Circuit and Electromagnetic System Design Notes

Note 48

27 March 2004

Cleared
AFRL/DE 04-267
7 Jun 04

Compression of Sinusoidal Pulses for High-Power Microwaves

Carl E. Baum
Air Force Research Laboratory
Directed Energy Directorate

Abstract

A technique for generation of high-power microwaves consists of taking a low-power CW source and shortening it into a high-power pulse. One way to do this is to build up a resonance in some kind of cavity and then switch this into a load. Here we exhibit sections of coaxial TEM transmission lines with switches for this purpose. An alternate approach known as binary pulse compression is also discussed.

This work was sponsored in part by the Air Force Office of Scientific Research, and in part by the Air Force Research Laboratory, Directed Energy Directorate.

1. Introduction

In generating pulses of sinusoidal electromagnetic waves one may use high-power microwave tubes for hypoband (narrow band ~1% or so). For shorter pulses, such as from the switched oscillator [1, 3, 4], one has a mesoband source [2, 10] (medium band ratio). This is to be distinguished from hyperband (impulsive) sources with band ratios over a decade.

The basic switched oscillator [4] consists of a section of low-characteristic-impedance transmission which is charged to high voltage and then shorted to the outer conductor through a closing switch at the end opposite the antenna load. This gives the basic frequency with quarter wavelength given by the length of the oscillator. This decays like a damped sinusoid as energy is fed to the load. There are various approaches to increasing the output [1, 3]. This is limited by the inductance of the switch to frequencies in the range of hundreds of MHz if one wishes very high powers [7]. One can also increase the power by use of transmission-line transformers, at the expense of shortening the pulse [5, 6].

For high-power mesoband sources one would like to have some alternate kinds of sources which could achieve higher powers, and perhaps higher frequencies. This paper explores some possibilities based on building up resonances from lower power hypoband sources, sometimes called pulse compression.

This concept has been applied to the design of particle accelerators [8-12, 16, 17]. It has been suggested as a way to generate ultra-wideband pulses [19]. Pulse compression, as it has been termed, has been achieved in two general ways which might be called resonance buildup and binary pulse compression.

2. Simple Transmission-Line Oscillator Switched Out at One End

Applying the resonance-buildup concept, consider the scheme in Fig. 2.1, which may be appropriate in the hundreds of MHz regime. In this concept the transmission-line oscillator is an odd number of wavelengths in the length, ℓ, appropriate to boundary conditions of short circuit at one end and open circuit at the other end. The oscillator is excited by a low-power oscillator (or amplifier) at the resonance frequency

$$f_0 = \frac{\mathrm{v}}{\lambda_0} = \frac{\mathrm{v}}{\ell}\,\frac{2n+1}{4}$$

v ≡ propagation speed in transmission line, depending on dielectric constant (or c in air) (2.1)

The power is introduced at an appropriate location for buildup with f_0 maintained closely tuned to the coaxial cavity resonance (with closeness depending on the cavity Q). The achievable power multiplication is proportional to the cavity Q.

As the energy in the cavity builds up the output switch will eventually be overvolted and self break. The RF energy will then exit to the load. If the characteristic impedance Z_c of the oscillator is matched to the load, and if the switch impedance after closure can be neglected, the resulting RF pulse will have length

$$\begin{aligned} T &= 2t_r \\ t_r &= \frac{\ell}{\mathrm{v}} = \frac{\lambda_0}{\mathrm{v}}\,\frac{2n+1}{4} = \text{transit time of oscillator} \end{aligned} \tag{2.2}$$

One sees this by considering the fields as a superposition of left-going and right-going waves. This gives a number N of RF cycles

$$N = f_0 T = \frac{\mathrm{v}}{\lambda_0} T = \frac{2n+1}{2} \tag{2.3}$$

This assumes that the switch closure time is short compared to a half period of the oscillation. Otherwise the start and stop of the output pulse will be smoothed (stretched out).

One of the limitations on oscillator Q is power (real, not reactive) that leaks past the switch (toward the load). This is associated with the switch capacitance which should be minimized. This can be accomplished in part by shunting some of this capacitance to local ground (outer conductor) with an iris as indicated in Fig. 2.2. One can also partly overcome this by increasing the cavity length, ℓ, thereby increasing the stored energy and the Q. Note that the switch capacitance also affects the electrical length of the oscillator.

