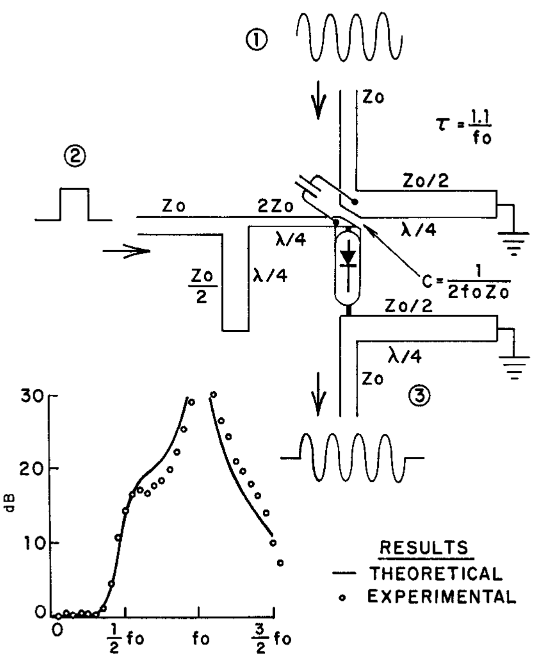

Fig. 7. Low-pass filter for TEM diode modulators.

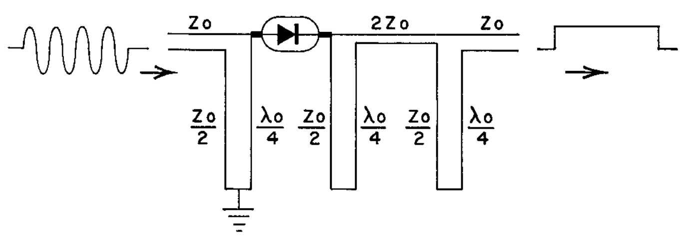

Fig. 8. Low-pass filter for diode demodulators.

section and the port 3 $\lambda/4$ stub removed) is shown at the bottom of Fig. 7. The low-frequency π equivalent circuit of the modulation input filter is a 1 dB ripple Chebychev filter with a 3 dB bandwidth of $2/3\, f_0$. The same basic filter is shown in Fig. 8 as it would be used in a demodulator. The series inductance in the π equivalent circuit is derived from the high-impedance $2Z_0$, $\lambda_0/4$ line section ($L_{2Z_0}=4Z_0/f_0$). Each shunt capacitance is derived from the low-impedance $Z_0/2$ open-circuit stubs ($C_{Z_0/2}=1/2f_0Z_0$) and half the $2Z_0$ line ($C_{2Z_0}/2=1/16f_0Z_0$).

The ground returns in both figures are $Z_0/2$, $\lambda_0/4$ stubs. One of these by itself has a critically damped response to a step waveform. If the characteristic impedance of the stub were higher, the transient response to a step voltage would be a fast-rising slowly decaying pulse. If the characteristic impedance were lower, the response would be a damped oscillation at f_0. The response of one $Z_0/2$ stub to a step voltage is a pulse having half the voltage of the step and one half an RF cycle at f_0 in duration. When more than one stub is used without attenuation between them, they interact and produce transients of longer duration. The simpler analysis of multiple stub makes use of conventional filter theory. A single $Z_0/2$ stub provides 3 dB attenuation at $f_0/2$ and has a rise time of about $0.7/f_0$. Adding rise time as before, the theoretical rise time of 2.0 ns was measured when the modulator of Fig. 7 was used with the demodulator of Fig. 8. Hot carrier diodes were used as modulator and demodulator at 1 GHz. Some difficulty was encountered in arranging filter spacing and attenuators so that transient voltages were suppressed yet pulse shape was preserved.