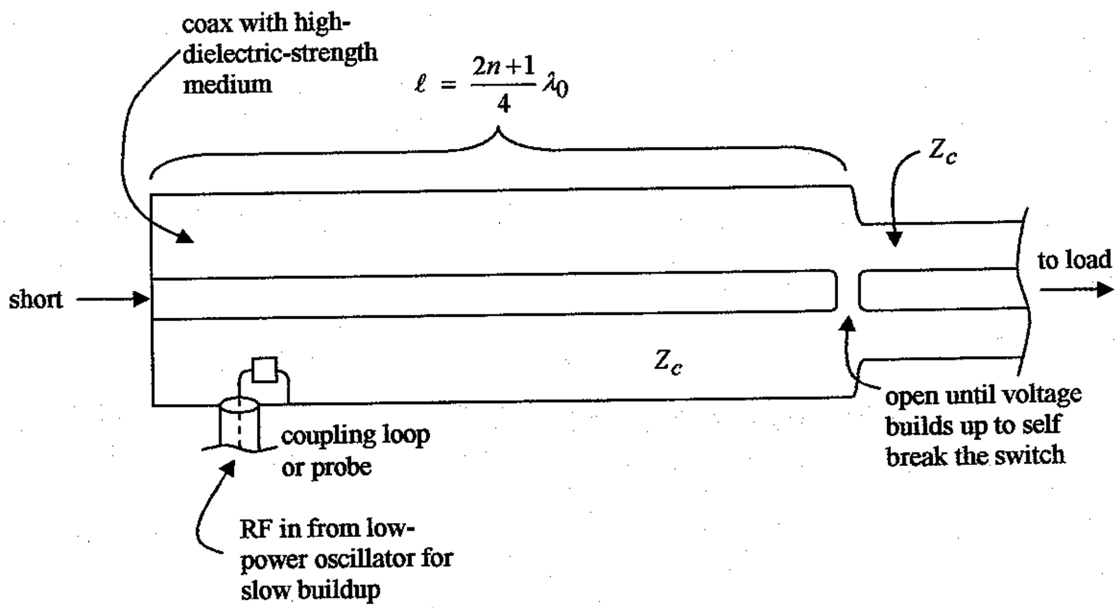


Fig 2.1. Resonance-Buildup Transmission-Line Oscillator Switched Out at One End.

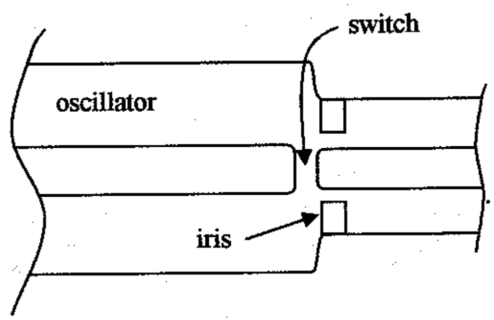


Fig. 2.2. Reduction of Switch Capacitance to Output Transmission Line.

There are various considerations for the design of the transmission-line conductors. For high Q one should minimize losses. Consider the case of a circular coax. We then have

$$
\begin{aligned}
Z_c &= \frac{1}{2\pi} Z_w \, \ell n\left(\frac{\Psi_2}{\Psi_1}\right) \simeq \frac{Z_0}{2\pi\varepsilon_r^{1/2}} \ell n\left(\frac{\Psi_2}{\Psi_1}\right) \\
Z_0 &= [\mu_0 / \varepsilon_0]^{1/2} \equiv \text{wave impedance of free space} \\
\varepsilon_r &\equiv \text{relative dielectric constant of dielectric medium} \\
&\simeq \begin{cases} 1 \text{ for air} \\ 2.25 \text{ for transformer oil or polyethylene} \end{cases} \\
\Psi_2 &= \text{inside diameter of outer conductor} \\
\Psi_1 &= \text{outside diameter of inner conductor} \\
\xi &\equiv \Psi_2 / \Psi_1 \\
Z_w &= \frac{Z_0}{\varepsilon_r^{1/2}} \equiv \text{wave impedance}
\end{aligned}
\tag{2.4}
$$

Losses occur by the skin effect. Neglect the small inductive part (compared to the inductance per unit length, L', of the coax). The resistance per unit length has the general form

$$
R' = \frac{R_1}{2\pi\Psi_1} + \frac{R_2}{2\pi\Psi_2} \tag{2.5}
$$

The propagation constant of the coax is

$$
\begin{aligned}
\gamma &= \left[[sL' + R']sC'\right] = s[sL']^{1/2}\left[1 + \frac{R'}{sL'}\right]^{1/2} \\
&= \gamma_0 \left[1 + \frac{R'}{2sL'} + O\left(\left[\frac{R'}{sL'}\right]^2\right)\right] \quad \text{as} \quad \frac{R'}{sL'} \to 0 \\
s &= j\omega = j2\pi f_0 \quad , \quad \gamma_0 = \frac{s}{\mathrm{v}} \\
C' &\equiv \text{capacitance per unit length}
\end{aligned}
\tag{2.6}
$$

The propagation takes the form (for propagation in the +z direction)

$$
e^{-\gamma z} \simeq e^{-\gamma_0 z} \, e^{-\frac{R'}{2Z_c} z} \tag{2.7}
$$

To minimize the attenuation we need to minimize (for fixed outer radius (size))

$$\frac{R'}{Z_c} = \frac{1}{Z_0}\left[\frac{R_1}{\Psi_1} + \frac{R_2}{\Psi_2}\right]\ell n^{-1}\left(\frac{\Psi_2}{\Psi_1}\right) = \frac{R'}{Z_0\Psi_2}[\xi R_1 + R_2]\ell n^{-1}(\xi) \tag{2.8}$$

If we neglect R_2 compared to R_1 due to the larger outer radius, we then need to minimize $\xi \ell n^{-1}(\xi)$. This occurs at

$$\xi = \frac{\Psi_2}{\Psi_1} = e \simeq 2.718 \tag{2.9}$$

for which

$$\begin{aligned} Z_c &\simeq \frac{120}{\varepsilon_r^{1/2}}\ell n(e) = 120\varepsilon_r^{-1/2} \\ &= \begin{cases} 120\,\Omega \text{ for air} \\ 80\,\Omega \text{ for oil or polyethylene} \end{cases} \end{aligned} \tag{2.10}$$

If we do not neglect R_2, the situation is more complicated. The case of $R_2 = R_1$ has been treated [14] giving an optimization condition of

$$\begin{aligned} &\xi \simeq 3.6 \\ &Z_c \simeq \begin{cases} 77\,\Omega \text{ for air} \\ 51\,\Omega \text{ for oil or polyethylene} \end{cases} \end{aligned} \tag{2.11}$$

In some cases silver is used to coat the center conductor to further reduce the attenuation. Placing appropriate relative values on R_1 and R_2 can give other optima.

A different optimization condition concerns electrical breakdown. For a given Ψ_2 and breakdown electric field E_0 (on the center conductor) one would like to have the maximum power from the oscillator. For the coaxial geometry we have

$$E = E_0 \frac{\Psi_1}{\Psi} \qquad \text{(electric field)}$$

$$\Psi_1 \le \Psi \le \Psi_2 \qquad \text{(cylindrical radius)}$$

$$V = -\int_{\Psi_1}^{\Psi_2} \mathrm{E d}\Psi = -E_0\,\Psi_1 \ell n\left(\frac{\Psi_2}{\Psi_1}\right) \qquad \text{(maximum voltage)}$$

$$P = \frac{V^2}{Z_c} \qquad \text{(power)} \tag{2.12}$$

$$= \frac{2\pi}{Z_w}\left[E_0\Psi_2\right]^2 \xi^{-2} \ell n(\xi)$$

Maximizing $\xi^{-2}\ell n(\xi)$ we have

$$\xi = \frac{\Psi_2}{\Psi_1} = e^{1/2} \simeq 1.65 \tag{2.13}$$

for which

$$\begin{aligned} Z_c &\simeq \frac{120}{\varepsilon_r^{1/2}} \ell n\left(e^{1/2}\right) = 60\varepsilon_r^{-1/2} \\ &= \begin{cases} 60\,\Omega \text{ for air} \\ 40\,\Omega \text{ for oil or polyethylene} \end{cases} \end{aligned} \tag{2.14}$$