Suppression of Modulation Transients

Ideally the application of step voltage modulation to port 2 of Fig. 1 would cause no voltage to appear at port 3. Due to the broad spectrum of the step and the finite attenuation of the combination of low-pass filter and high-pass filter, the step causes a transient to emerge from port 3 as shown in Fig. 9. Transmission measurements of combined low-pass and high-pass filters on a "time domain reflectometer" have demonstrated that the transient response is similar to that generated by the application of a step voltage to a bandpass filter. This combination of low-pass filter and high-pass filter also gives attenuation similar to that of a bandpass filter but with high attenuation inside the frequency band as shown approximately in Fig. 10. The cutoff frequency (3 dB point) of the low-pass filter is f_1 and of the high-pass filter is f_2. The attenuation formulas shown are approximations for maximally flat filters with n elements in their lumped-element low-pass prototypes. The attenuation for constant-time-delay filters is slightly lower and for Chebychev filters is slightly higher [14]. Note that increasing the separation between f_1 and f_2 increases the attenuation inside the passband, thus reducing the magnitude of the transients. The increased separation also increases the rise time. Therefore, a compromise has to be made between transient suppression, number of elements in the filter, and rise time.

Approximations are made in the calculation of this compromise relationship because the exact solution of the transient as portrayed in Fig. 9 required a computer program [15]. The analytic expressions for the envelopes shown in Fig. 9 are approximations of the transients calculated by Henderson and Kautz [15]. The symbol f_B is the 3 dB bandwidth of the filter and f_c is its center frequency. For the purpose of calculating the transients the input is assumed to be a square wave as shown in Fig. 11. Time duration τ_p is made long enough that the transients from one step decay before the transients from the next step begin. The suppression is defined as

$$S = 20 \log \frac{A}{V_{0\,\max}}. \tag{4}$$

The voltage of each harmonic $V(n')$ of a perfect square wave, as shown in Fig. 11, is

$$V(n') = \frac{A}{2}\,\frac{\sin\dfrac{\pi n'}{2}}{\dfrac{\pi n'}{2}} \tag{5}$$

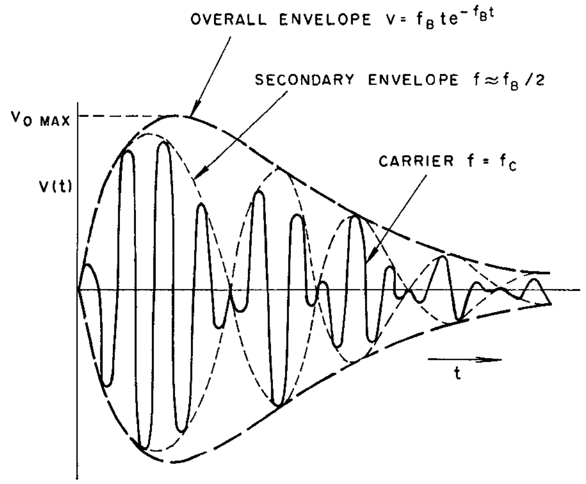

Fig. 9. Typical waveform of the switching transient appearing on the RF terminals of a diode switch.

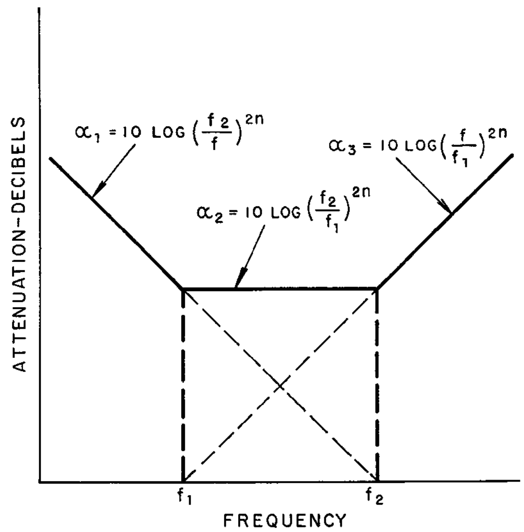

Fig. 10. Attenuation of a low-pass filter and a high-pass filter in series (by which modulation voltage is reduced to suppress transients).