We can observe from (2.10), (2.11), and (2.14) that these impedances are in the range that we would like for feeding antennas. For the switched oscillator one would like a few ohms for Z_c. Here we have Z_c of the order of 50 to 100 ohms. This has a significant impact on the switch. For a given switch inductance L the time constant L/Z_c for the switched oscillator versus $L/[2Z_c]$ for the present concept is a reduction of over an order of magnitude. This may allow us to go to higher voltages (implying larger L) or to higher frequencies. Remember now that pulse width is gained by increased oscillator length, ℓ.

3. Use of Symmetry Plane

Symmetry plays an important role in the design of electromagnetic devices [18]. The point symmetry groups (rotation and reflection) are commonly encountered. Symmetry planes (reflection) are often used for microwave devices, for example, in magic tees.

For present purposes consider a symmetry plane, say $z = 0$ in Cartesian (x, y, z) coordinates. This gives a reflection dyadic

$$\overleftrightarrow{R}_z = \overleftrightarrow{1} - 2\,\vec{1}_z\,\vec{1}_z = \vec{1}_x\,\vec{1}_x + \vec{1}_y\,\vec{1}_y - \vec{1}_z\,\vec{1}_z \tag{3.1}$$

to characterize the symmetry. In this case the electromagnetic fields can be divided into two parts, symmetric (sy) and antisymmetric (as), which are characterized by

$$\begin{aligned}
\vec{E}_{\substack{\mathrm{sy}\\ \mathrm{as}}}(\vec{r}_m,t) &= \pm\overleftrightarrow{R}_z \cdot \vec{E}_{\substack{\mathrm{sy}\\ \mathrm{as}}}(\vec{r},t)\\
\vec{H}_{\substack{\mathrm{sy}\\ \mathrm{as}}}(\vec{r}_m,t) &= \mp\overleftrightarrow{R}_z \cdot \vec{H}_{\substack{\mathrm{sy}\\ \mathrm{as}}}(\vec{r},t)\\
\vec{r}_m &= \overleftrightarrow{R} \cdot \vec{r} = \text{mirror coordinate}
\end{aligned} \tag{3.2}$$

These two parts do not couple to each other and it is quite possible to excite either one of these alone.

In the more general microwave context one can have a resonant cavity with a symmetry plane as in Fig. 3.1. One can also have waveguide ports for coupling to the cavity. For standard rectangular waveguides with only one propagating mode (the lowest mode), these can also be coupled to the cavity at locations and orientations which possess the same symmetry plane. In the example one waveguide couples to only symmetric modes while the other couples only to antisymmetric modes. There are higher order evanescent modes in the waveguides, but these transport no real power. If one sends power in through one guide to build up a cavity resonance (with the same symmetry as the waveguide mode), no power exists through the second waveguide. If now one destroys the cavity symmetry (say by switching some nonlinear element appropriately located in the cavity) then the modal structure is changed and power exits via the second guide to a load. Note that the exciting port need not have this symmetry since the resonant mode that one is exciting (by tuning to the resonant frequency) need only have this symmetry.

Applying this concept to the transmission-line oscillator, consider the configurations in Fig. 3.2. In this case the symmetry plane dividing the transmission line in its center is also a symmetry plane for the transmission line (such as a coax) to the load. As in Section 2 the power is fed in at one or more positions along the transmission

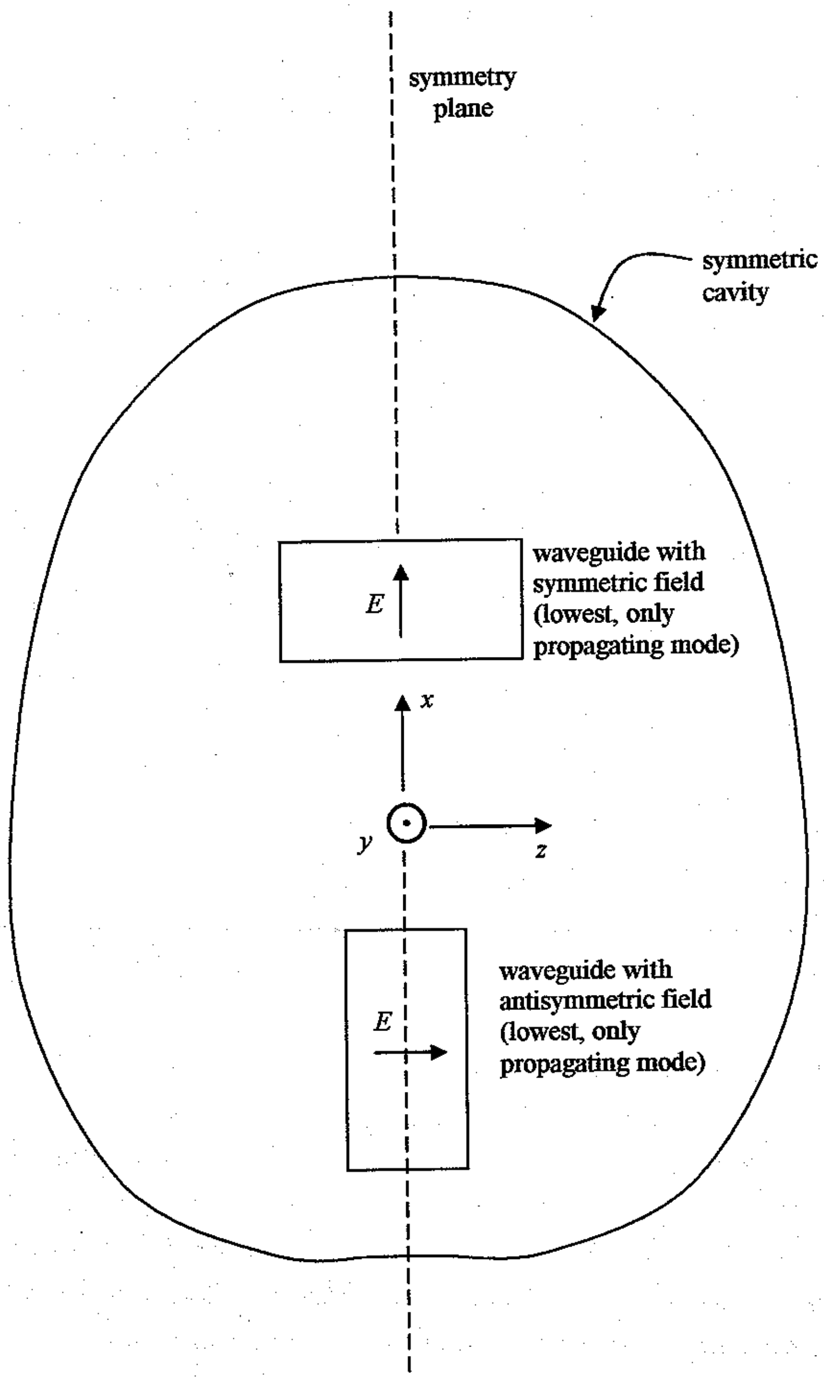
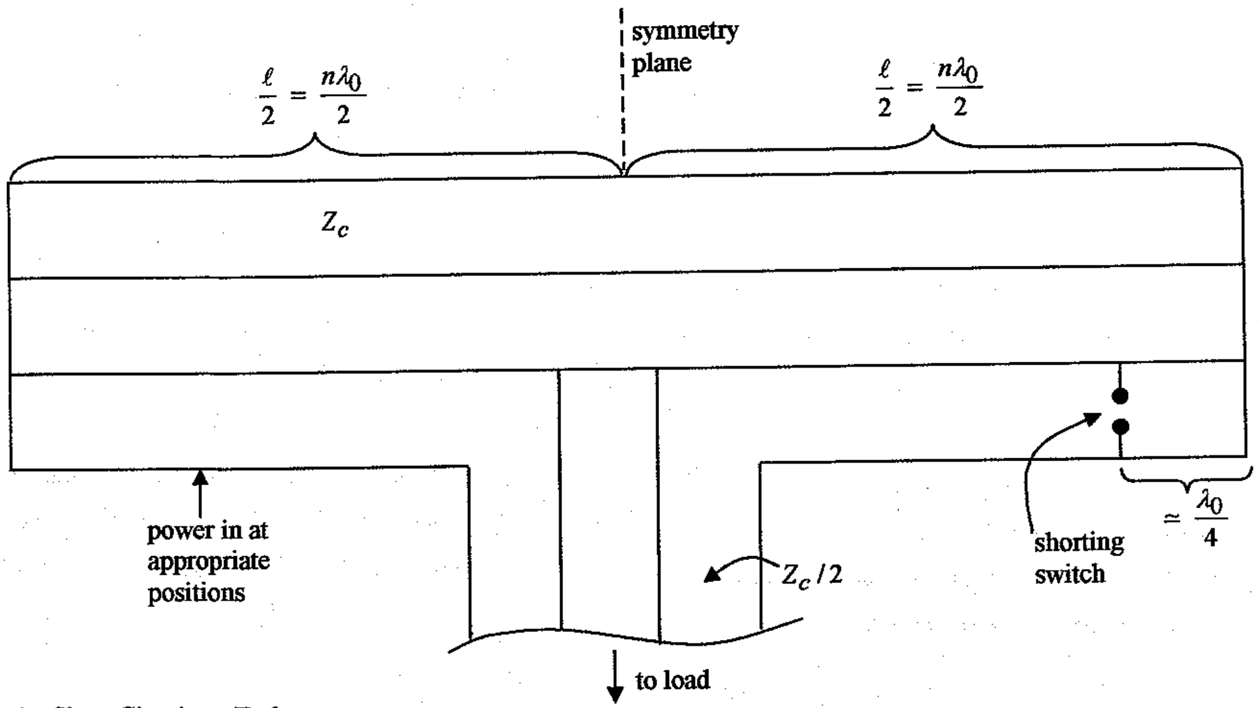