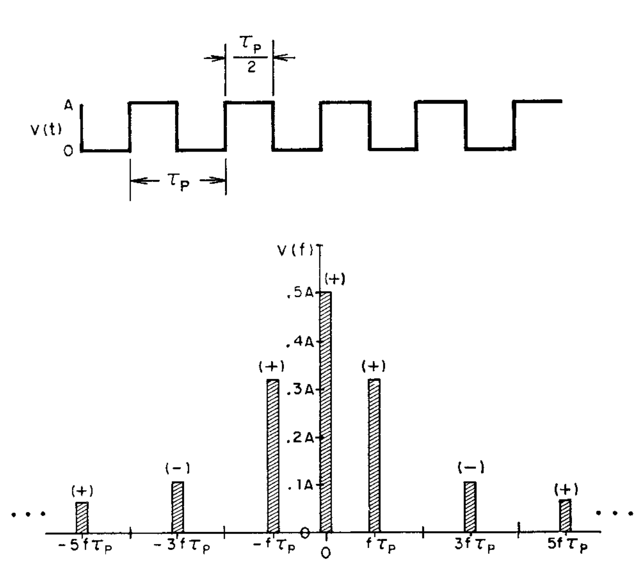

Fig. 11. Time and frequency characteristics of modulation for transient suppression analysis.

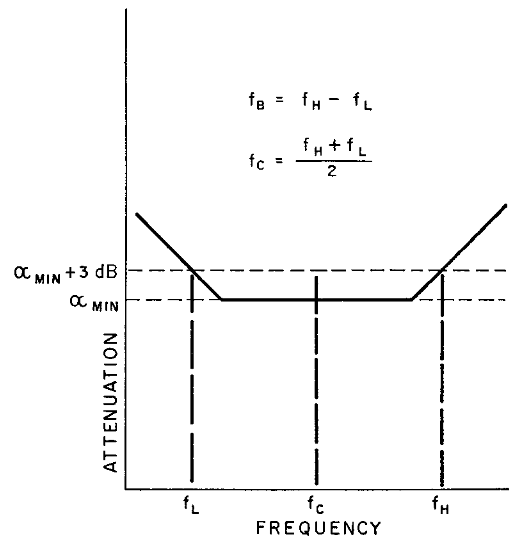

Fig. 12. Portrayal of frequencies of Fig. 9 from attenuation characteristics of Fig. 10.

in which n' is an integer varying from $-\infty$ to $+\infty$. For n' an even number, the voltage is zero. For n' odd,

$$|V(n')| = \frac{A}{\pi n'} \cdot \tag{6}$$

For n' sufficiently large,

$$\frac{A}{\pi n'} \approx \frac{A}{\pi(n'+1)} \cdot \tag{7}$$

Thus the voltage terms of the negative odd n' may be approximately folded into the empty places left by the positive even n', and the spectrum of the square wave can be written

$$|V(n')| \approx \frac{A}{\pi n'} \tag{8}$$

for n' from 0 to ∞. Thus the power at each harmonic can be written

$$P(f\tau_p) = \frac{A^2}{\pi^2(f\tau_p)^2} \cdot \tag{9}$$

The power past the filters can be found by adding up the power at each $f\tau_p$ after it has been attenuated. To facilitate integration, the attenuation of a maximally flat filter

$$\alpha = 10 \log \left[1 + (f/f_0)^{\pm 2n}\right] \tag{10}$$

is approximated by

$$\alpha = 10 \log (f/f_0)^{\pm 2n} \tag{11}$$

as shown in Fig. 10 in the attenuation band, or $\alpha = 0$ otherwise. The approximation introduces negligible error for

$$(f/f_0)^{\pm 2n} > 10 \tag{12}$$

which corresponds to attenuation for each filter greater than 10 dB. The average power getting past the filters is obtained by integration over the sections of Fig. 10 as follows:

$$P_{AV} = \int_0^{f_1\tau_p} \frac{P(f\tau_p)}{\left(\frac{f_2\tau_p}{f\tau_p}\right)^{2n}} d(f\tau_p) + \int_{f_1\tau_p}^{f_2\tau_p} \frac{P(f\tau_p)}{\left(\frac{f_2\tau_p}{f_1\tau_p}\right)^{2n}} d(f\tau_p)$$

$$+ \int_{f_2\tau_p}^{\infty} \frac{P(f\tau_p)}{\left(\frac{f\tau_p}{f_1\tau_p}\right)^{2n}} d(f\tau_p)$$

$$= \frac{A^2}{\pi^2}\cdot\left(\frac{f_1\tau_p}{f_2\tau_p}\right)^{2n}\left[\frac{f_2\tau_p}{f_1\tau_p}\frac{1}{1-\frac{1}{2n}} - \frac{1}{1+\frac{1}{2n}}\right]\frac{1}{f_2\tau_p}\cdot \quad (13)$$

The f_B must be determined. Referring to Fig. 10, the minimum attenuation of the two overlapping filters is the ratio $(f_2/f_1)^{2n}$. Referring to Fig. 12, the 3 dB bandwidth will be the difference between the f_H and f_L as determined from the following equations:

$$10 \log\left(\frac{f_H}{f_1}\right)^{2n} = 3.0 + 10 \log\left(\frac{f_2}{f_1}\right)^{2n} \quad (18)$$

$$10 \log\left(\frac{f_2}{f_L}\right)^{2n} = 3.0 + 10 \log\left(\frac{f_2}{f_1}\right)^{2n} \quad (19)$$

Solving

$$f_B = \left(2^{1/2n}\frac{f_2}{f_1} - \frac{1}{2^{1/2n}}\right)f_1. \quad (20)$$

The ratio of A and $V_{0\,max}$ which gives the attenuation may be derived by equating $P_{AV} = P_T$

$$\alpha = 10 \log \frac{A^2}{V_{0\,max}^2} = 10 \log \left\{\frac{\pi^2 e^2 (f_2/f_1)^{2n+1}}{4\left(2^{1/2n}\frac{f_2}{f_1} - 2^{-1/2n}\right)\left(\frac{1}{1-\frac{1}{2n}}\frac{f_2}{f_1} - \frac{1}{1+\frac{1}{2n}}\right)}\right\}. \quad (21)$$

The expression is not correct for $n=1$ because the waveform is not as assumed. Using Laplace transforms, the attenuation of two single-element filters consisting of a series inductance and a shunt inductance is given exactly by

$$\alpha = 20 \log\left\{\left(2\frac{f_2}{f_1} + 1 - M\right)\left[\frac{2\frac{f_2}{f_1} + 1 + M}{2\frac{f_2}{f_1} + 1 - M}\right]^{\frac{1}{2}(2(f_2/f_1)+1+M)/M}\right\}, \quad M = \sqrt{\left(2\frac{f_2}{f_1}\right)^2 + 1}. \quad (22)$$

The general output transient response is assumed to be of the form

$$V_0 = K(f_B t)e^{-f_B t}\cos\omega_c t. \quad (14)$$

Dropping the secondary envelope term from the expression does not significantly alter the average power in each wave packet or the maximum voltage of the wave packet. The maximum voltage is given by

$$V_{0\,max} = \frac{K}{e}\cdot \quad (15)$$

The energy in each wave packet is given by

$$\int_0^\infty V_0^2 dt = \frac{K^2}{8f_B}\cdot \quad (16)$$

Since there are $2/\tau_p$ envelopes per second, the average transient power is

$$P_T = \frac{e^2 V_0^2{}_{max}}{4 f_B \tau_p}\cdot \quad (17)$$

The attenuation is a function of n and f_2/f_1. A more convenient relationship would give attenuation as a function of increased switching time.

From the equations shown in Fig. 13, the following relationship is derived:

$$\frac{f_2}{f_1} = 2\frac{\tau}{\tau_{min}} - 1. \quad (23)$$

The attenuation for various n is shown in Fig. 14.

As an example of an application of the relationship demonstrated in Fig. 14, how fast can a 3 GHz carrier be switched if the required switching voltage for a very fast switching diode is 10 V and transients below -100 dBm peak will not disturb the system? Since 10 V correspond to $+33$ dBm in a 50 ohm transmission line, the suppression of switching transients must be 133 dB. For 3 GHz minimum rise time (τ_{min}) is 0.41 ns. Three-element filters would give the required suppression with 41 ns rise times. Seven-element filters would give the required suppression with 1.9 ns rise times. If a conventional two-element filter were designed