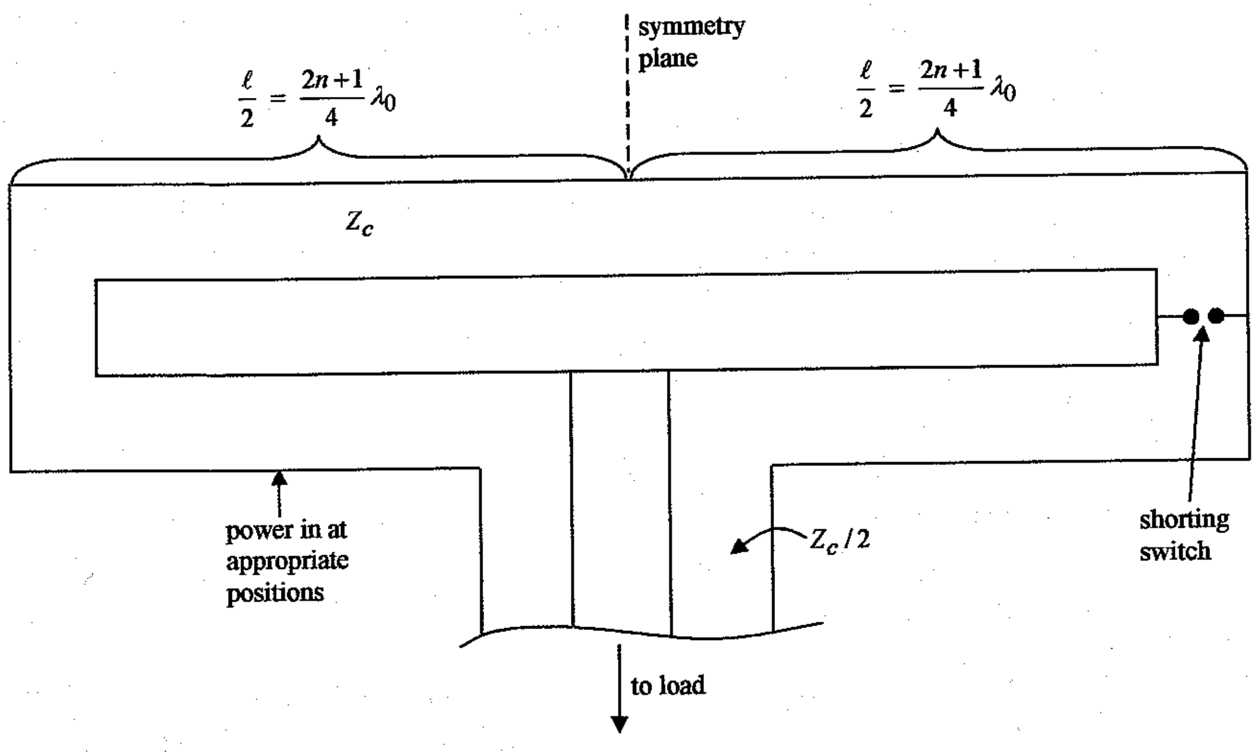