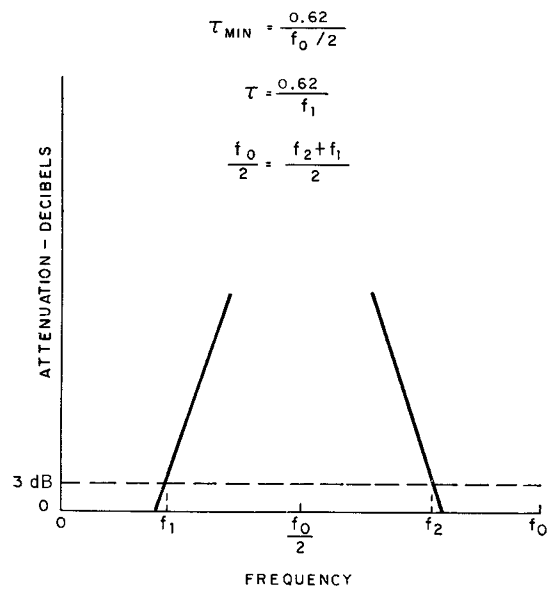

Fig. 13. Attenuation of overlapping filters and their resulting switching times.

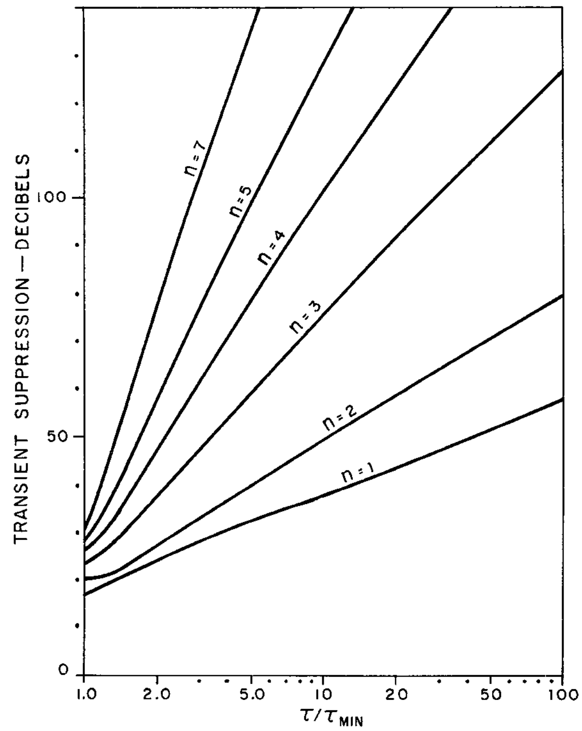

Fig. 14. Transient suppression versus switching time of baseband prototype filter for increasing number of elements.

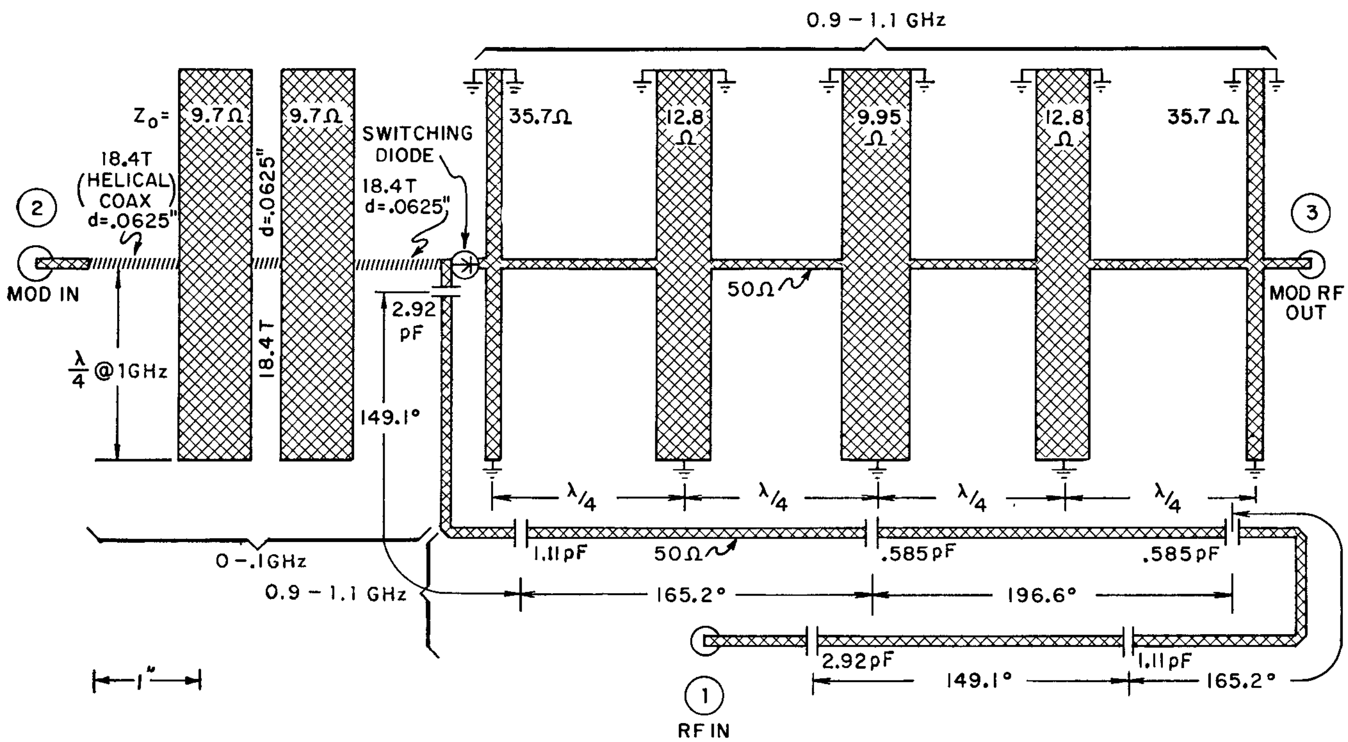

Fig. 15. Stripline circuit of a 6.2 ns diode switch designed for 1 GHz to provide 100 dB transient suppression.

for 10 ns switching speed at this frequency, the suppression would be 61 dB. Any desired suppression of switching transients can be obtained at the cost of filter complexity and increased switching time.

As another example, suppose a diode switch is needed to protect a 1 GHz radar receiver from "transmit" signal leakage. The diode requires 10 V to cause switching, and the transients must be below 200 μV to prevent receiver saturation. The switching pulse is generated by a matched 50 ohm generator so that any of the pulse reflected by filters or the diode is absorbed back in the pulse generator. Therefore, 94 dB suppression is needed. Figure 14 indicates that 100 dB suppression may be obtained by using five-element filters at a sacrifice of increasing the switching time by a factor of 5. Thus the switching speed becomes 6.2 ns. The filters for a series diode switch in stripline are shown in Fig. 15. The design of the low-pass and series capacitor filters are taken from *The Microwave Engineers Handbook* [16], and the design of the shorted quarter-wavelength stub filter is taken from Mumford [17]. When used with a hot-carrier diode as a switch, the rise time was 3.6 ns and the transient suppression was 50 dB. The pulse-shaping effect of the diode as evidenced by the shortened rise time decreased the modulation suppression by a factor of two.

Conclusion

To obtain high-speed modulation with a diode the following rationale is suggested.

1) To keep modulation transients low, a modulating diode should be used requiring only very low voltages for modulation, and modulation voltage should be kept as low as possible.

2) The filters for further suppressing the modulation transients may be designed with the help of Fig. 14.

3) Modulation circuit simplification may be accomplished by using a modulation source with a rise time no faster than possible according to Fig. 14. The slow rise time then accomplishes the same effect as the low-pass filter.

4) Waveguide for a high-pass filter should be used if possible, since this corresponds to an $n=\infty$ type filter.

5) Balanced transmission-line techniques may be used to obtain additional suppression of modulation transients when required.

6) Diodes no faster than required should be used otherwise they will cause pulse shaping and reduction of desired transient suppression. Diodes demonstrating step-recovery characteristics should not be used when transients will be a problem. The transients of a step-recovery diode will not be reduced by a slowly rising pulse whether from the pulse generator or a low-pass filter.

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