Fig. 3.1. Cavity With Waveguide Ports With Common Symmetry Plane



A. Short Circuits at Ends.



B. Open Circuits at Ends.

Fig. 3.1. Transmission-Line Oscillator Switched Near One End With Output in Center

line. A special case would have a single position for a coupling loop on the symmetry plane (a voltage null during feed in of power) to give the required antisymmetric resonant mode. With the output transmission line designed to propagate only the TEM mode, no power is fed to the load during resonant charging.

In one configuration (A, shorts at both ends) one can place a shorting switch (center to outer conductor) at about $\lambda_0/4$ from one end. On closing, a wave with reversed phase is sent toward the center. At the center we no longer have a voltage null, but a voltage maximum. To quickly feed the power from the oscillator one might choose the characteristic impedance as about $Z_c/2$ to account for the two halves of the oscillator feeding out in parallel.

In the other configuration (B, open at both ends) the situation is similar, except that now each half of the oscillator needs to be an odd number of quarter wavelengths. In this case the shorting switch connects at one end to achieve the phase reversal. In both of these configurations the power can be switched to the load in one transit time $t_r = \ell/\mathrm{v}$ of the full oscillator length.

One potential advantage of these configurations concerns the location of the switch. In Section 2 the output wave passes through the switch and it does not have zero impedance after closure. In the present configurations there is no switch in series with the output. While a practical switch has inductance after closure, this can be compensated by adjustment of its position. Note that it also has capacitance before closure which needs to be tuned out by adjustment of the transmission-line parameters.

In order not to introduce higher-order (non TEM) propagating modes in the oscillator or output transmission line, the outer coaxial radius Ψ_2 should not be too large. While one can compute the cutoff wavelengths for such modes from Bessel-function formulae, the lowest order of these has a cutoff wavelength

$$\lambda_c \simeq \pi(\Psi_1 + \Psi_2) \tag{3.3}$$

i.e., roughly the average circumference [15]. Thus we require

$$\lambda_c < \lambda_0 \tag{3.4}$$

by some factor greater than one. This limits the outer radius and thereby the energy per unit length that can be stored (given some maximum field E_0 in (2.11)). Lower frequencies are then favored by this kind of oscillator.

4. Binary Pulse Compression

Now consider a very different kind of pulse compression. The reader can consult [9] for a more complete discussion.

In simplified terms what one has is two phase-locked equal-power signals, each of length T. One of these has its phase reversed half way through the pulse (time $T/2$). These are then fed into two ports of a 4-port 3db directional coupler. Then the signals are combined in phase for $0 < t$, $T/2$ out of one port and out of phase for $T/2 < t < T$ out of the other output port. Delaying the first signal by a delay line of length T/2 the two waves are combined in phase to give twice the power in half the time.

Such a single-stage device can be extended to M stages giving 2^{M} times the power in a pulse of width 2^{-MT}. This is done by a "coding" of the phase reversals to achieve the power doubling at successive stages.

To achieve the fast phase reversals Klystron amplifiers have been used to take their inputs from low-power waveform synthesizers. For our applications we may need to revisit this question so as to design appropriate sources.

5. Concluding Remarks

Here we have begun the exproration of a new kind of source for our high-power-microwave applications. Potentially it can complement or even extend the capability represented by the switched oscillator.

I would like the thank J. Benford and J. Swegle for discussions and assistance in finding some of the references.

References

1. C. E. Baum, "Differential Switched Oscillators and Associated Antennas", Sensor and Simulation Note 457, June 2001.

2. C. E. Baum, "Modification of TEM-Fed Reflector for Increased Efficiency", Sensor and Simulation Note 458, July 2001.

3. C. E. Baum, "Differential Switched Oscillators and Associated Antennas, Part 2", Sensor and Simulation Note 484, November 2003.

4. C. E. Baum, "Switched Oscillators", Circuit and Electromagnetic System Design Note 45, September 2001.

5. C. E. Baum, "A Transmission-Line Transformer for Matching the Switched Oscillator to a Higher Impedance Load", Circuit and Electromagnetic System Design Note 46, August 2001; Proce. 15th. Zurich EMC Symposium, February 2003, pp. 133-136.

6. C. E. Baum, "Sequential Quarter-Wave Transmission-Line Transformers", Circuit and Electromagnetic System Design Note 47, June 2002.

7. J. M. Lehr, C. E. Baum, and W. D. Prather, "Fundamental Physical Considerations for Ultrafast Spark Gap Switching", Switching Note 28, June 1997; pp. 11-20, in E. Heyman et al (eds.), *Ultra-Wideband, Short-Pulse Electromagnetics 4*, Kluwer Academic/Plenum Publishers, 1999.

8. N. D. Deuyatkov, A. N. Didenko, L. Ya. Zamyatina, S. V. Razin, and Yu. G. Yushkov, "Formation of Powerful Pulses with Accumulation of UHF Energy in a Resonator", Radiotekh. Electron., Vol. 25, 1980, p. 1227ff; Radio Engineering and Electronic Physics, Vo. 25, No. 6, June 1980, pp. 87-89.

9. Z. D. Farkas, "Binary Peak Power Multiplier and its Application to Linear Accelerator Design", IEEE Trans. Microwave Theory and Techniques, 1986, pp. 1036-1043.

10. R. A. Alvarez, D. P. Byrne, and R. M. Johnson, "Prepulse Suppression in Micrwave Pulse-Compression Cavities", Rev. Sci. Instrum., 1986, pp. 2475-2480.

11. R. A. Alvarez, "Some Properties of Microwave Resonant Cavities Relevant to Pulse-Compression Power Amplification", Rev. Sci. Instrum., 1986, pp. 2481-2488.

12. Z. D. Farkas, T. L. Lavine, A. Menegat, R. H. Miller, C. Nantista, G. Spalek, and P. B. Wilson, "Radio Frequency Pulse Compression Experiments at SLAC", SPIE Vol. 1407, Intense Microwave and Particle Beams II, 1991, pp. 502-511.

13. D. V. Giri and F. M. Tesche, "Classification of Intentional Electromagnetic Interference (IEMI)", IEEE Trans. EMC, Special Issue on High-Power Electromagnetics and Intentional Electromagnetic Interference, 2004, in publication.

14. T. Moreno, *Microwave Transmission Design Data*, Dover, 1948.

15. S. Ramo, J. R. Whinnery, and T. Van Duzer, "*Fields and Waves in Communication Electronics*", Wiley, 1965.

16. J. Benford and J. Swegle, "*High Power Microwaves'*, Artech House, 1992.

17. A. V. Gaponov-Grekhov and V. L. Granatstein, *Applications of High-Power Microwaves*, Artech House, 1994, pp. 290-297.

18. C. E. Baum and H. N. Kritikos (eds.) *Electromagnetic Symmetry*, Taylor & Francis, 1995.

19. R. Yu. Chumerin, A. N. Didenko, S. A. Novikov, and Yu. G. Yushkov, "Resonance Pulse Compression as an Effective Method for Generation of UWB High Repetitive Microwave Pulses", pp. 427-434, in E. Mokole et al, *Ultra-Wideband, Short-Pulse Electromagnetics 6*, Kluwer Academic/Plenum Publishers, 2003.